PTO/SB/14 (07-07)
Approved for use through 06/30/2010. OMB 0651-0032
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE

Application Data Sheet 37 CFR 1.76	Attorney Docket Number	WTCY-0026-P07
	Application Number	
Title of Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS	

The application data sheet is part of the provisional or nonprovisional application for which it is being submitted. The following form contains the bibliographic data arranged in a format specified by the United States Patent and Trademark Office as outlined in 37 CFR 1.76.
This document may be completed electronically and submitted to the Office in electronic format using the Electronic Filing System (EFS) or the document may be printed and included in a paper filed application.

Secrecy Order 37 CFR 5.2

☐ Portions or all of the application associated with this Application Data Sheet may fall under a Secrecy Order pursuant to 37 CFR 5.2 (Paper filers only. Applications that fall under Secrecy Order may not be filed electronically.)

Applicant Information:

Applicant 1 Remove

Applicant Authority	◉ Inventor	○ Legal Representative under 35 U.S.C. 117	○ Party of Interest under 35 U.S.C. 118	
Prefix	**Given Name**	**Middle Name**	**Family Name**	**Suffix**
	Aristeidis		Karalis	

Residence Information (Select One)	◉ US Residency	○ Non US Residency	○ Active US Military Service		
City	Boston	**State/Province**	MA	**Country of Residence** i	US
Citizenship under 37 CFR 1.41(b) i	GR				
Mailing Address of Applicant:					
Address 1	151 Tremont Street				
Address 2	Apartment 21F				
City	Boston	**State/Province**	MA		
Postal Code	02111	**Country** i	US		

Applicant 2 Remove

Applicant Authority	◉ Inventor	○ Legal Representative under 35 U.S.C. 117	○ Party of Interest under 35 U.S.C. 118	
Prefix	**Given Name**	**Middle Name**	**Family Name**	**Suffix**
	Andre	B.	Kurs	

Residence Information (Select One)	◉ US Residency	○ Non US Residency	○ Active US Military Service		
City	Chestnut Hill	**State/Province**	MA	**Country of Residence** i	US
Citizenship under 37 CFR 1.41(b) i	BR				
Mailing Address of Applicant:					
Address 1	250 Hammond Pond Parkway				
Address 2	Apartment 1203S				
City	Chestnut Hill	**State/Province**	MA		
Postal Code	02467	**Country** i	US		

Applicant 3 Remove

Applicant Authority	◉ Inventor	○ Legal Representative under 35 U.S.C. 117	○ Party of Interest under 35 U.S.C. 118	
Prefix	**Given Name**	**Middle Name**	**Family Name**	**Suffix**
	Andrew	J.	Campanella	

Residence Information (Select One)	◉ US Residency	○ Non US Residency	○ Active US Military Service		
City	Waltham	**State/Province**	MA	**Country of Residence** i	US

Application Data Sheet 37 CFR 1.76	Attorney Docket Number	WTCY-0026-P07
	Application Number	

Title of Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS

Citizenship under 37 CFR 1.41(b) i	US		
Mailing Address of Applicant:			
Address 1	30 Orange Street		
Address 2			
City	Waltham	**State/Province**	MA
Postal Code	02453	**Country**i	US

Applicant 4 Remove

Applicant Authority ◉Inventor ○Legal Representative under 35 U.S.C. 117 ○Party of Interest under 35 U.S.C. 118

Prefix	**Given Name**	**Middle Name**	**Family Name**	**Suffix**
	Konrad		Kulikowski	

Residence Information (Select One) ◉ US Residency ○ Non US Residency ○ Active US Military Service

City	Somerville	**State/Province**	MA	**Country of Residence** i	US
Citizenship under 37 CFR 1.41(b) i	US				

Mailing Address of Applicant:			
Address 1	91 Pearson Avenue		
Address 2	Apartment 1		
City	Somerville	**State/Province**	MA
Postal Code	02144	**Country**i	US

Applicant 5 Remove

Applicant Authority ◉Inventor ○Legal Representative under 35 U.S.C. 117 ○Party of Interest under 35 U.S.C. 118

Prefix	**Given Name**	**Middle Name**	**Family Name**	**Suffix**
	Katherine	L.	Hall	

Residence Information (Select One) ◉ US Residency ○ Non US Residency ○ Active US Military Service

City	Westford	**State/Province**	MA	**Country of Residence** i	US
Citizenship under 37 CFR 1.41(b) i	US				

Mailing Address of Applicant:			
Address 1	21 Griffin Road		
Address 2			
City	Westford	**State/Province**	MA
Postal Code	01886	**Country**i	US

Applicant 6 Remove

Applicant Authority ◉Inventor ○Legal Representative under 35 U.S.C. 117 ○Party of Interest under 35 U.S.C. 118

Prefix	**Given Name**	**Middle Name**	**Family Name**	**Suffix**
	Marin		Soljacic	

Residence Information (Select One) ◉ US Residency ○ Non US Residency ○ Active US Military Service

City	Belmont	**State/Province**	MA	**Country of Residence** i	US
Citizenship under 37 CFR 1.41(b) i	HR				

Application Data Sheet 37 CFR 1.76	Attorney Docket Number	WTCY-0026-P07
	Application Number	
Title of Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS	

Mailing Address of Applicant:			
Address 1	44 Westlund Road		
Address 2			
City	Belmont	State/Province	MA
Postal Code	02478	Country i	US

Applicant 7 — Remove

Applicant Authority	◉ Inventor	◯ Legal Representative under 35 U.S.C. 117	◯ Party of Interest under 35 U.S.C. 118	
Prefix	Given Name	Middle Name	Family Name	Suffix
	Morris	P.	Kesler	

Residence Information (Select One) ◉ US Residency ◯ Non US Residency ◯ Active US Military Service

City	Bedford	State/Province	MA	Country of Residence i	US
Citizenship under 37 CFR 1.41(b) i	US				

Mailing Address of Applicant:			
Address 1	95 Hancock Street		
Address 2			
City	Bedford	State/Province	MA
Postal Code	01730	Country i	US

All Inventors Must Be Listed - Additional Inventor Information blocks may be generated within this form by selecting the **Add** button. — Add

Correspondence Information:

Enter either Customer Number or complete the Correspondence Information section below.
For further information see 37 CFR 1.33(a).

☐ **An Address is being provided for the correspondence Information of this application.**

Customer Number	43520
Email Address	patents@strategic-patents.com

Add Email — Remove Email

Application Information:

Title of the Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS		
Attorney Docket Number	WTCY-0026-P07	Small Entity Status Claimed	☐
Application Type	Nonprovisional		
Subject Matter	Utility		
Suggested Class (if any)		Sub Class (if any)	
Suggested Technology Center (if any)			
Total Number of Drawing Sheets (if any)	53	Suggested Figure for Publication (if any)	

Application Data Sheet 37 CFR 1.76	Attorney Docket Number	WTCY-0026-P07
	Application Number	
Title of Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS	

Publication Information:

☐ Request Early Publication (Fee required at time of Request 37 CFR 1.219)

☐ **Request Not to Publish.** I hereby request that the attached application not be published under 35 U.S.C. 122(b) and certify that the invention disclosed in the attached application **has not and will not** be the subject of an application filed in another country, or under a multilateral international agreement, that requires publication at eighteen months after filing.

Representative Information:

Representative information should be provided for all practitioners having a power of attorney in the application. Providing this information in the Application Data Sheet does not constitute a power of attorney in the application (see 37 CFR 1.32). Enter either Customer Number or complete the Representative Name section below. If both sections are completed the Customer Number will be used for the Representative Information during processing.

Please Select One:	◉ Customer Number	○ US Patent Practitioner	○ Limited Recognition (37 CFR 11.9)
Customer Number	43520		

Domestic Benefit/National Stage Information:

This section allows for the applicant to either claim benefit under 35 U.S.C. 119(e), 120, 121, or 365(c) or indicate National Stage entry from a PCT application. Providing this information in the application data sheet constitutes the specific reference required by 35 U.S.C. 119(e) or 120, and 37 CFR 1.78(a)(2) or CFR 1.78(a)(4), and need not otherwise be made part of the specification.

Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/100721	2008-09-27
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/108743	2008-10-27
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/147386	2009-01-26
Prior Application Status	Pending		Remove
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12/567716	non provisional of	61/152086	2009-02-12
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/178508	2009-05-15
Prior Application Status	Pending		Remove

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Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/182768	2009-06-01
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/121159	2008-12-09
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/142977	2009-01-07
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/142885	2009-01-06
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/142796	2009-01-06
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Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/142880	2009-01-06
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/142818	2009-01-06
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Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/142887	2009-01-06
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/156764	2009-03-02
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Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/143058	2009-01-07
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/152390	2009-02-13
Prior Application Status	Pending		Remove

Application Data Sheet 37 CFR 1.76	Attorney Docket Number	WTCY-0026-P07
	Application Number	
Title of Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS	

Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/163695	2009-03-26
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/172633	2009-04-24
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/169240	2009-04-14
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/173747	2009-04-29
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
	Continuation in part of	12/567716	2009-09-25

Additional Domestic Benefit/National Stage Data may be generated within this form by selecting the **Add** button. Add

Foreign Priority Information:

This section allows for the applicant to claim benefit of foreign priority and to identify any prior foreign application for which priority is not claimed. Providing this information in the application data sheet constitutes the claim for priority as required by 35 U.S.C. 119(b) and 37 CFR 1.55(a).

Remove

Application Number	Country i	Parent Filing Date (YYYY-MM-DD)	Priority Claimed
			(●) Yes ( ) No

Additional Foreign Priority Data may be generated within this form by selecting the **Add** button. Add

Assignee Information:

Providing this information in the application data sheet does not substitute for compliance with any requirement of part 3 of Title 37 of the CFR to have an assignment recorded in the Office.

Assignee 1 Remove

If the Assignee is an Organization check here. ☐

Prefix	**Given Name**	Middle Name	**Family Name**	Suffix

Application Data Sheet 37 CFR 1.76	Attorney Docket Number	WTCY-0026-P07
	Application Number	
Title of Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS	

Mailing Address Information:			
Address 1			
Address 2			
City		**State/Province**	
Country i		Postal Code	
Phone Number		Fax Number	
Email Address			

Additional Assignee Data may be generated within this form by selecting the **Add** button.

Add

Signature:

A signature of the applicant or representative is required in accordance with 37 CFR 1.33 and 10.18. Please see 37 CFR 1.4(d) for the form of the signature.

Signature	/John Nortrup/			Date (YYYY-MM-DD)	2009-12-28
First Name	John	Last Name	Nortrup	Registration Number	59063

Privacy Act Statement

The Privacy Act of 1974 (P.L. 93-579) requires that you be given certain information in connection with your submission of the attached form related to a patent application or patent. Accordingly, pursuant to the requirements of the Act, please be advised that: (1) the general authority for the collection of this information is 35 U.S.C. 2(b)(2); (2) furnishing of the information solicited is voluntary; and (3) the principal purpose for which the information is used by the U.S. Patent and Trademark Office is to process and/or examine your submission related to a patent application or patent. If you do not furnish the requested information, the U.S. Patent and Trademark Office may not be able to process and/or examine your submission, which may result in termination of proceedings or abandonment of the application or expiration of the patent.

The information provided by you in this form will be subject to the following routine uses:

1. The information on this form will be treated confidentially to the extent allowed under the Freedom of Information Act (5 U.S.C. 552) and the Privacy Act (5 U.S.C. 552a). Records from this system of records may be disclosed to the Department of Justice to determine whether the Freedom of Information Act requires disclosure of these records.

2. A record from this system of records may be disclosed, as a routine use, in the course of presenting evidence to a court, magistrate, or administrative tribunal, including disclosures to opposing counsel in the course of settlement negotiations.

3. A record in this system of records may be disclosed, as a routine use, to a Member of Congress submitting a request involving an individual, to whom the record pertains, when the individual has requested assistance from the Member with respect to the subject matter of the record.

4. A record in this system of records may be disclosed, as a routine use, to a contractor of the Agency having need for the information in order to perform a contract. Recipients of information shall be required to comply with the requirements of the Privacy Act of 1974, as amended, pursuant to 5 U.S.C. 552a(m).

5. A record related to an International Application filed under the Patent Cooperation Treaty in this system of records may be disclosed, as a routine use, to the International Bureau of the World Intellectual Property Organization, pursuant to the Patent Cooperation Treaty.

6. A record in this system of records may be disclosed, as a routine use, to another federal agency for purposes of National Security review (35 U.S.C. 181) and for review pursuant to the Atomic Energy Act (42 U.S.C. 218(c)).

7. A record from this system of records may be disclosed, as a routine use, to the Administrator, General Services, or his/her designee, during an inspection of records conducted by GSA as part of that agency's responsibility to recommend improvements in records management practices and programs, under authority of 44 U.S.C. 2904 and 2906. Such disclosure shall be made in accordance with the GSA regulations governing inspection of records for this purpose, and any other relevant (i.e., GSA or Commerce) directive. Such disclosure shall not be used to make determinations about individuals.

8. A record from this system of records may be disclosed, as a routine use, to the public after either publication of the application pursuant to 35 U.S.C. 122(b) or issuance of a patent pursuant to 35 U.S.C. 151. Further, a record may be disclosed, subject to the limitations of 37 CFR 1.14, as a routine use, to the public if the record was filed in an application which became abandoned or in which the proceedings were terminated and which application is referenced by either a published application, an application open to public inspections or an issued patent.

9. A record from this system of records may be disclosed, as a routine use, to a Federal, State, or local law enforcement agency, if the USPTO becomes aware of a violation or potential violation of law or regulation.

Fig. 1

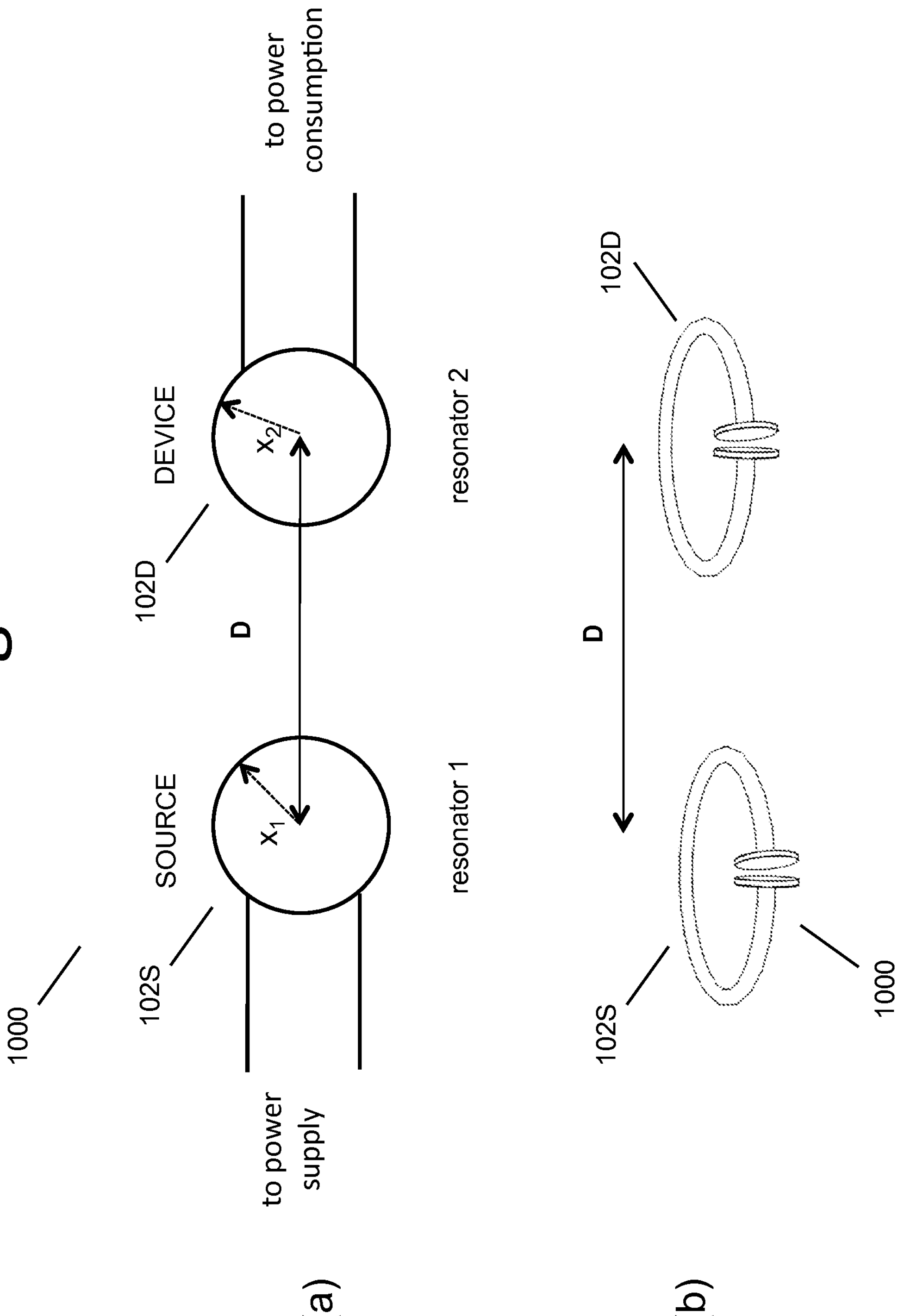

(a)
(b)
1000
102S
102D
SOURCE
DEVICE
x_1
x_2
D
resonator 1
resonator 2
to power supply
to power consumption
102S
102D
D
1000

Fig. 2

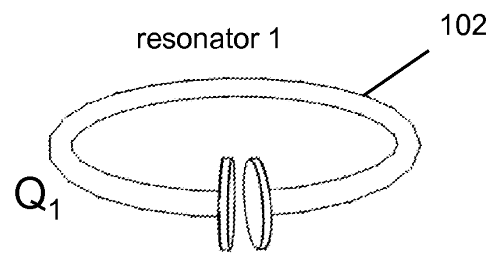


Fig. 3

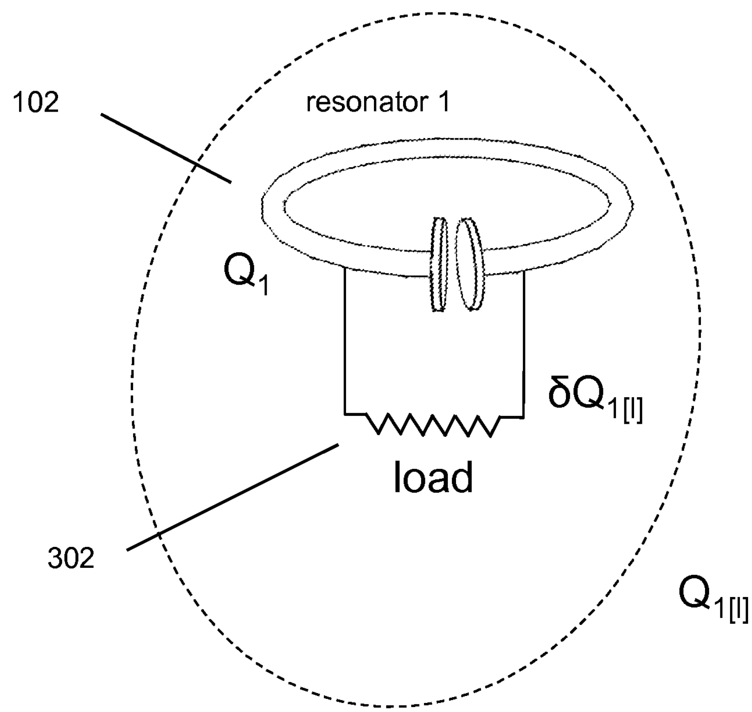


Fig. 4

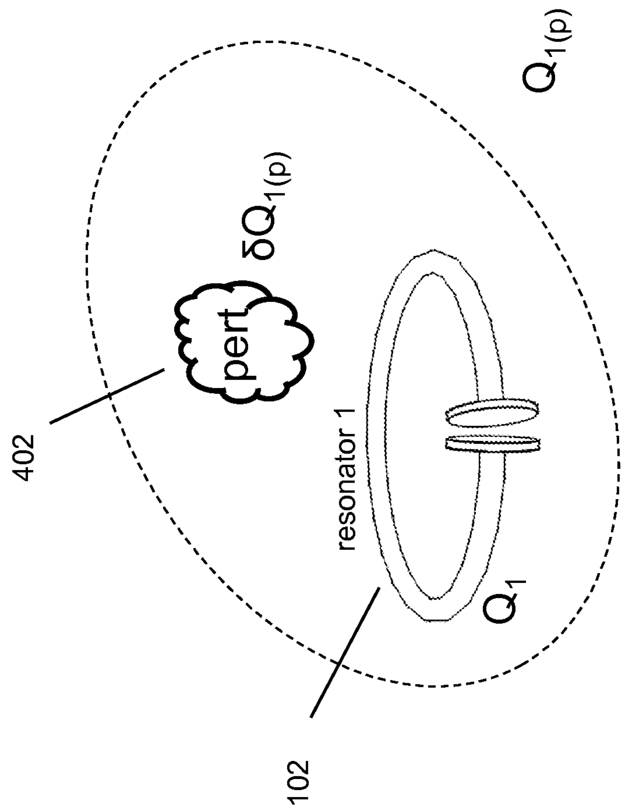
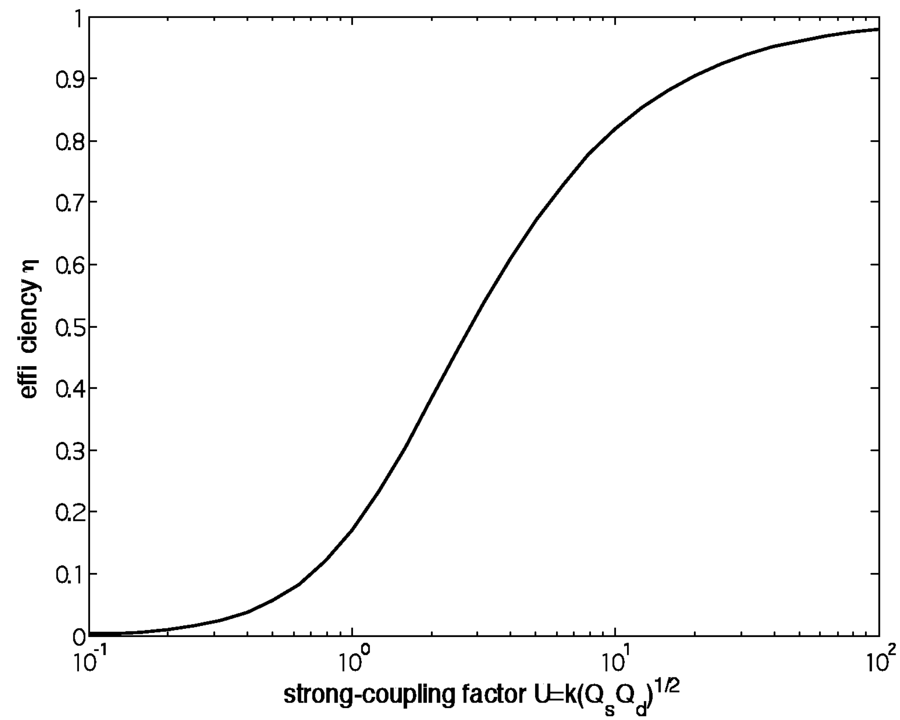


Fig. 5



1
0.9
0.8
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0
effi ciency η
10^-1
10^0
10^1
10^2
strong-coupling factor U=k(Q_s Q_d)^1/2

5

Fig. 6

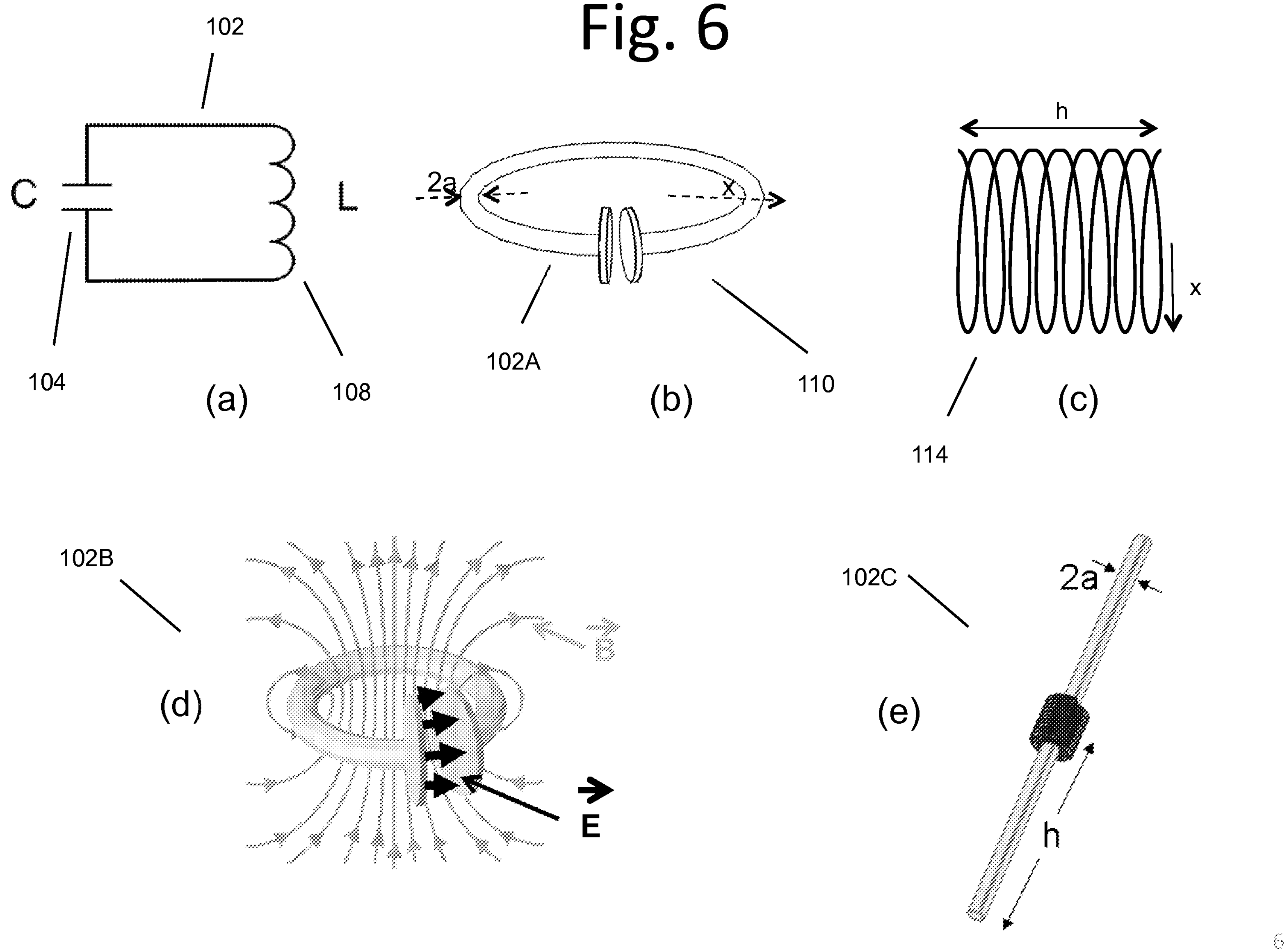
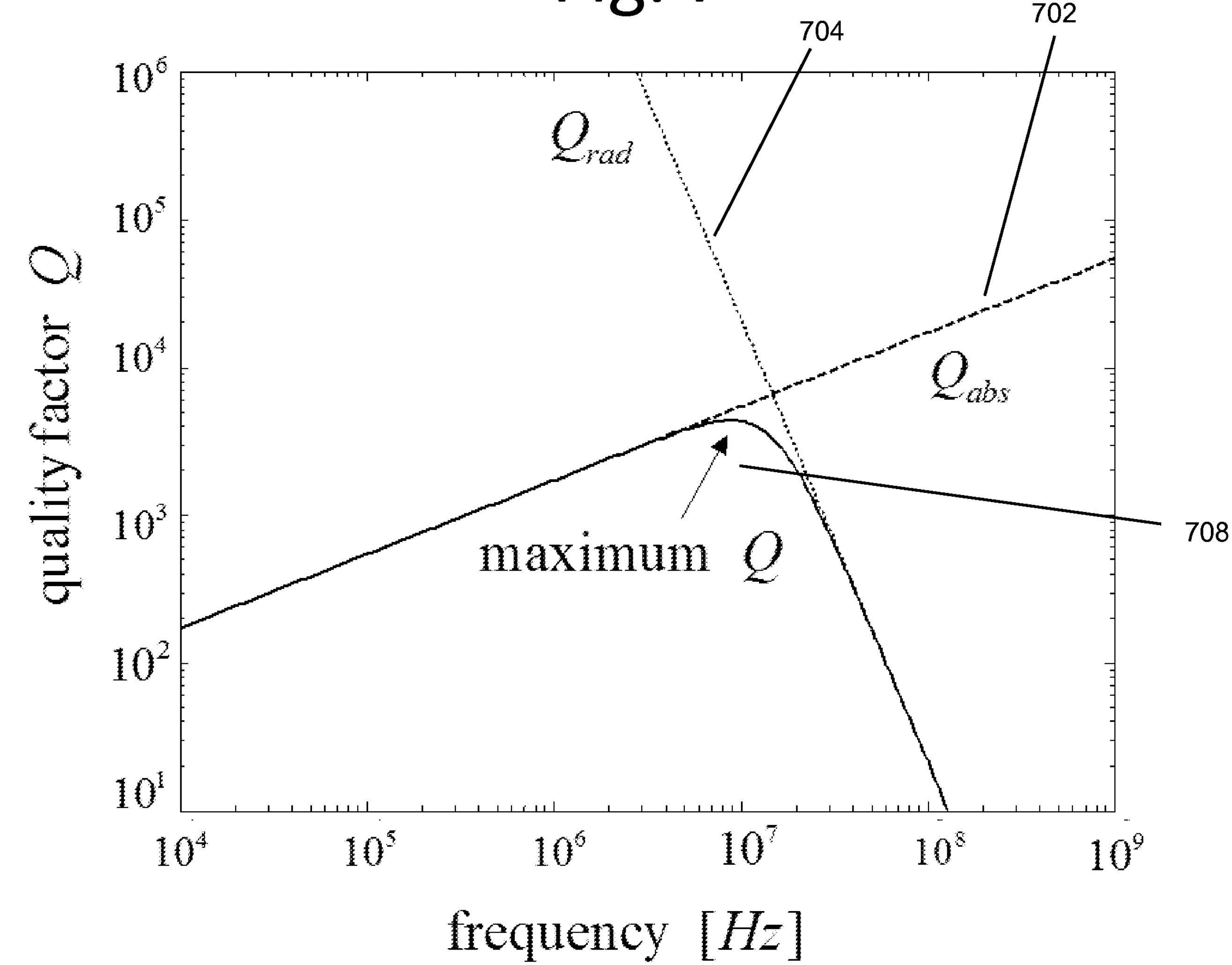


Fig. 7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702
Q_{rad}
Q_{abs}
maximum Q
708
quality factor Q
10^6
10^5
10^4
10^3
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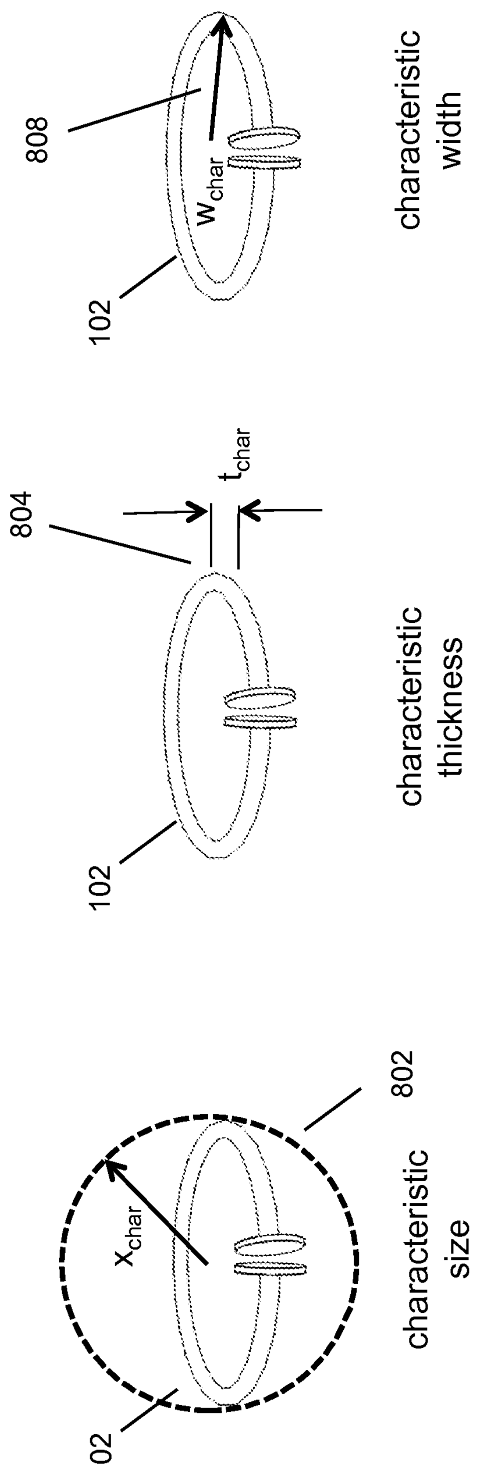
10^9
frequency $[Hz]$

Fig. 8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characteristic size
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804
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characteristic thickness
102
808
w_{char}
characteristic width

Fig. 9

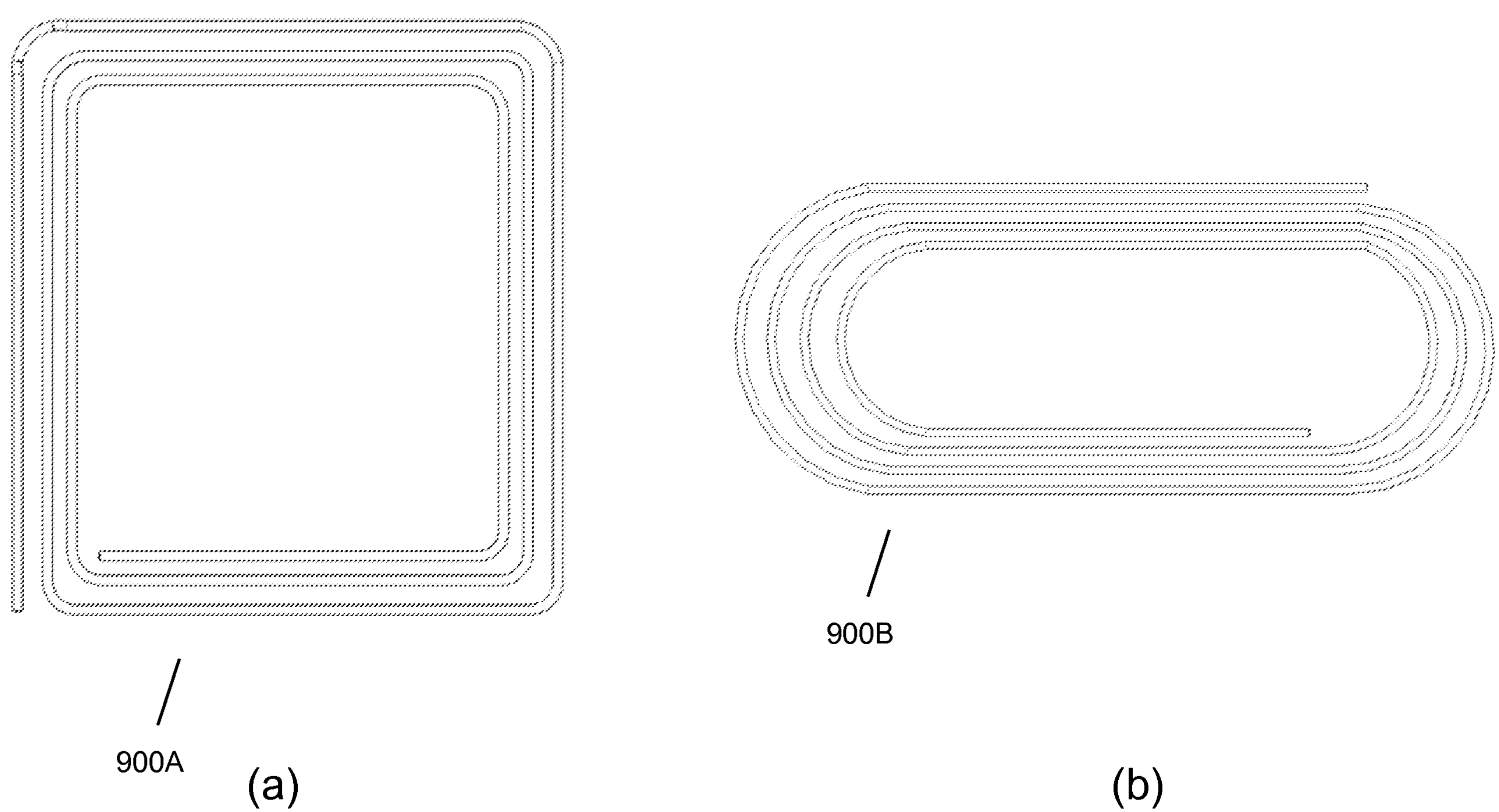


Fig. 10

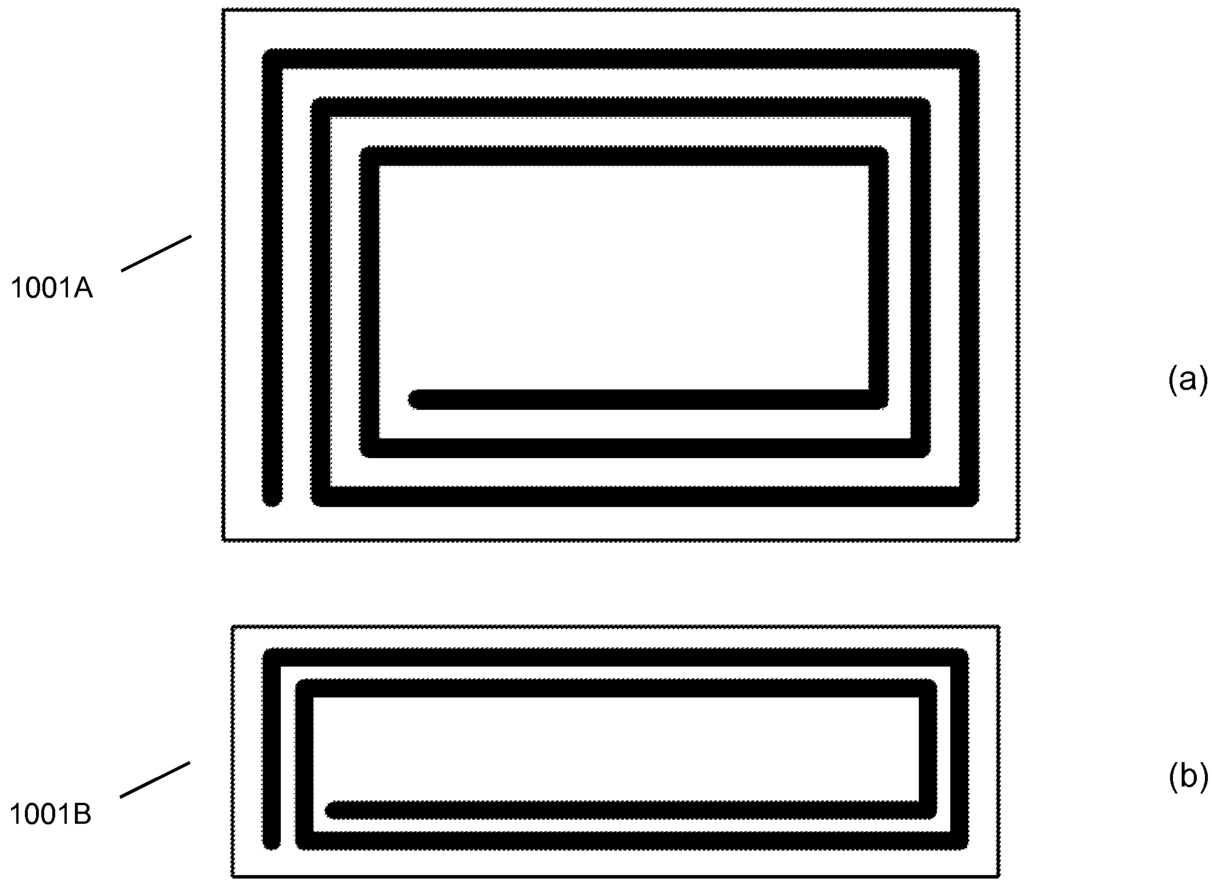
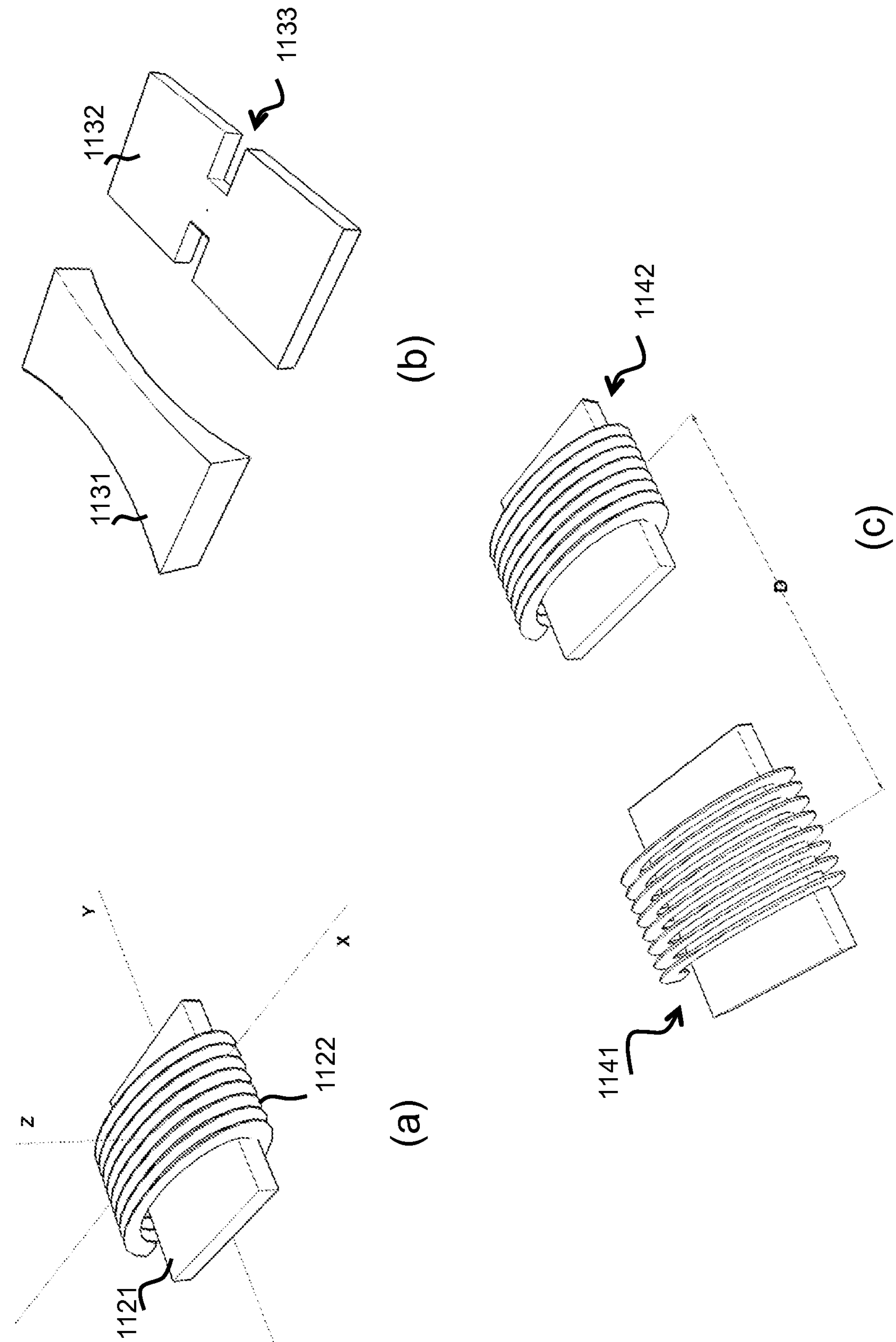


Fig. 11



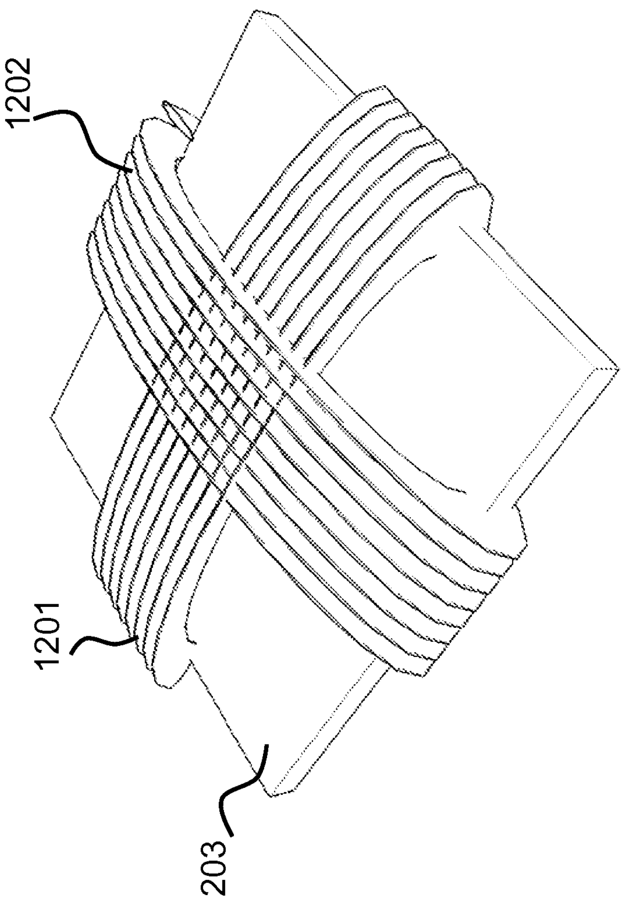
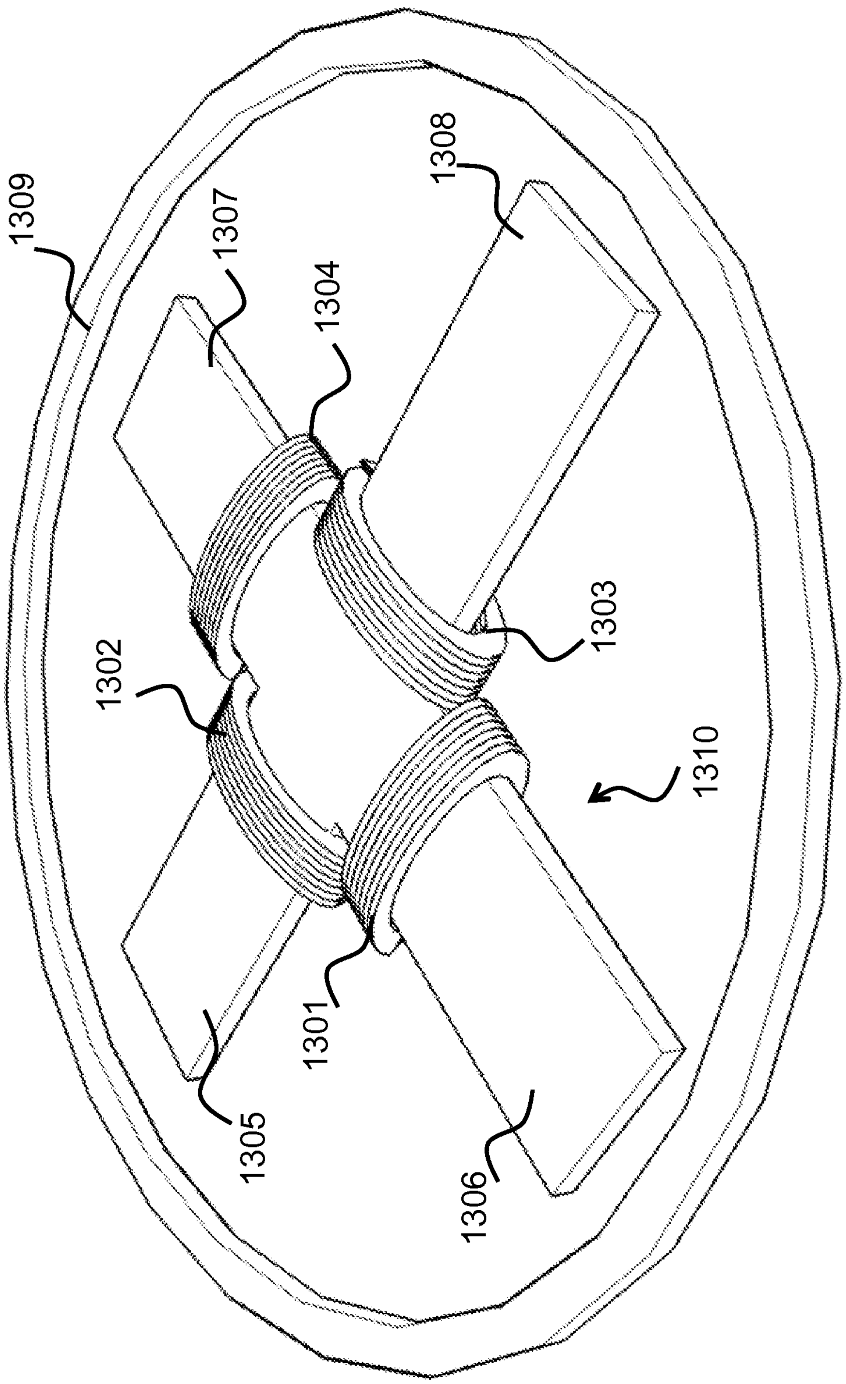


Fig. 12



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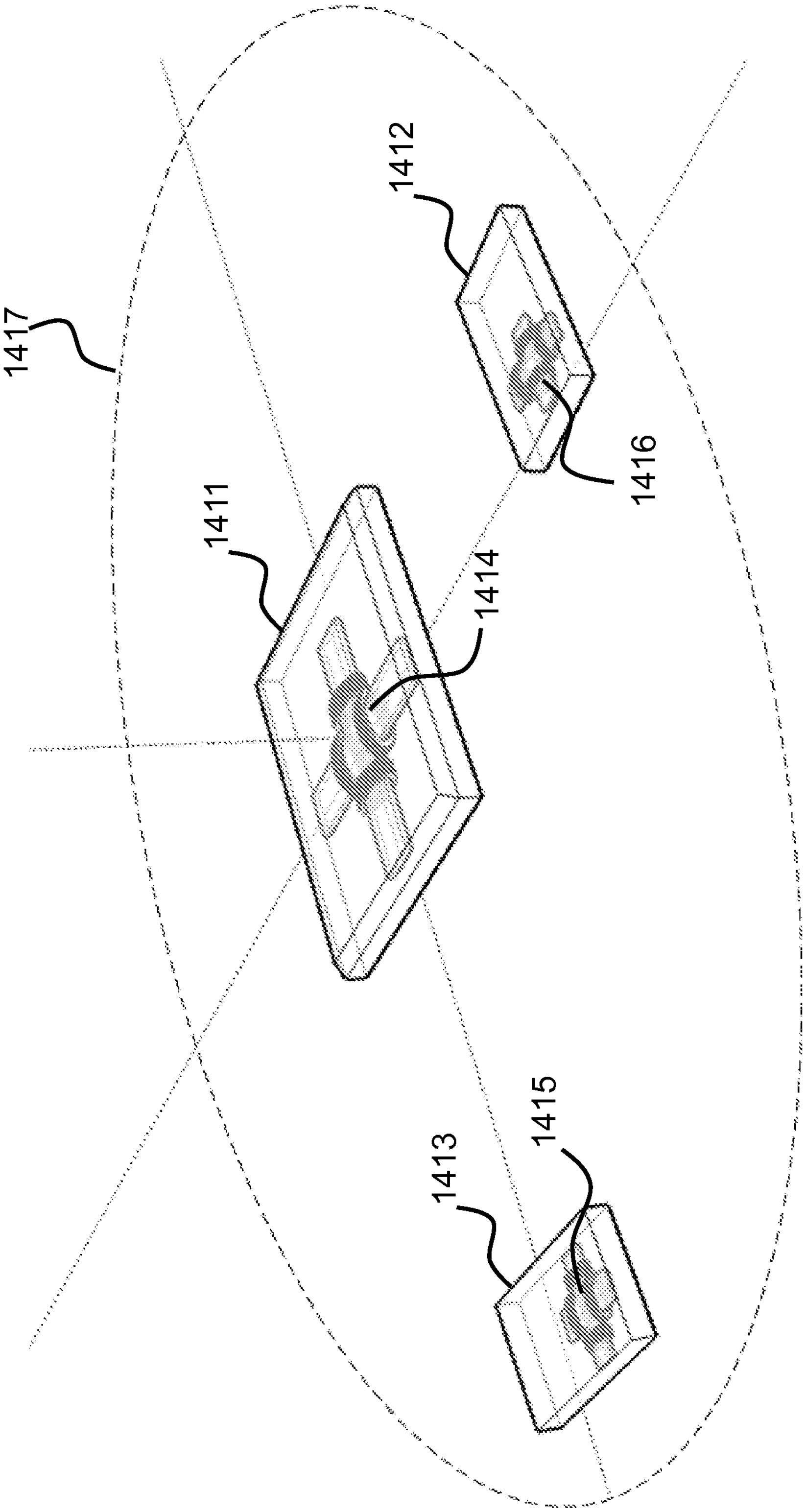


Fig. 14

15

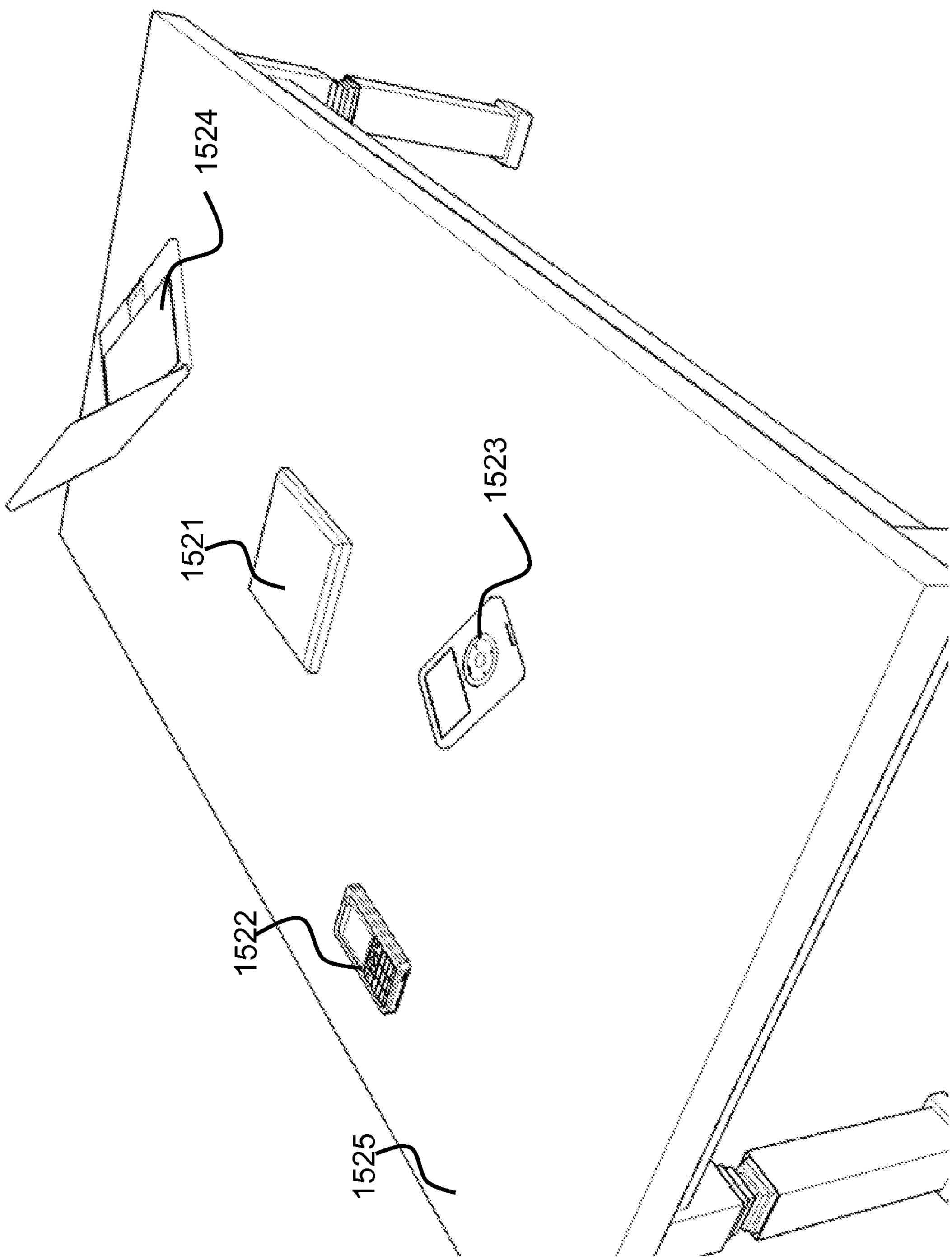
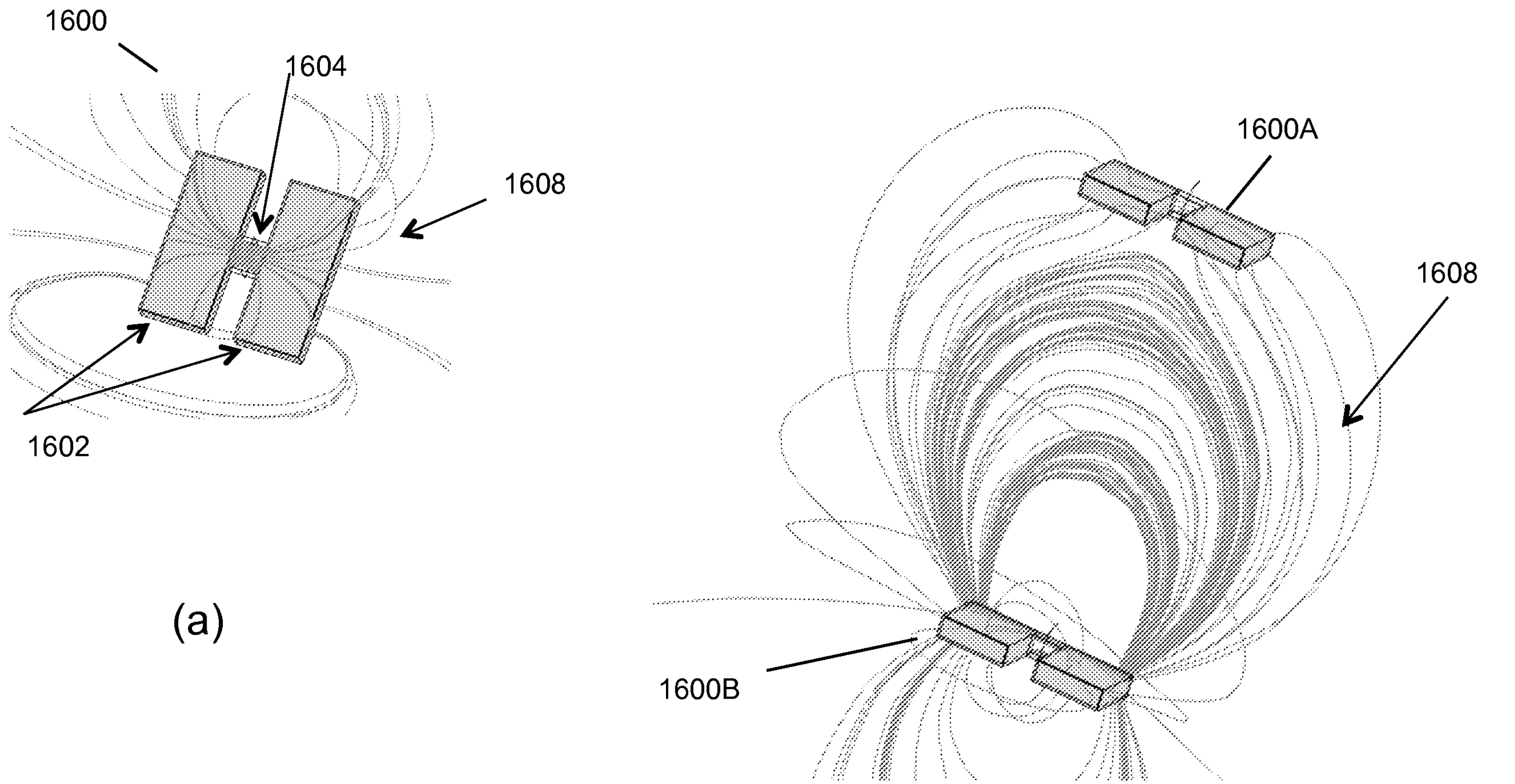


Fig. 15

Fig. 16



(a)

(b)

Fig. 17

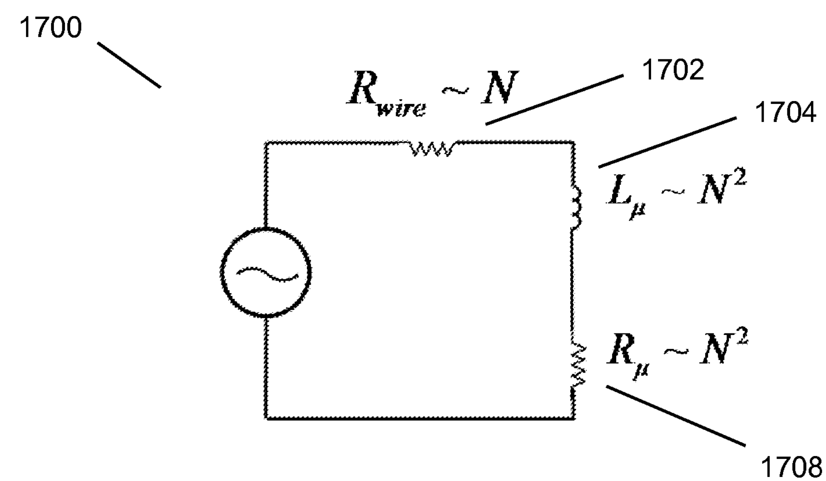
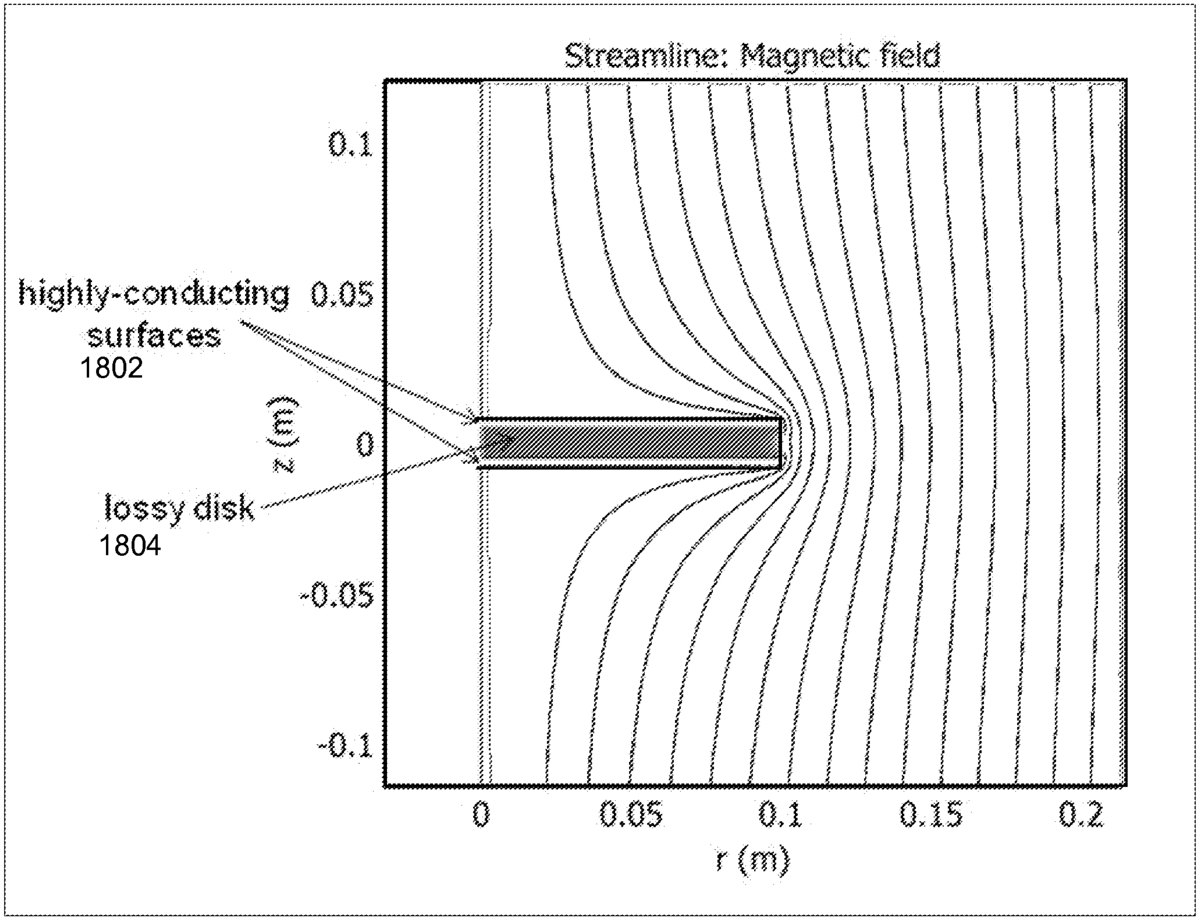


Fig. 18



Streamline: Magnetic field
highly-conducting surfaces
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1804
z (m)
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0
-0.05
-0.1
r (m)
0
0.05
0.1
0.15
0.2

Fig. 19

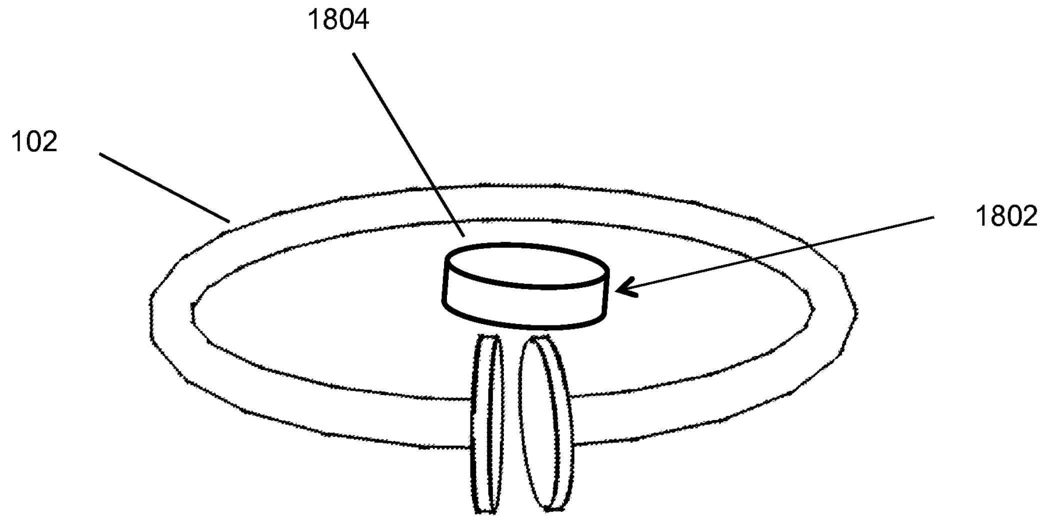


Fig. 20

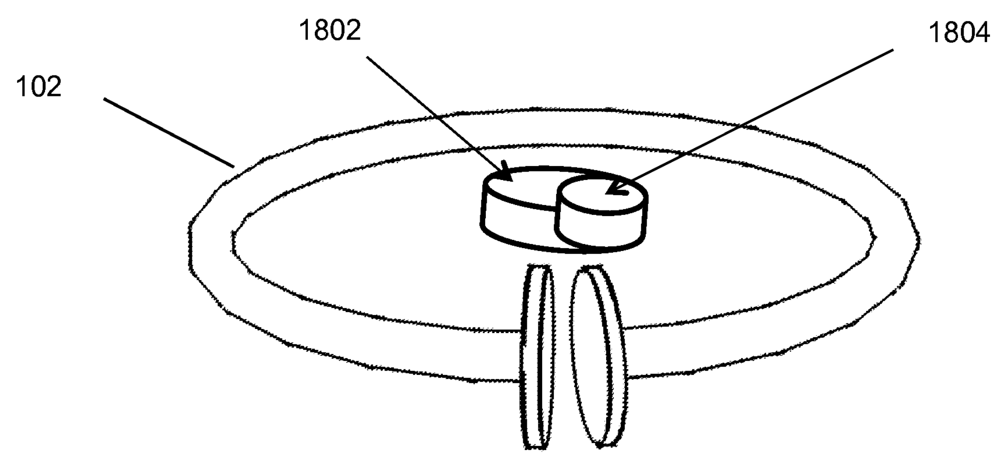


Fig. 21

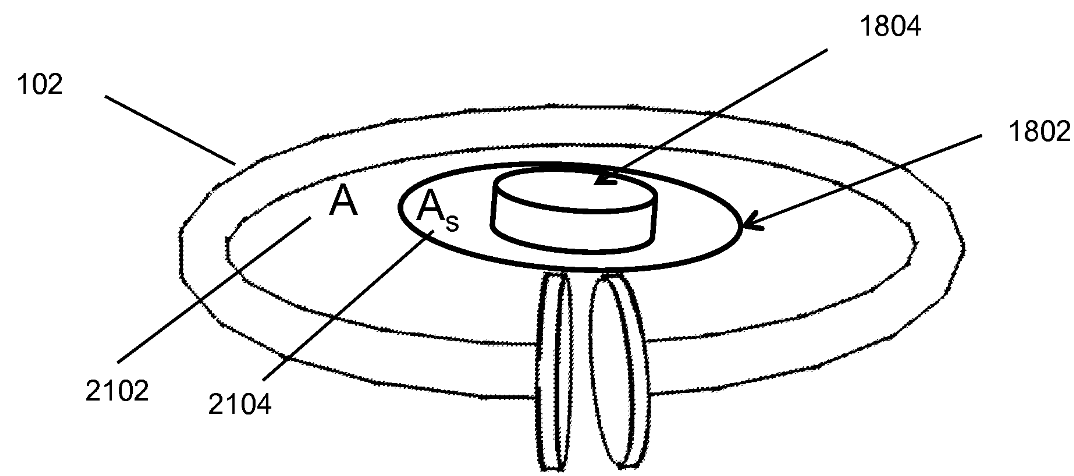


Fig. 22

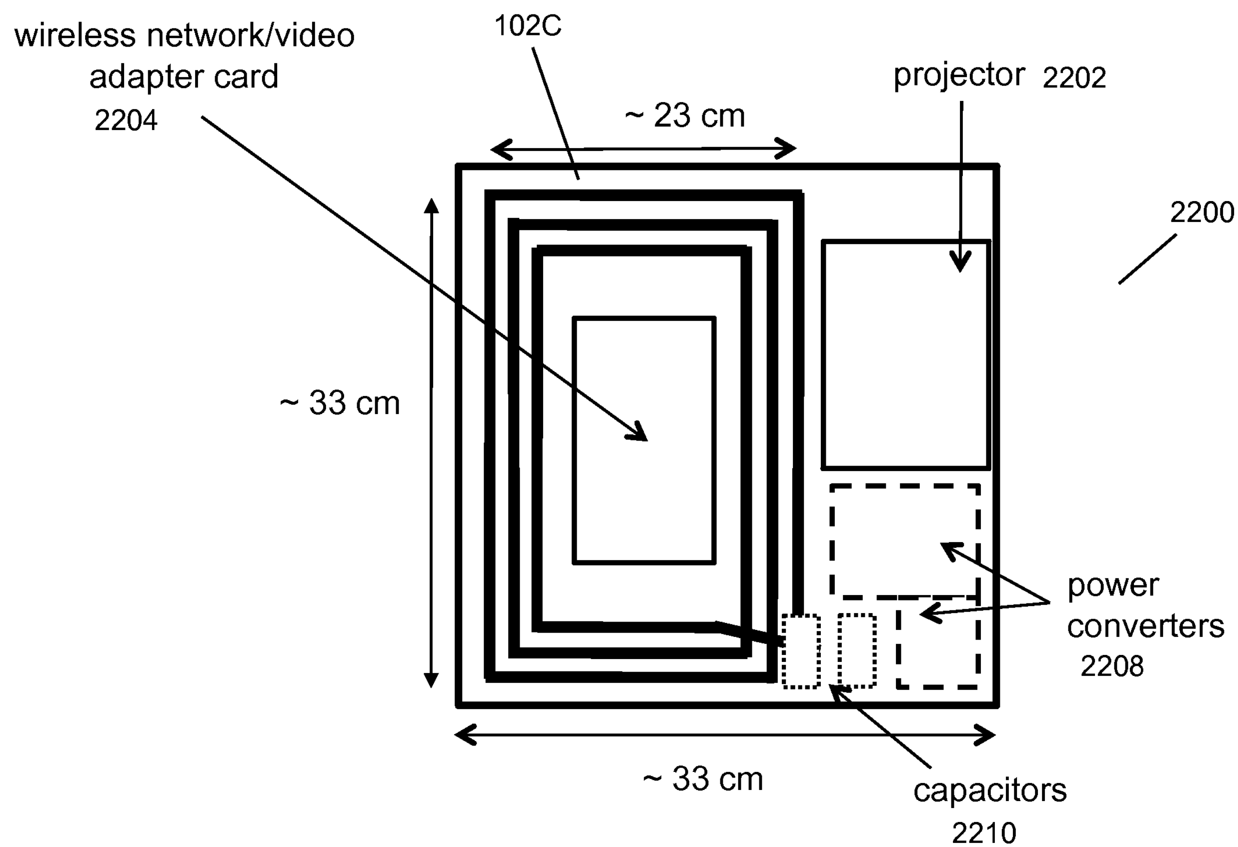

wireless network/video adapter card 2204
102C
~ 23 cm
projector 2202
2200
~ 33 cm
power converters 2208
~ 33 cm
capacitors 2210

Fig. 23

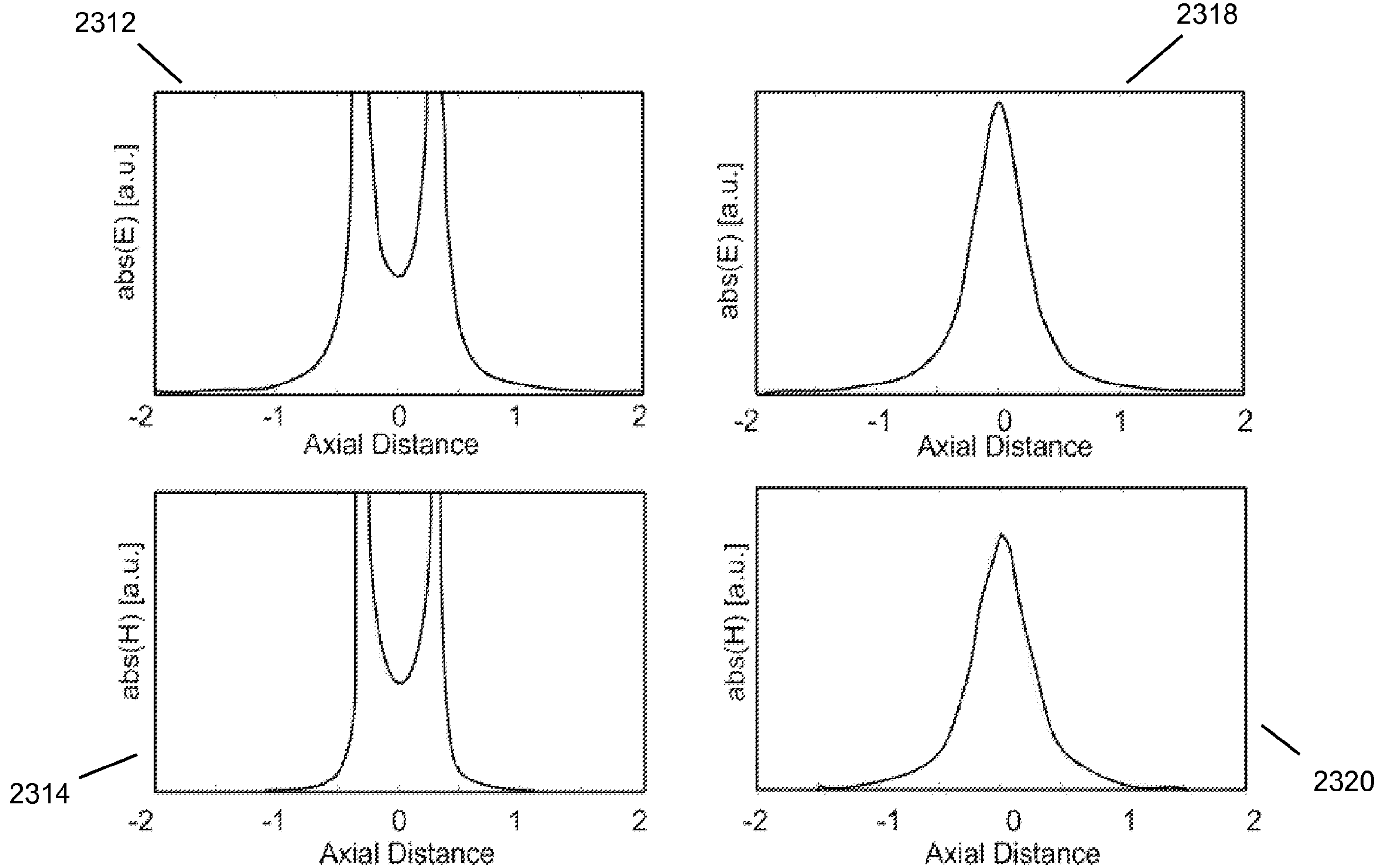
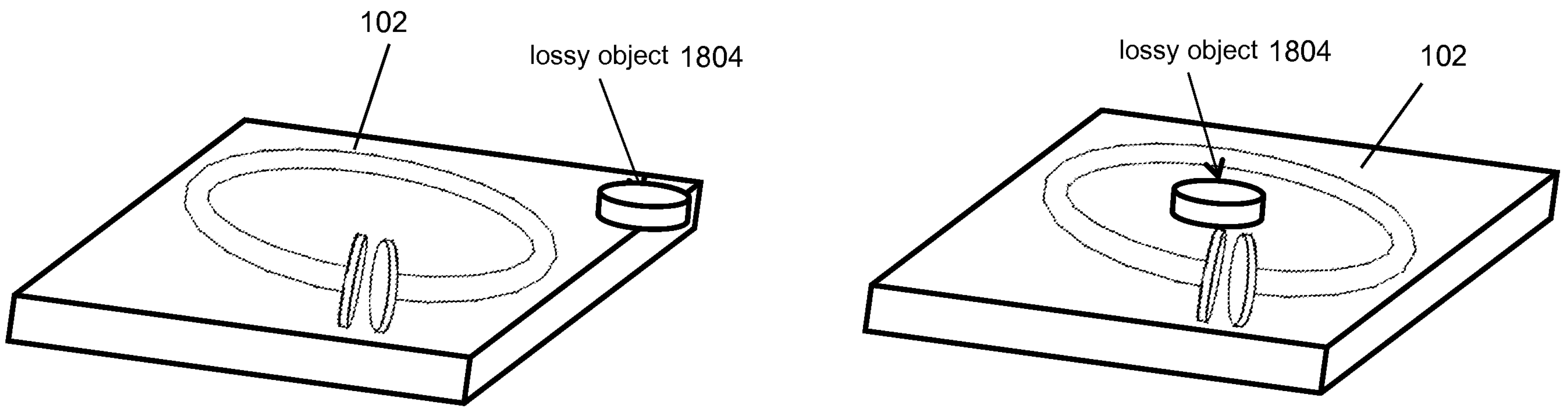


Fig. 24



(a) (b)

Fig. 25

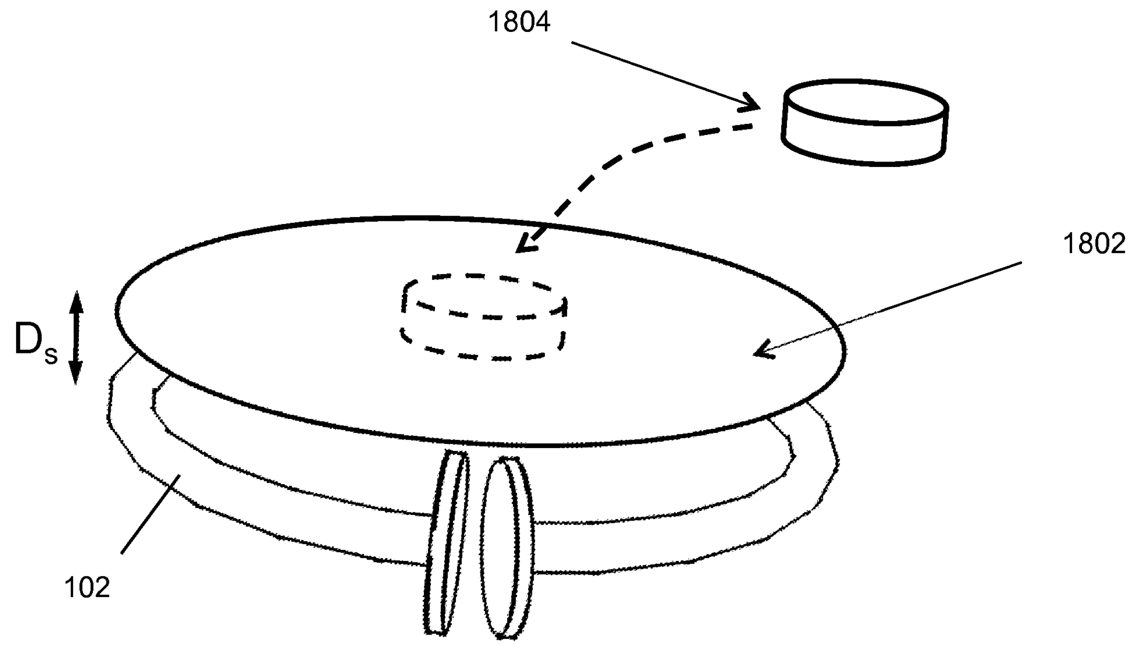
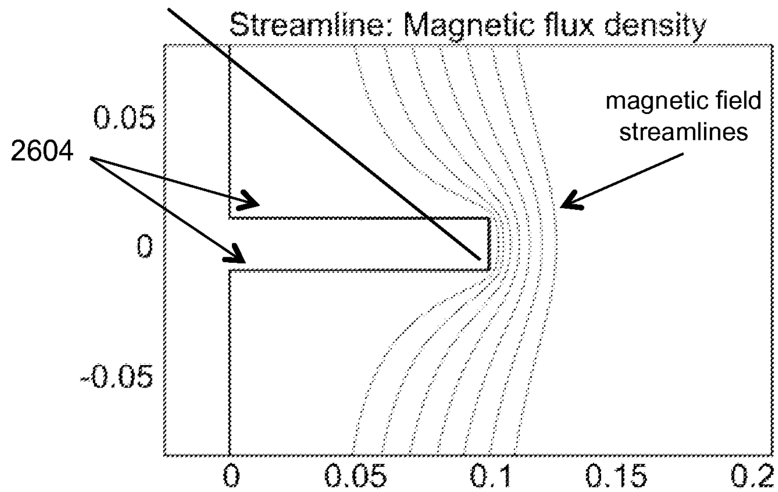
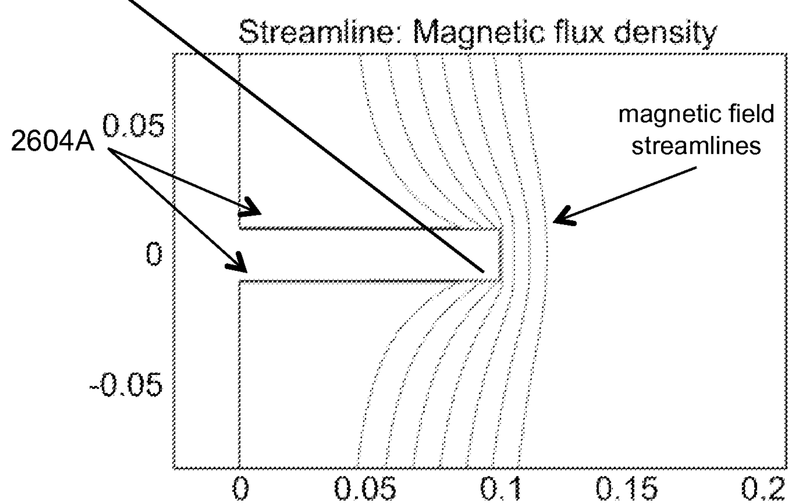


Fig. 26



2602A
Streamline: Magnetic flux density
2604
0.05
0
-0.05
magnetic field
streamlines
0
0.05
0.1
0.15
0.2

(a)



2602B
Streamline: Magnetic flux density
2604A
0.05
0
-0.05
magnetic field
streamlines
0
0.05
0.1
0.15
0.2

(b)

Fig. 27

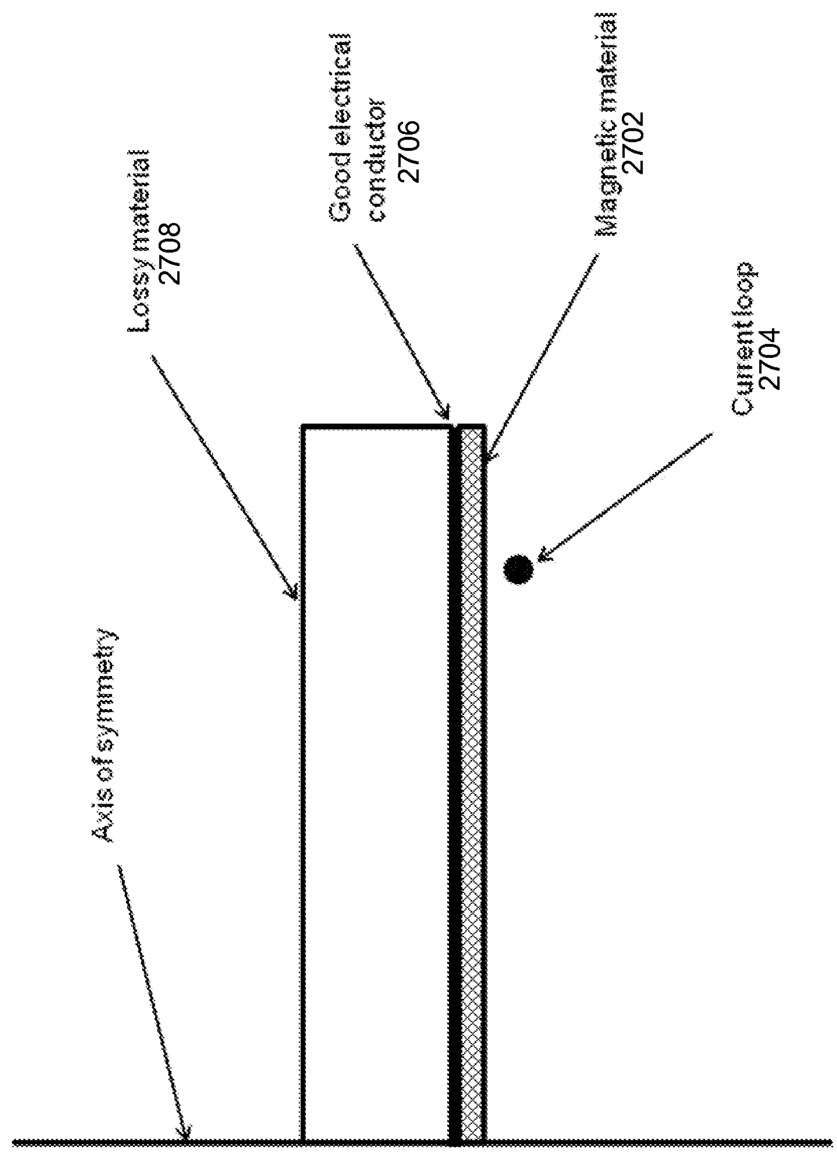
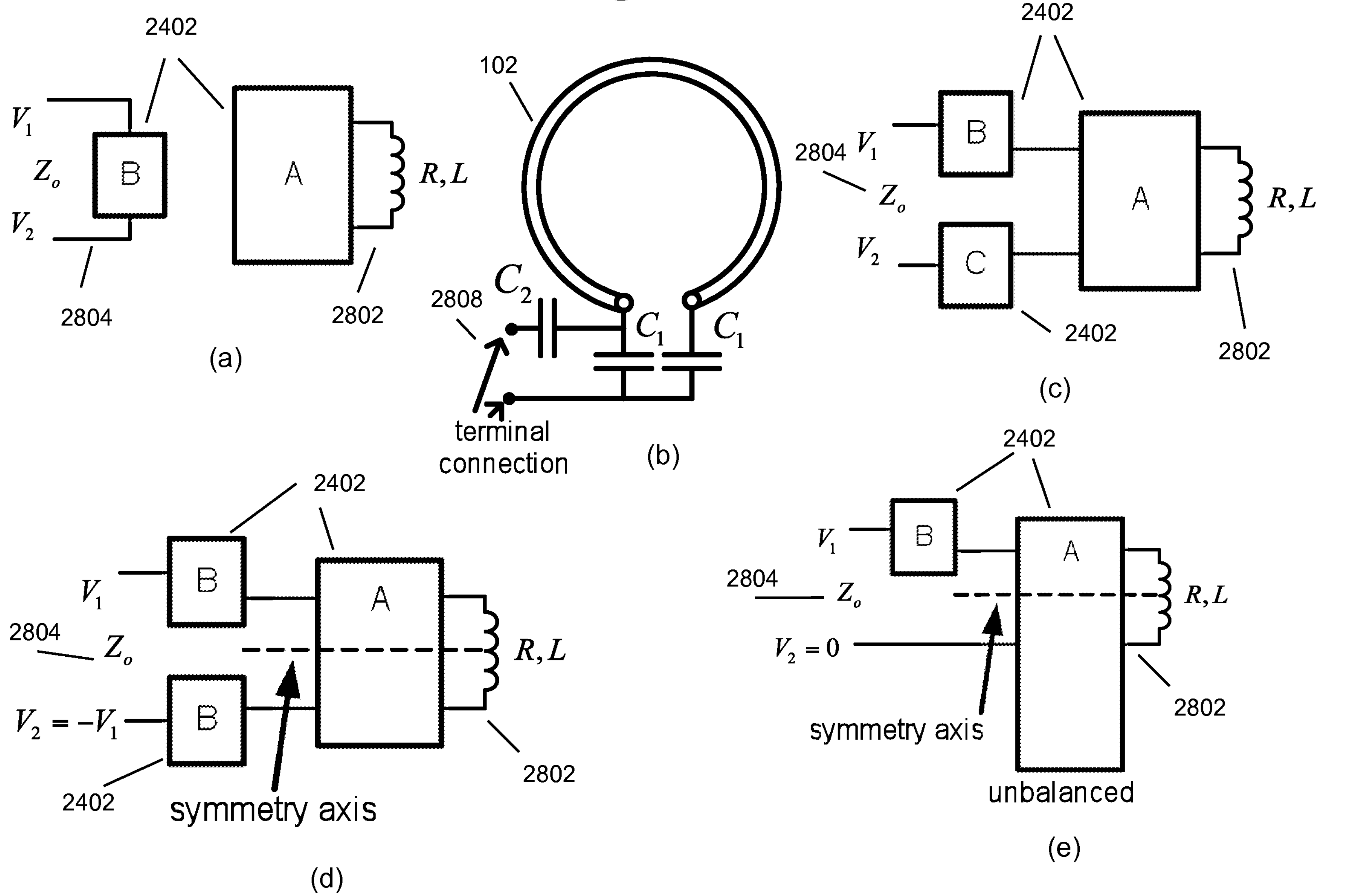

Axis of symmetry
Lossy material
2708
Good electrical
conductor
2706
Magnetic material
2702
Current loop
2704

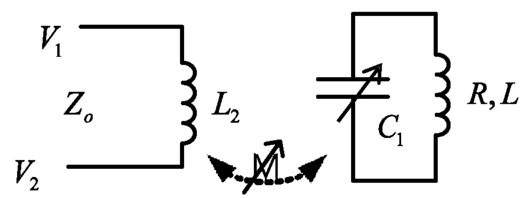
Fig. 28



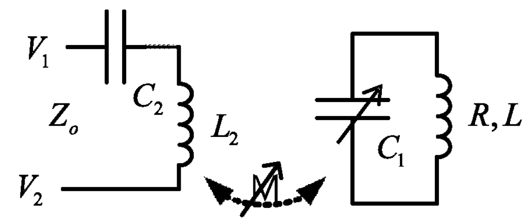
2402
V1
Zo
B
V2
A
R,L
2804
2802
(a)
102
C2
2808
C1
C1
terminal
connection
(b)
2402
2804 V1
B
Zo
A
R,L
V2
C
2402
2802
(c)
2402
B
V1
A
2804 Zo
R,L
V2 = −V1
B
2402
2802
symmetry axis
(d)
2402
V1
B
A
2804 Zo
R,L
V2 = 0
2802
symmetry axis
unbalanced
(e)

Fig. 29

(a)



(b)



(c)

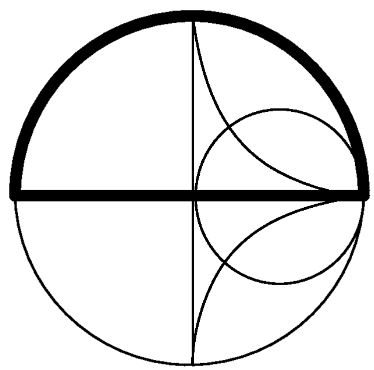
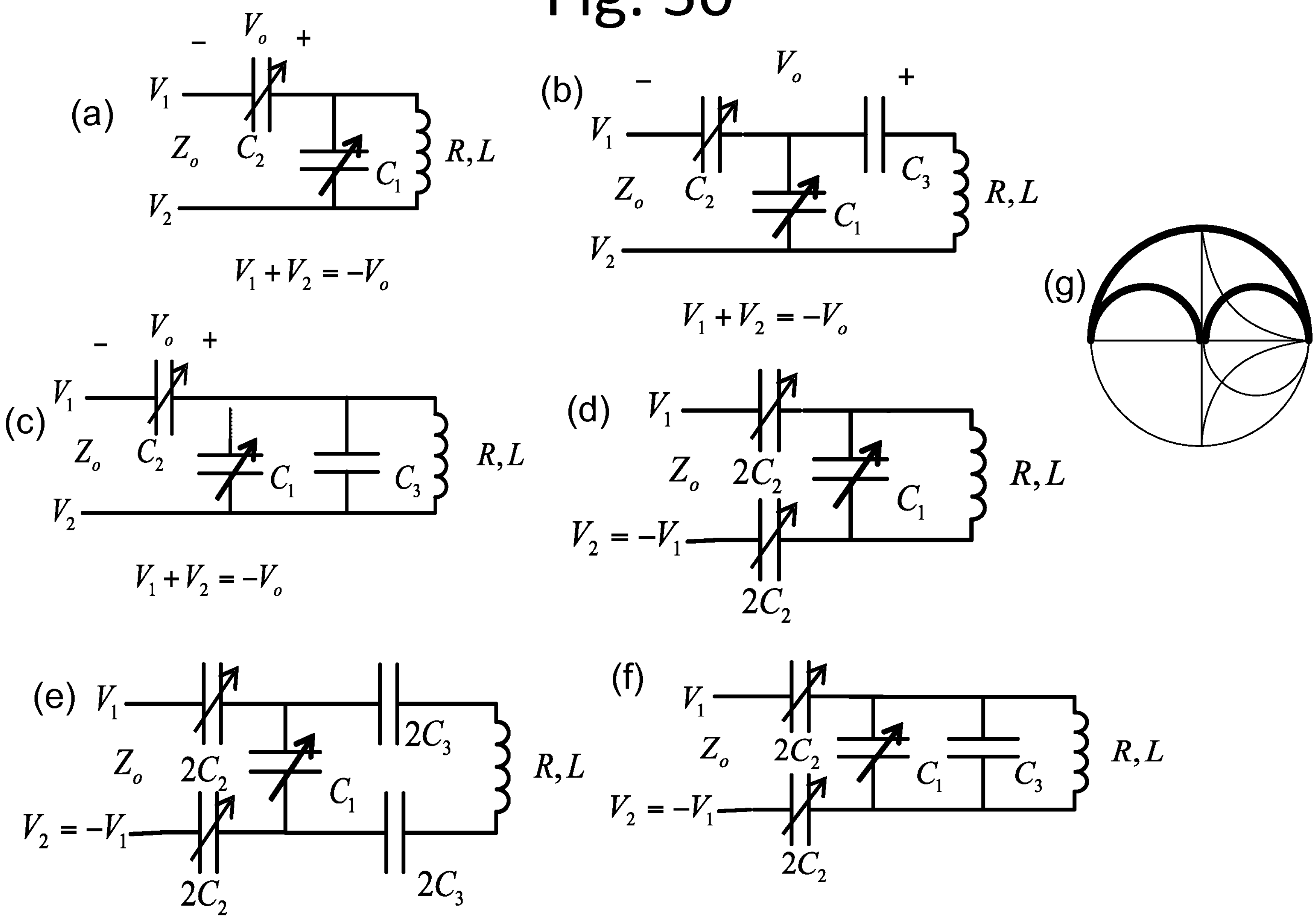


Fig. 30



(a)
V_o
V_1
Z_o
C_2
C_1
R, L
V_2
$V_1 + V_2 = -V_o$
(b)
V_1
Z_o
C_2
C_1
C_3
R, L
V_2
$V_1 + V_2 = -V_o$
(c)
V_1
Z_o
C_2
C_1
C_3
R, L
V_2
$V_1 + V_2 = -V_o$
(d)
V_1
Z_o
$2C_2$
C_1
R, L
$V_2 = -V_1$
$2C_2$
(e)
V_1
Z_o
$2C_2$
C_1
$2C_3$
R, L
$V_2 = -V_1$
$2C_2$
$2C_3$
(f)
V_1
Z_o
$2C_2$
C_1
C_3
R, L
$V_2 = -V_1$
$2C_2$
(g)

Fig. 30

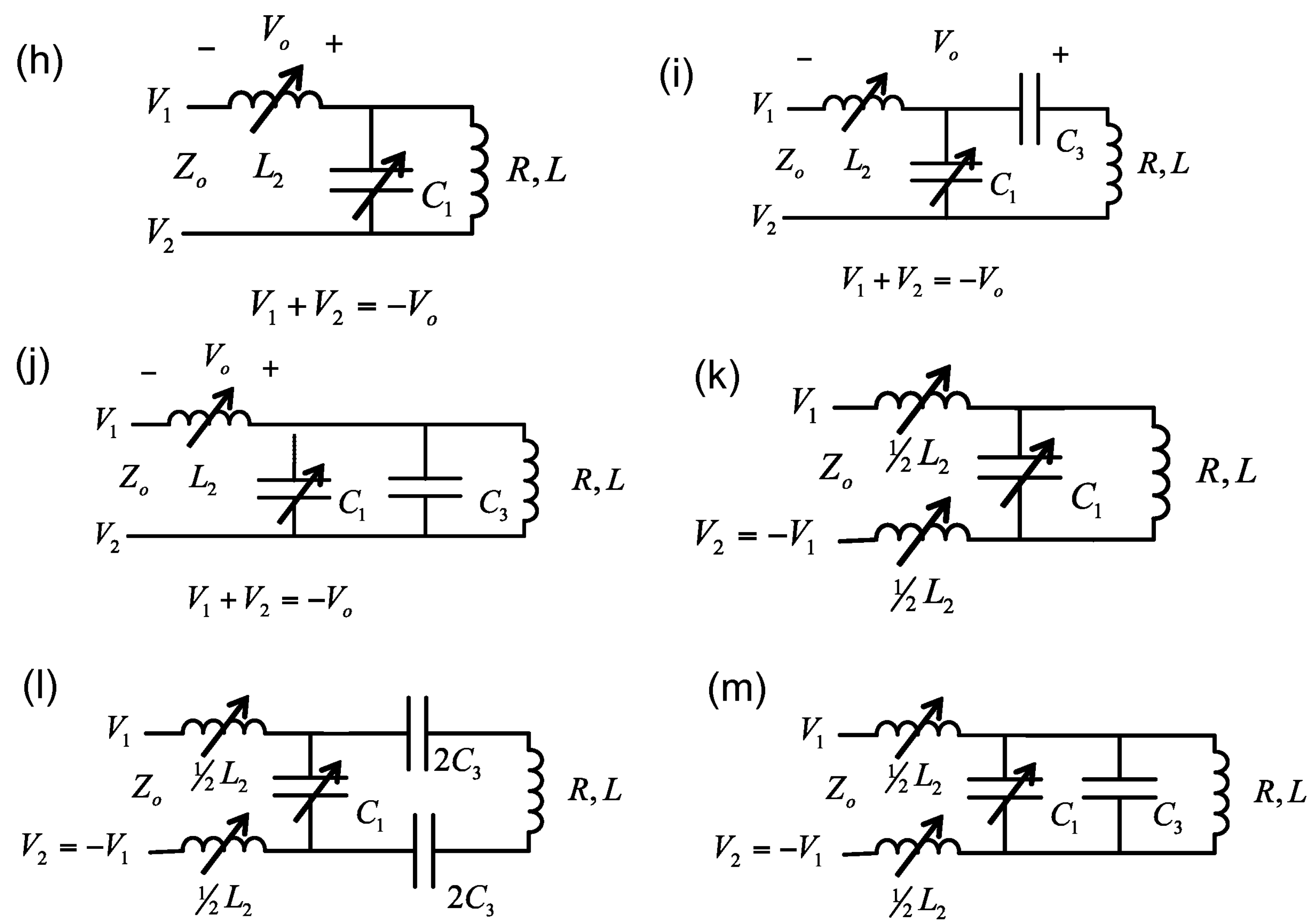
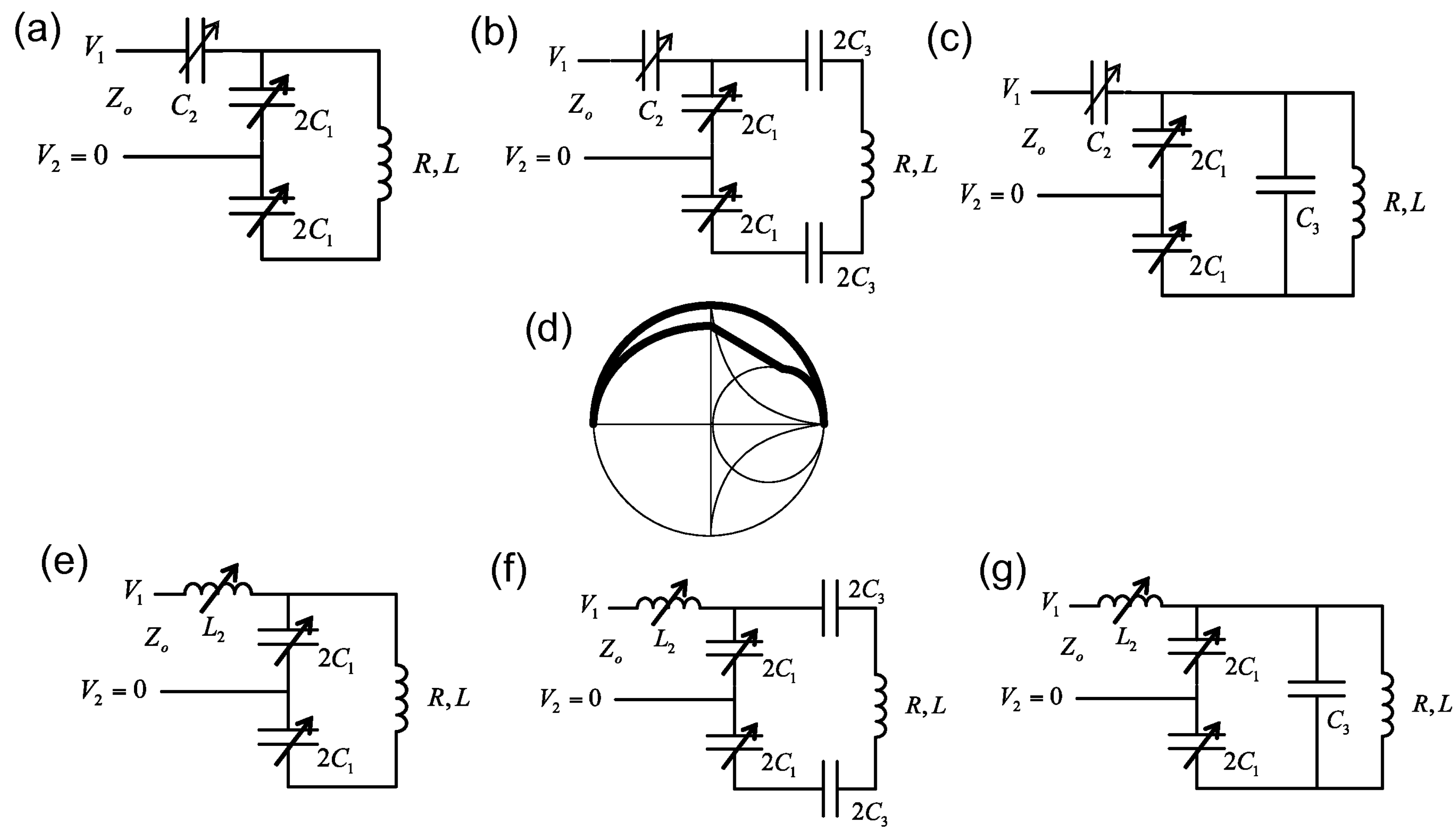
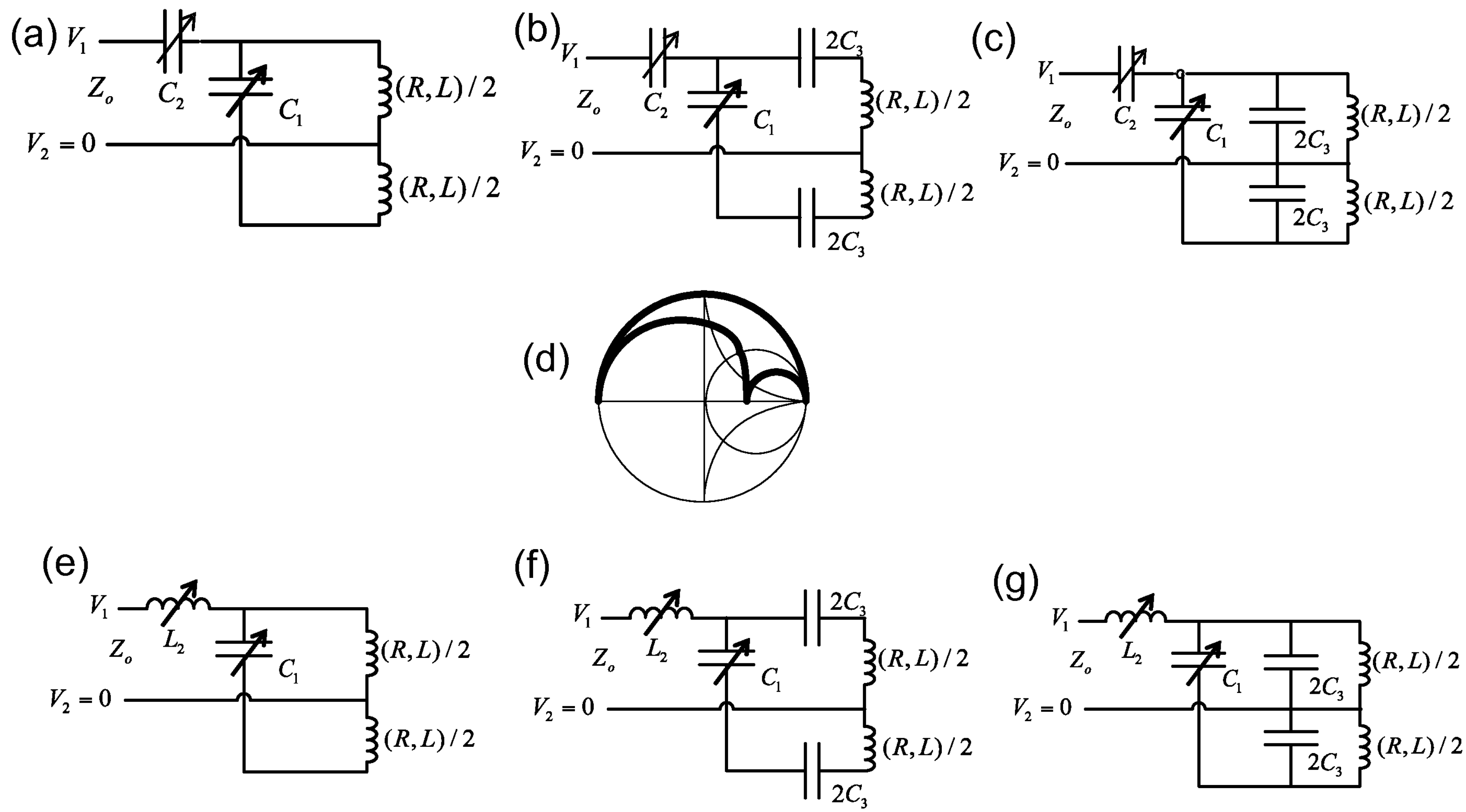

(h)
V_o
$-$ $+$
V_1
Z_o
L_2
C_1
R, L
V_2
$V_1 + V_2 = -V_o$
(i)
V_o
$-$ $+$
V_1
Z_o
L_2
C_1
C_3
R, L
V_2
$V_1 + V_2 = -V_o$
(j)
V_o
$-$ $+$
V_1
Z_o
L_2
C_1
C_3
R, L
V_2
$V_1 + V_2 = -V_o$
(k)
V_1
Z_o
$\frac{1}{2}L_2$
C_1
R, L
$V_2 = -V_1$
$\frac{1}{2}L_2$
(l)
V_1
Z_o
$\frac{1}{2}L_2$
C_1
$2C_3$
R, L
$V_2 = -V_1$
$\frac{1}{2}L_2$
$2C_3$
(m)
V_1
Z_o
$\frac{1}{2}L_2$
C_1
C_3
R, L
$V_2 = -V_1$
$\frac{1}{2}L_2$

Fig. 31


(a)
V_1
Z_o
C_2
$2C_1$
$V_2 = 0$
R, L
$2C_1$
(b)
V_1
Z_o
C_2
$2C_1$
$2C_3$
$V_2 = 0$
R, L
$2C_1$
$2C_3$
(c)
V_1
Z_o
C_2
$2C_1$
$V_2 = 0$
C_3
R, L
$2C_1$
(d)
(e)
V_1
Z_o
L_2
$2C_1$
$V_2 = 0$
R, L
$2C_1$
(f)
V_1
Z_o
L_2
$2C_1$
$2C_3$
$V_2 = 0$
R, L
$2C_1$
$2C_3$
(g)
V_1
Z_o
L_2
$2C_1$
$V_2 = 0$
C_3
R, L
$2C_1$

32

Fig. 32



33

Fig. 33

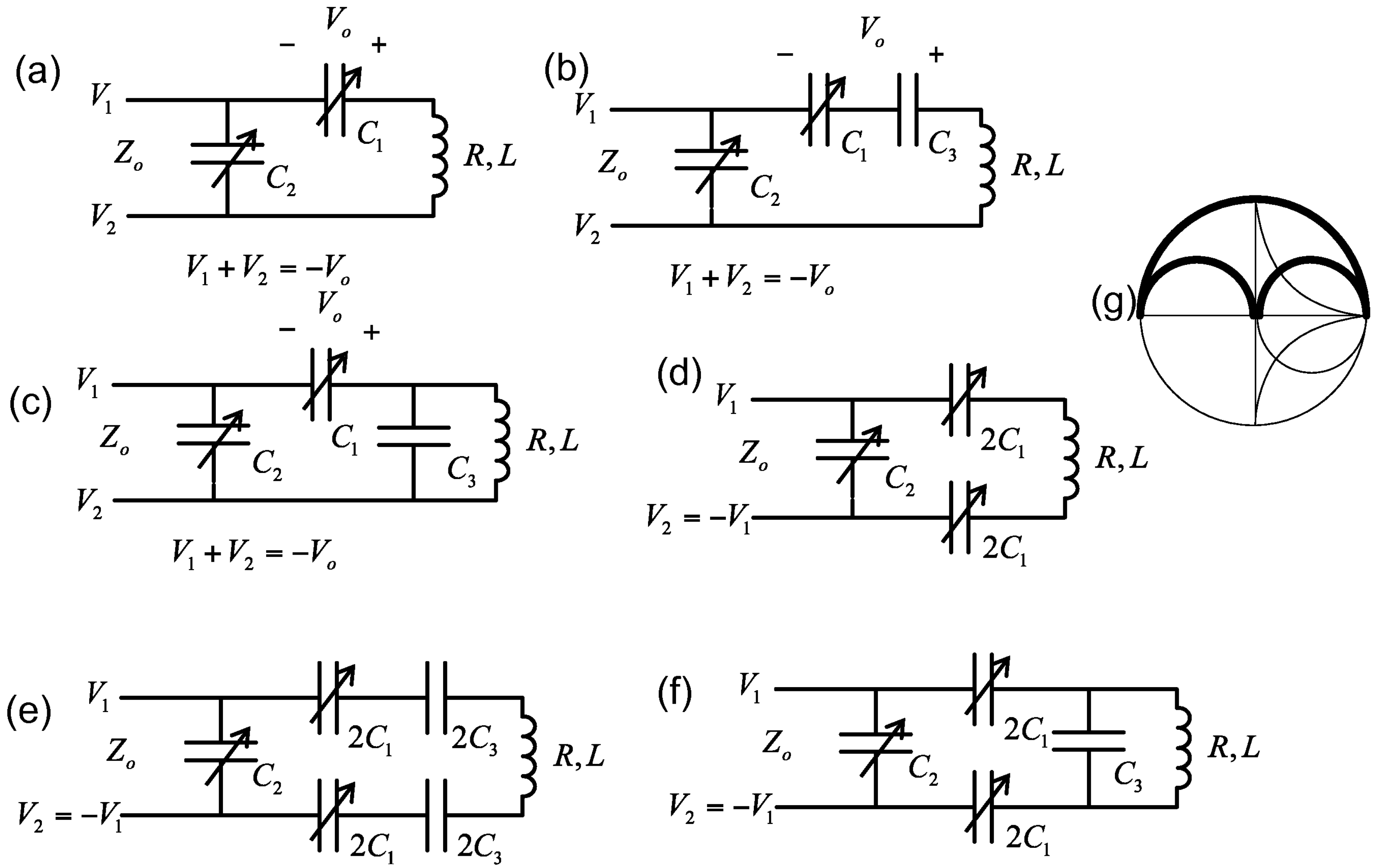

(a)
V_o
$-$ +
V_1
Z_o
V_2
C_1
C_2
R, L
$V_1 + V_2 = -V_o$
(b)
V_o
$-$ +
V_1
Z_o
V_2
C_1
C_3
C_2
R, L
$V_1 + V_2 = -V_o$
(c)
V_o
$-$ +
V_1
Z_o
V_2
C_1
C_2
C_3
R, L
$V_1 + V_2 = -V_o$
(d)
V_1
Z_o
$V_2 = -V_1$
C_2
$2C_1$
$2C_1$
R, L
(e)
V_1
Z_o
$V_2 = -V_1$
C_2
$2C_1$
$2C_3$
$2C_1$
$2C_3$
R, L
(f)
V_1
Z_o
$V_2 = -V_1$
C_2
$2C_1$
C_3
$2C_1$
R, L
(g)

Fig. 33

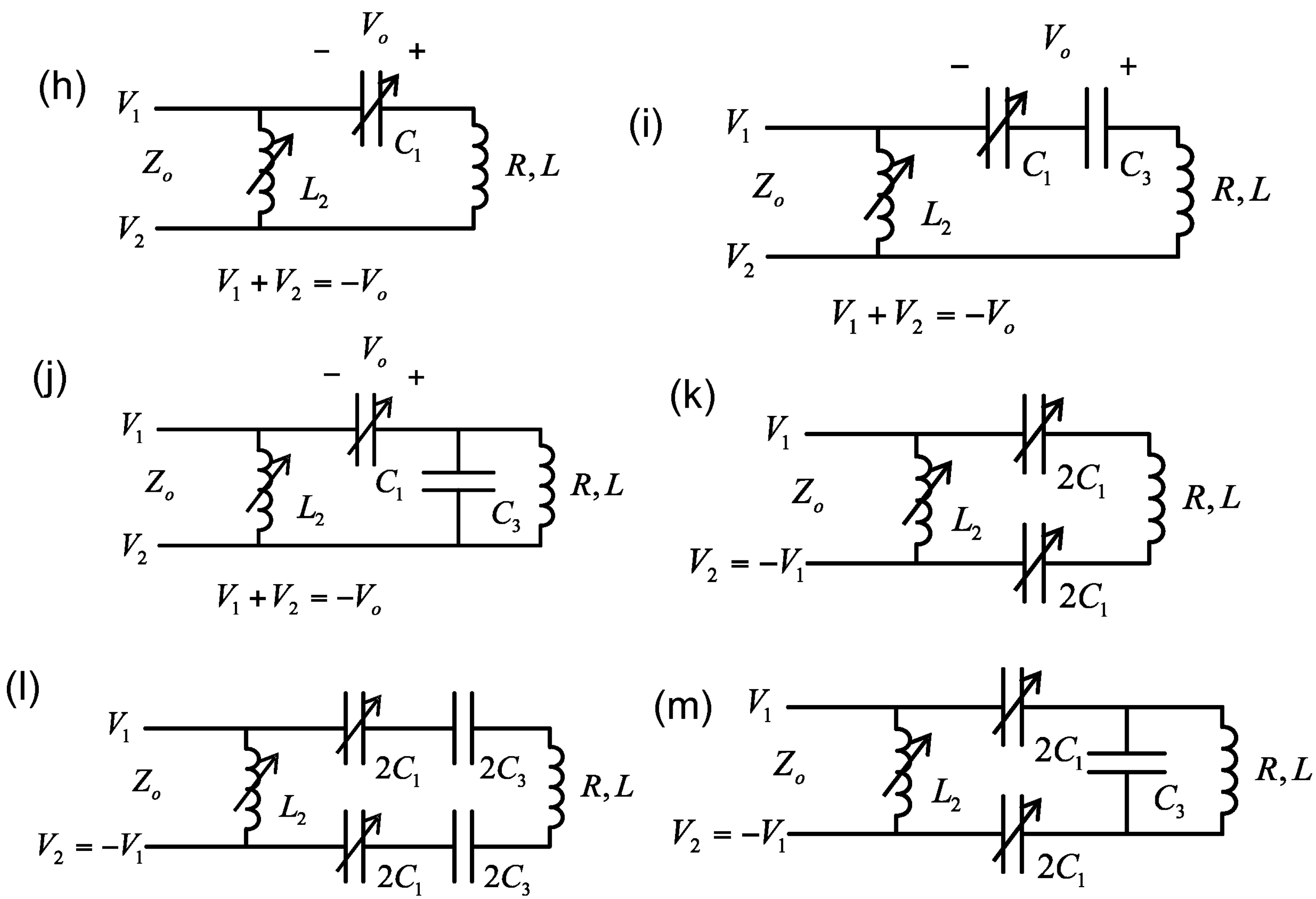

(h)
V_1
Z_o
V_2
V_o
– +
C_1
L_2
R, L
$V_1 + V_2 = -V_o$
(i)
V_1
Z_o
V_2
V_o
– +
C_1
C_3
L_2
R, L
$V_1 + V_2 = -V_o$
(j)
V_1
Z_o
V_2
V_o
– +
C_1
C_3
L_2
R, L
$V_1 + V_2 = -V_o$
(k)
V_1
Z_o
$V_2 = -V_1$
L_2
$2C_1$
$2C_1$
R, L
(l)
V_1
Z_o
$V_2 = -V_1$
L_2
$2C_1$
$2C_3$
$2C_1$
$2C_3$
R, L
(m)
V_1
Z_o
$V_2 = -V_1$
L_2
$2C_1$
C_3
$2C_1$
R, L

Fig. 34

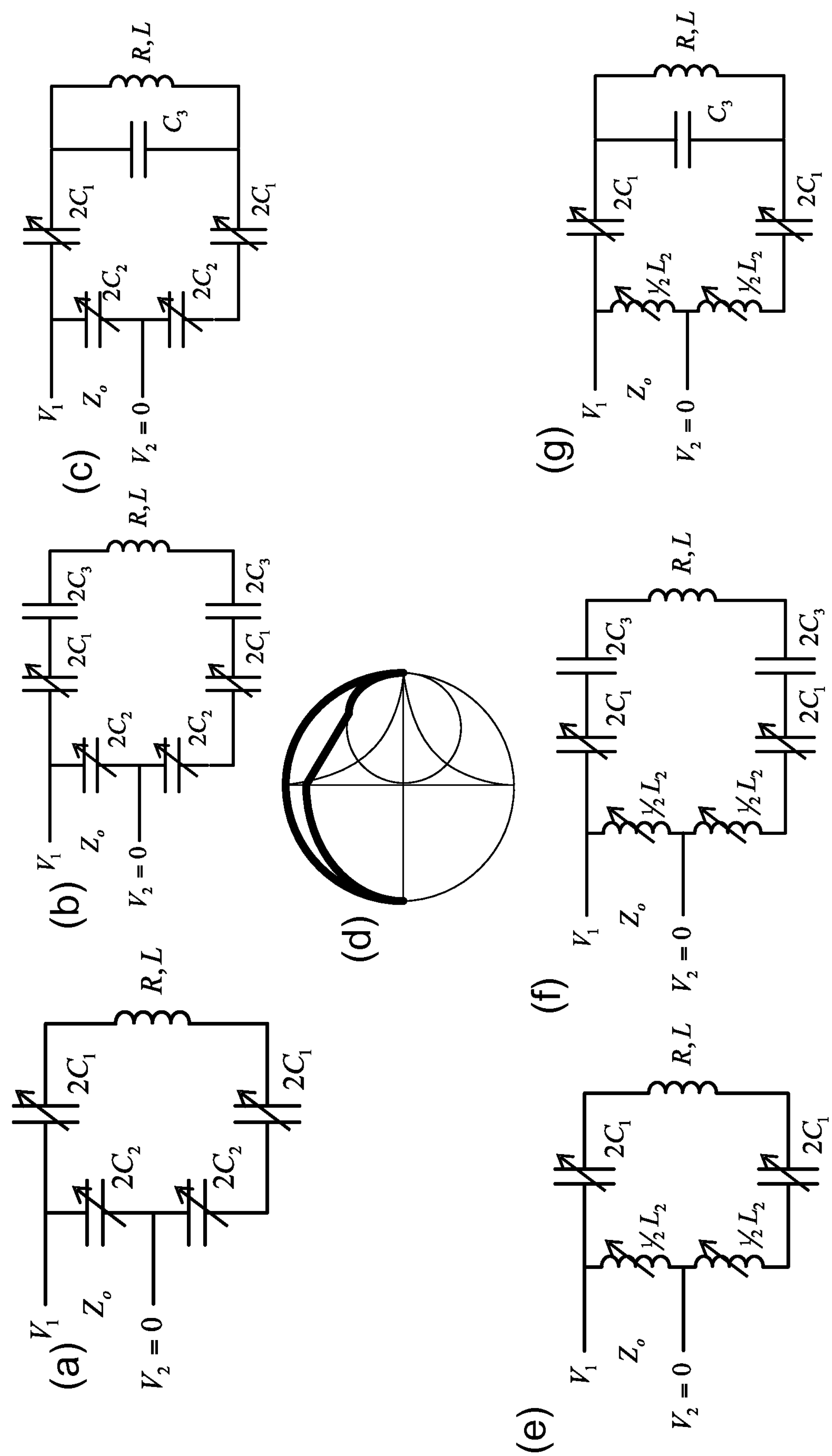
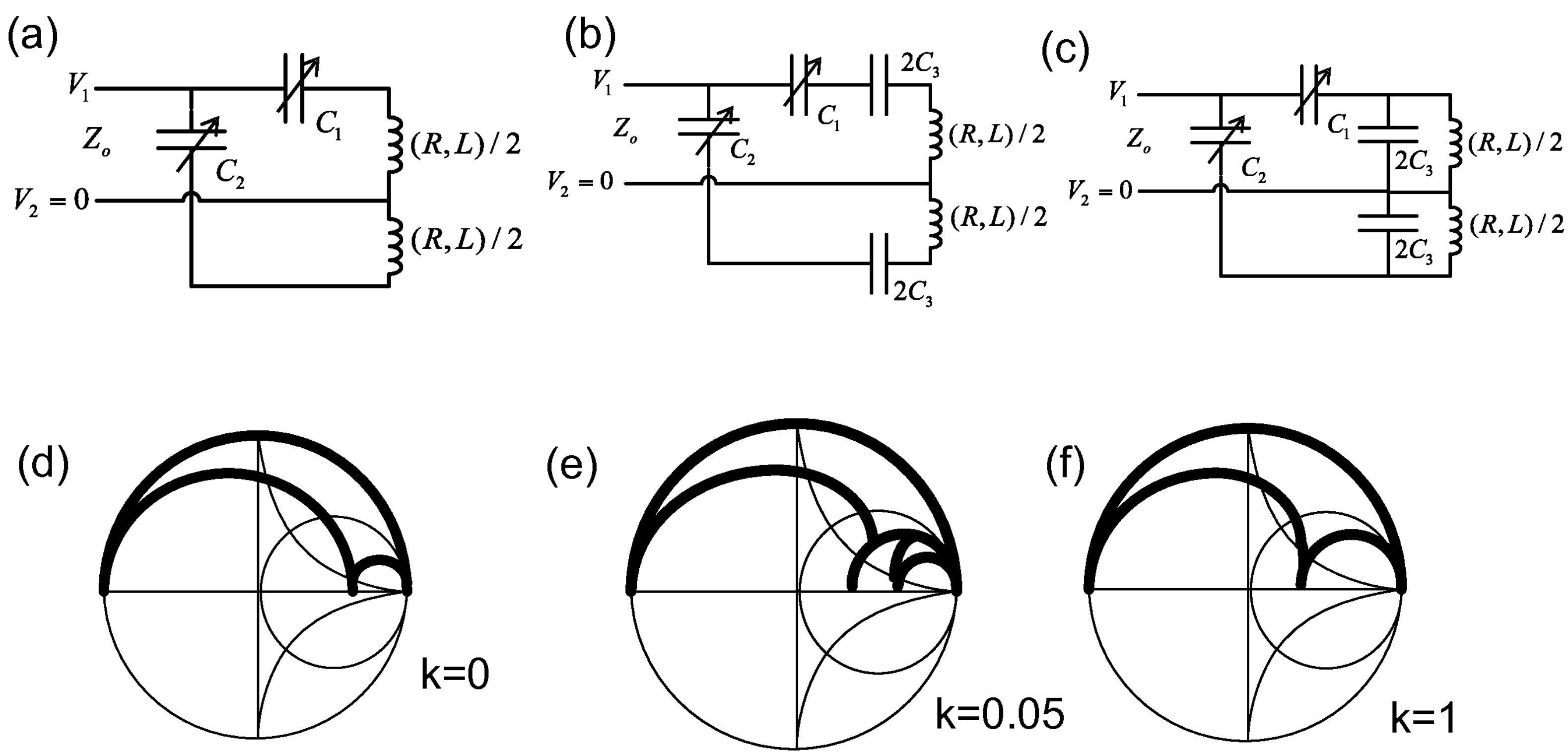


Fig. 35



(a)
V_1
Z_o
$V_2 = 0$
C_1
C_2
$(R,L)/2$
$(R,L)/2$
(b)
V_1
Z_o
$V_2 = 0$
C_1
C_2
$2C_3$
$2C_3$
$(R,L)/2$
$(R,L)/2$
(c)
V_1
Z_o
$V_2 = 0$
C_1
C_2
$2C_3$
$2C_3$
$(R,L)/2$
$(R,L)/2$
(d)
k=0
(e)
k=0.05
(f)
k=1

37

Fig. 36

(a) (b) (c) (d)

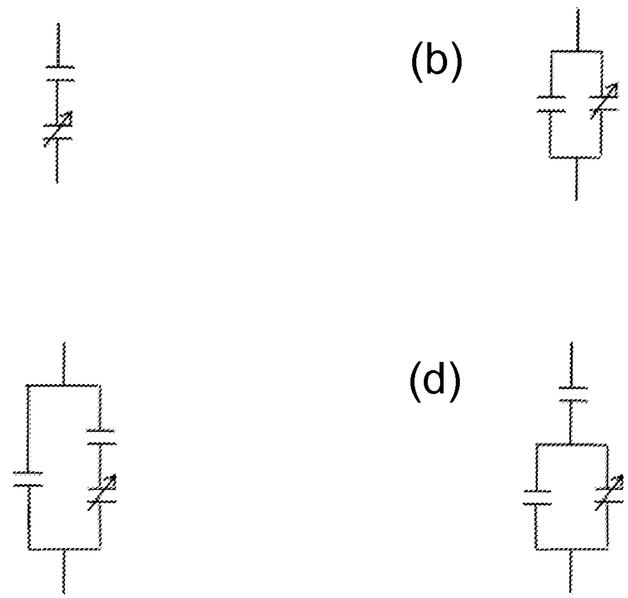


Fig. 37

(a)

(b)

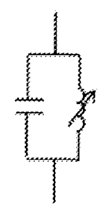


Fig. 38

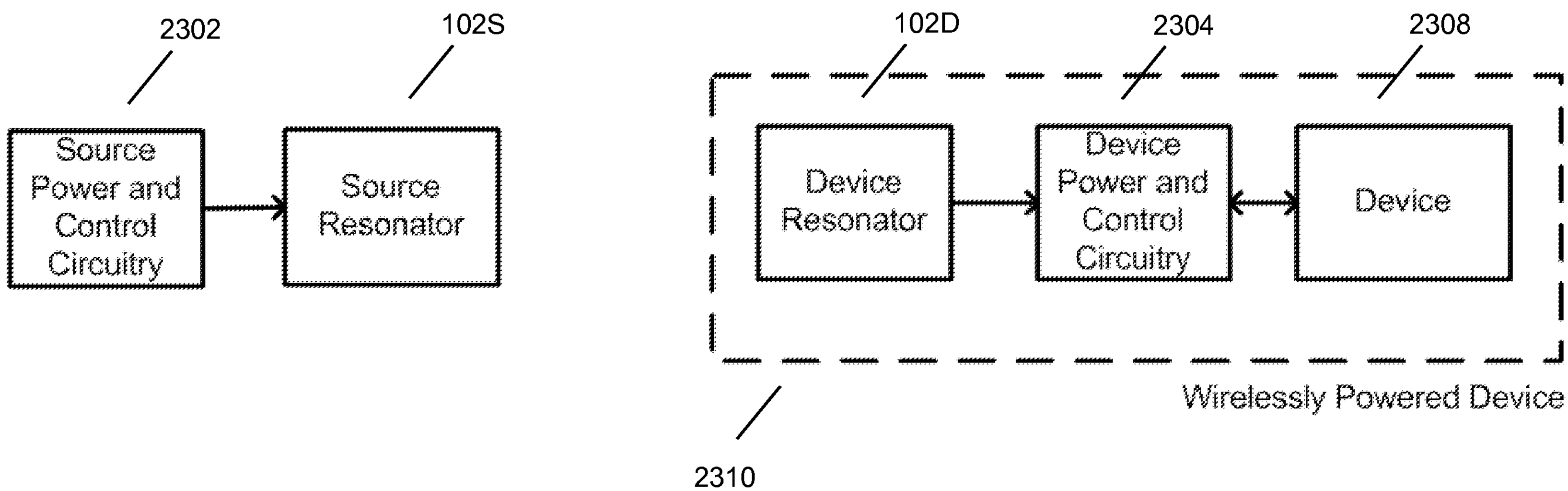

2302
102S
Source Power and Control Circuitry
Source Resonator
102D
2304
2308
Device Resonator
Device Power and Control Circuitry
Device
Wirelessly Powered Device
2310

Fig. 39

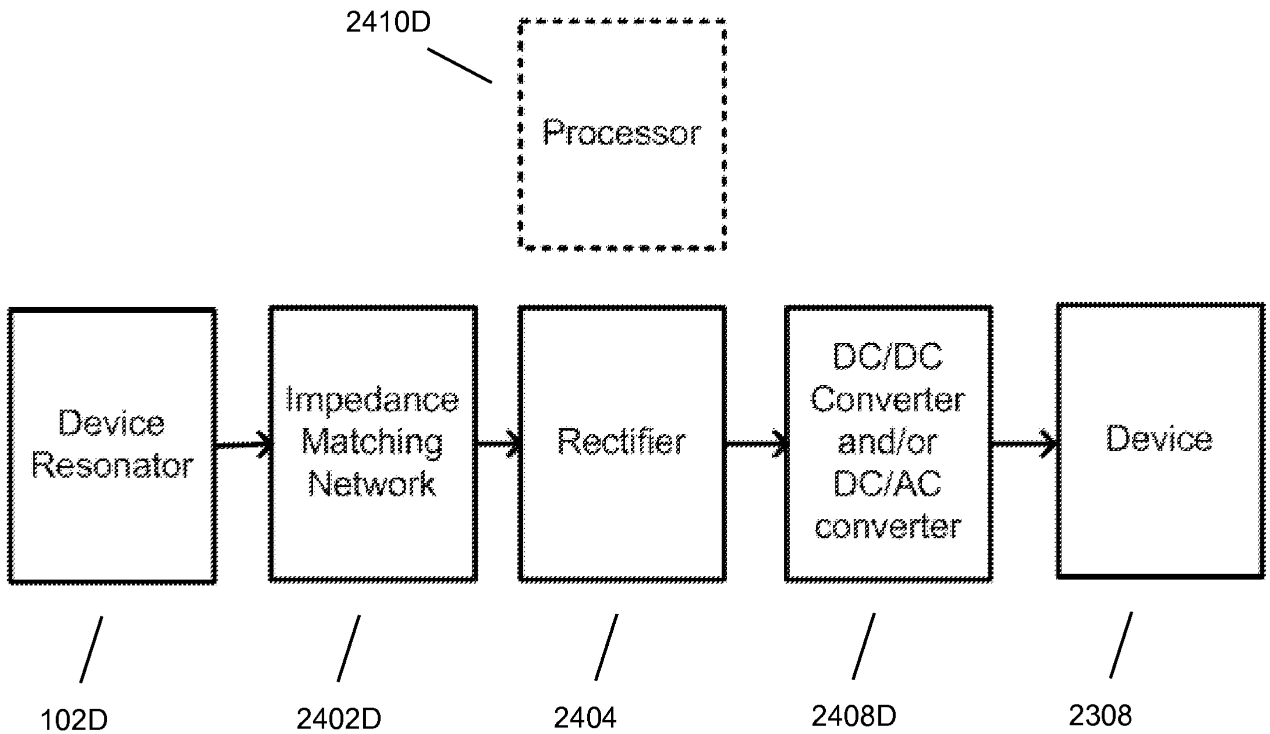

2410D
Processor
Device Resonator
Impedance Matching Network
Rectifier
DC/DC Converter and/or DC/AC converter
Device
102D
2402D
2404
2408D
2308

Fig. 40

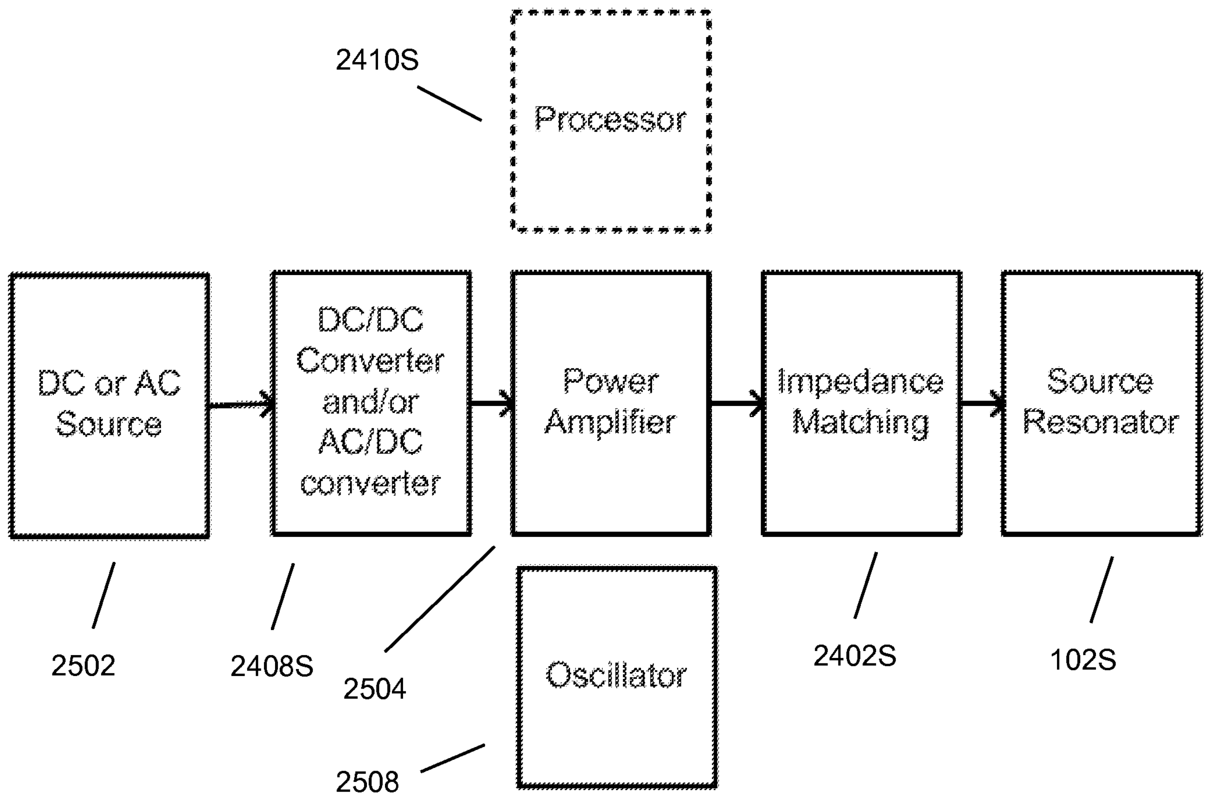

2410S
Processor
DC or AC Source
DC/DC Converter and/or AC/DC converter
Power Amplifier
Impedance Matching
Source Resonator
2502
2408S
2504
Oscillator
2402S
102S
2508

Fig. 41

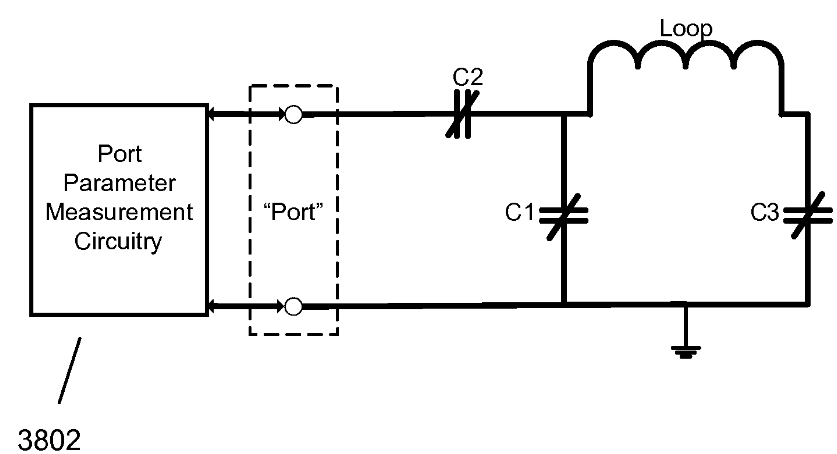
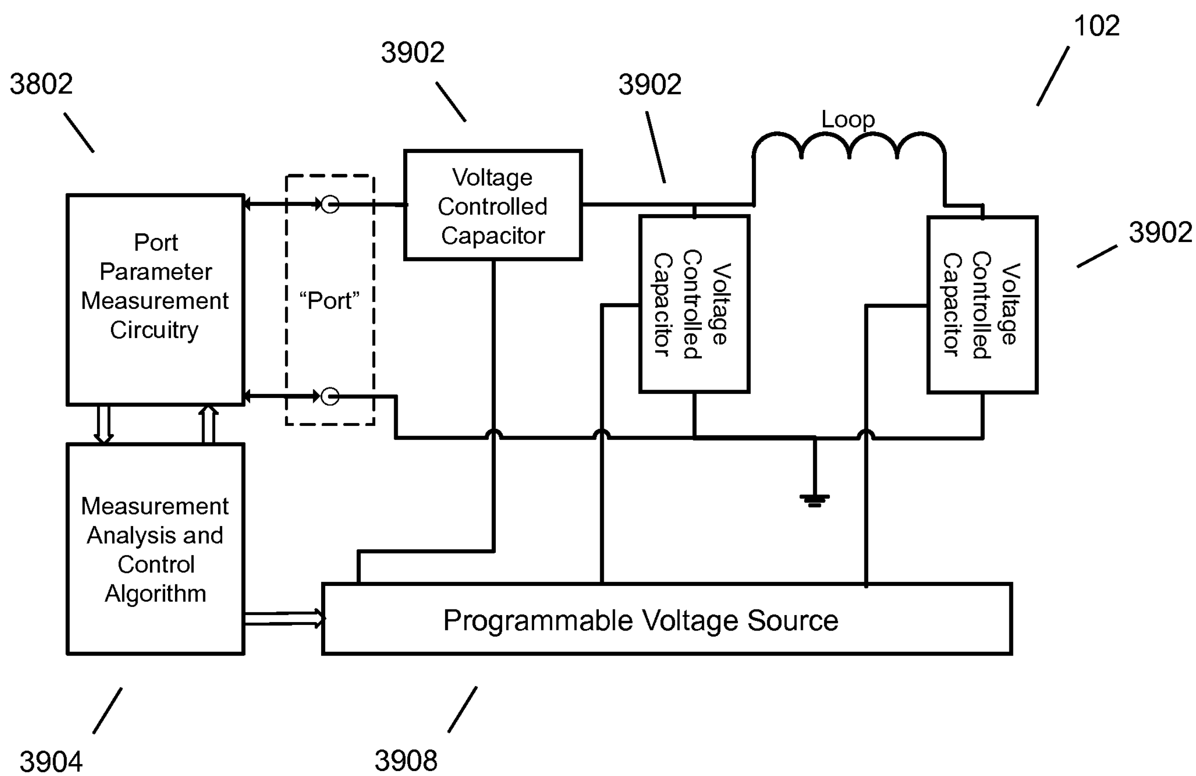

Port
Parameter
Measurement
Circuitry
"Port"
C2
Loop
C1
C3
3802

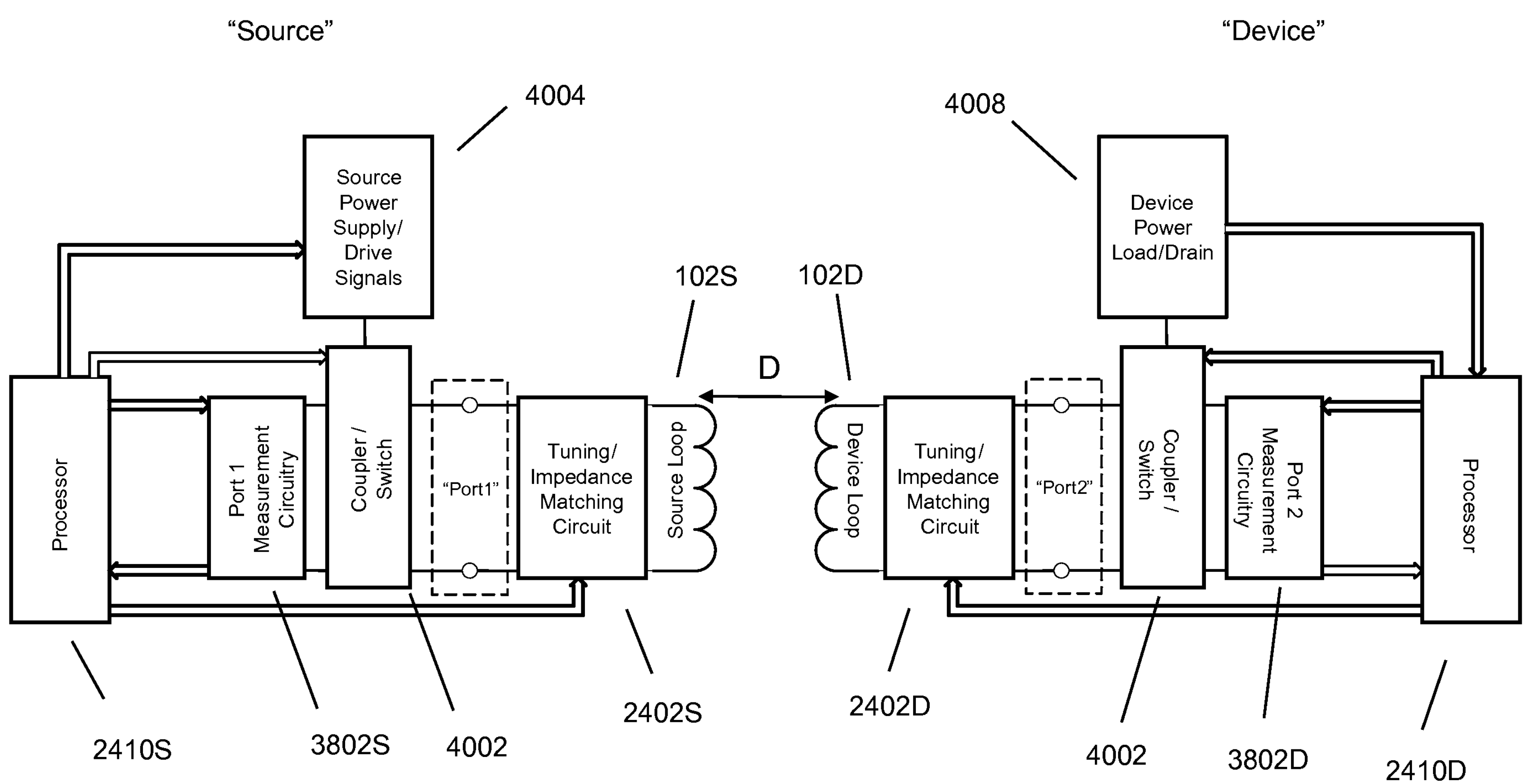
Fig. 42



3802
3902
3902
102
Loop
3902
Port
Parameter
Measurement
Circuitry
"Port"
Voltage
Controlled
Capacitor
Voltage
Controlled
Capacitor
Voltage
Controlled
Capacitor
Measurement
Analysis and
Control
Algorithm
Programmable Voltage Source
3904
3908

44

Fig. 43



"Source"
"Device"
4004
4008
Source Power Supply/ Drive Signals
Device Power Load/Drain
102S
102D
D
Processor
Port 1 Measurement Circuitry
Coupler / Switch
"Port1"
Tuning/ Impedance Matching Circuit
Source Loop
Device Loop
Tuning/ Impedance Matching Circuit
"Port2"
Coupler / Switch
Port 2 Measurement Circuitry
Processor
2410S
3802S
4002
2402S
2402D
4002
3802D
2410D

45

Fig. 44

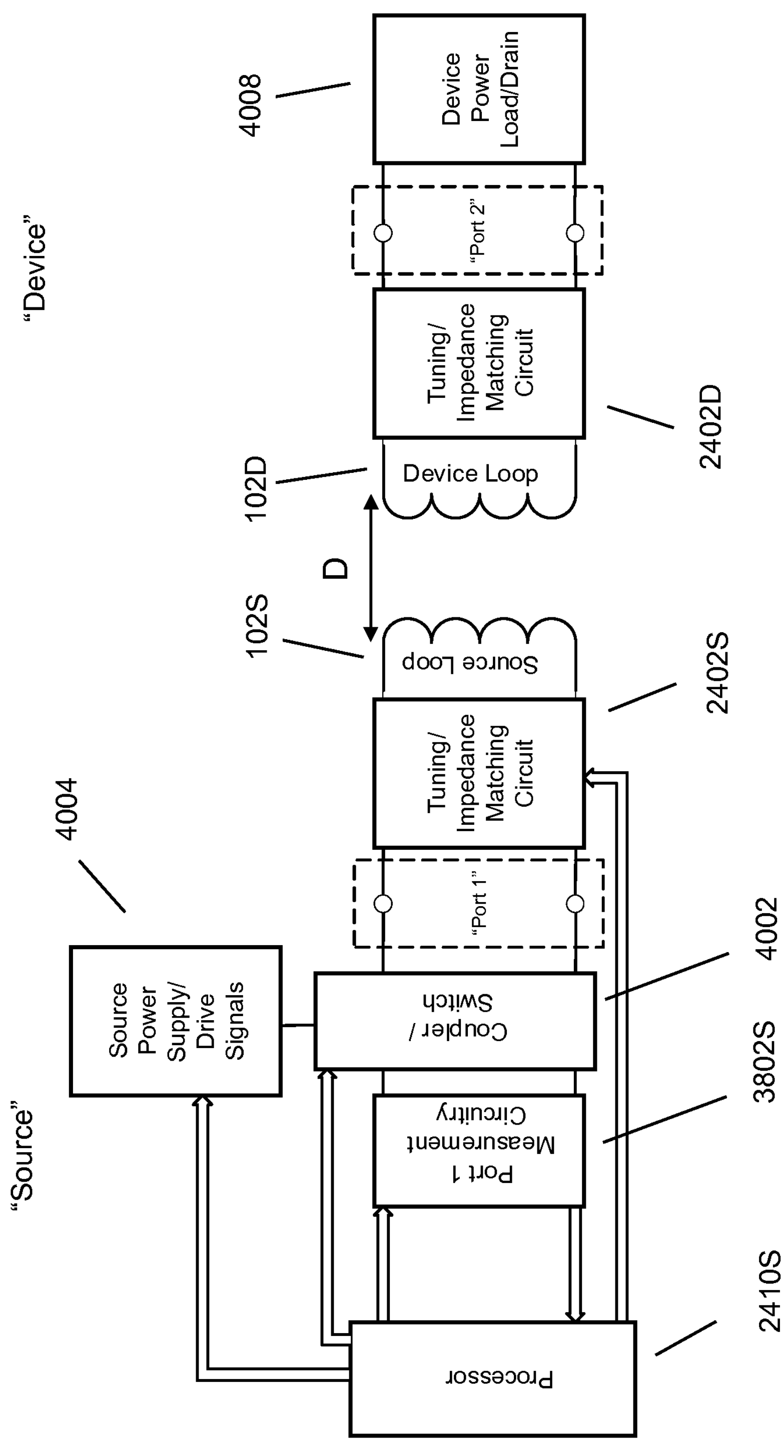
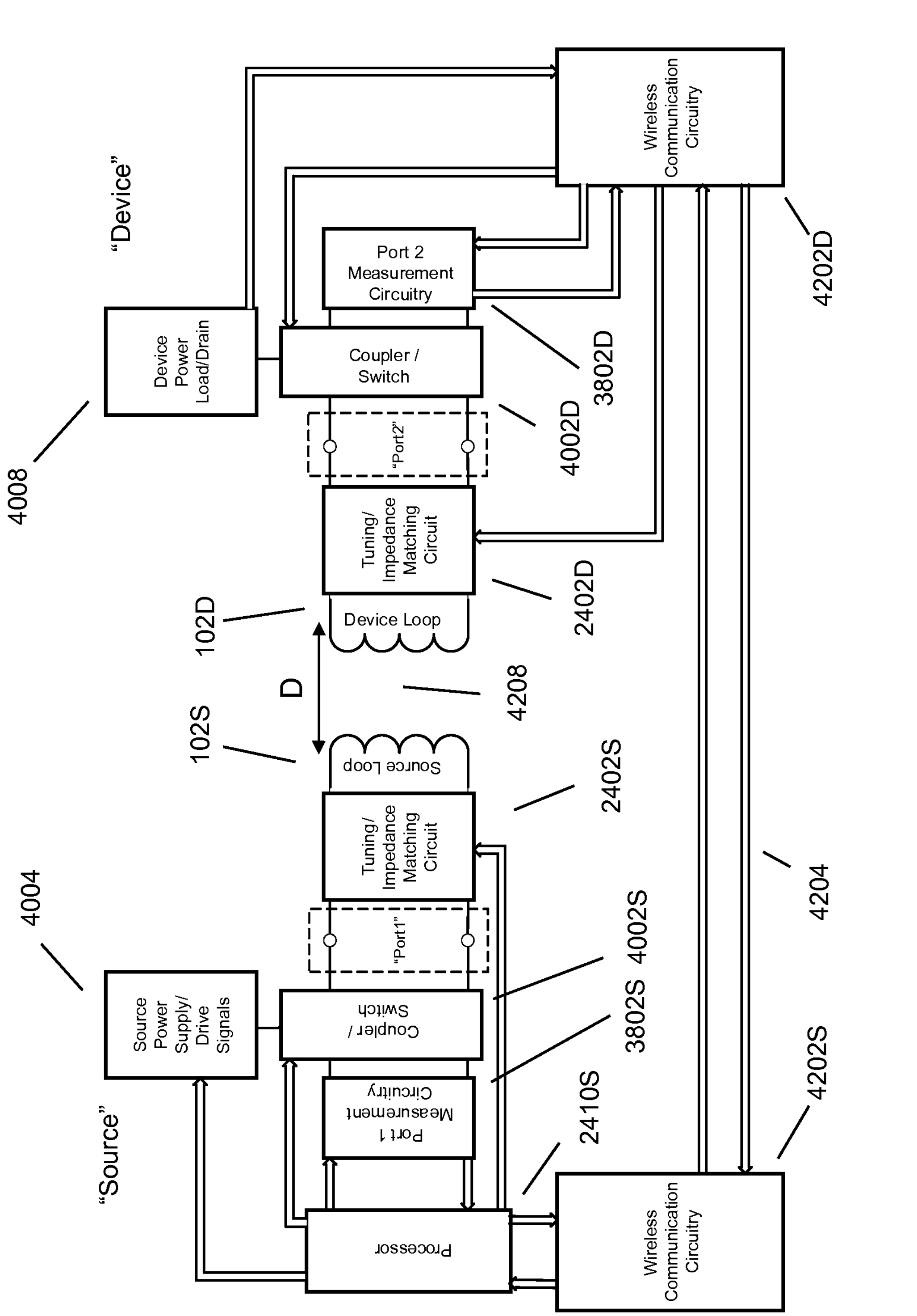
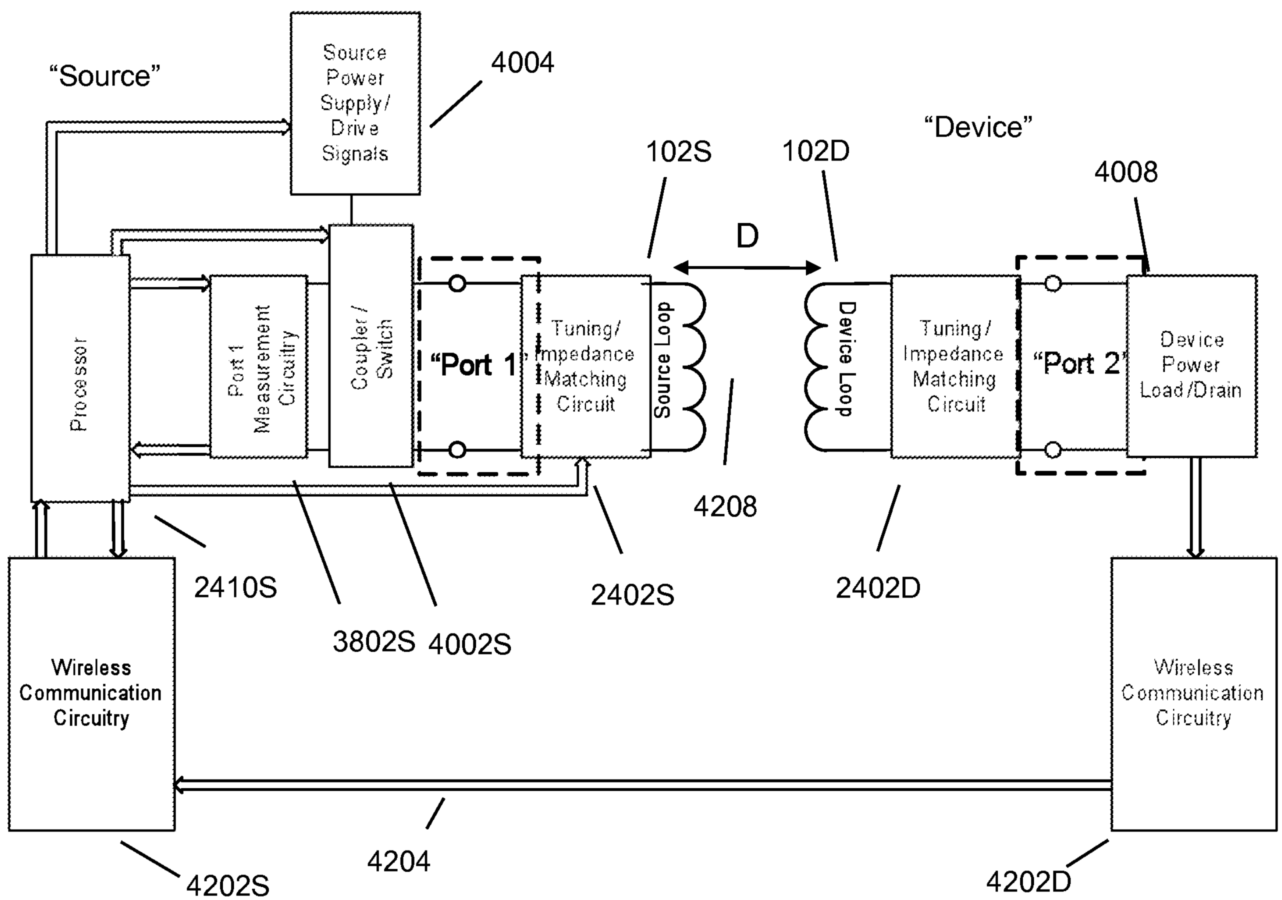

"Source"
"Device"
4004
Source Power Supply/ Drive Signals
Coupler / Switch
4002
Port 1 Measurement Circuitry
3802S
Processor
2410S
"Port 1"
Tuning/ Impedance Matching Circuit
2402S
Source Loop
102S
D
102D
Device Loop
Tuning/ Impedance Matching Circuit
2402D
"Port 2"
Device Power Load/Drain
4008

Fig. 45


"Source"
"Device"
4004
4008
Source Power Supply/ Drive Signals
Processor
Port 1 Measurement Circuitry
Coupler / Switch
"Port1"
Tuning/ Impedance Matching Circuit
Source Loop
102S
D
102D
4208
Device Loop
Tuning/ Impedance Matching Circuit
"Port2"
Coupler / Switch
Port 2 Measurement Circuitry
Device Power Load/Drain
Wireless Communication Circuitry
Wireless Communication Circuitry
2410S
3802S
4002S
2402S
2402D
4002D
3802D
4204
4202S
4202D

47

Fig. 46



"Source"
Source Power Supply/ Drive Signals
4004
"Device"
102S
102D
D
4008
Processor
Port 1 Measurement Circuitry
Coupler / Switch
"Port 1"
Tuning/ Impedance Matching Circuit
Source Loop
Device Loop
Tuning/ Impedance Matching Circuit
"Port 2"
Device Power Load/Drain
4208
2410S
3802S
4002S
2402S
2402D
Wireless Communication Circuitry
Wireless Communication Circuitry
4204
4202S
4202D

48

Fig. 47

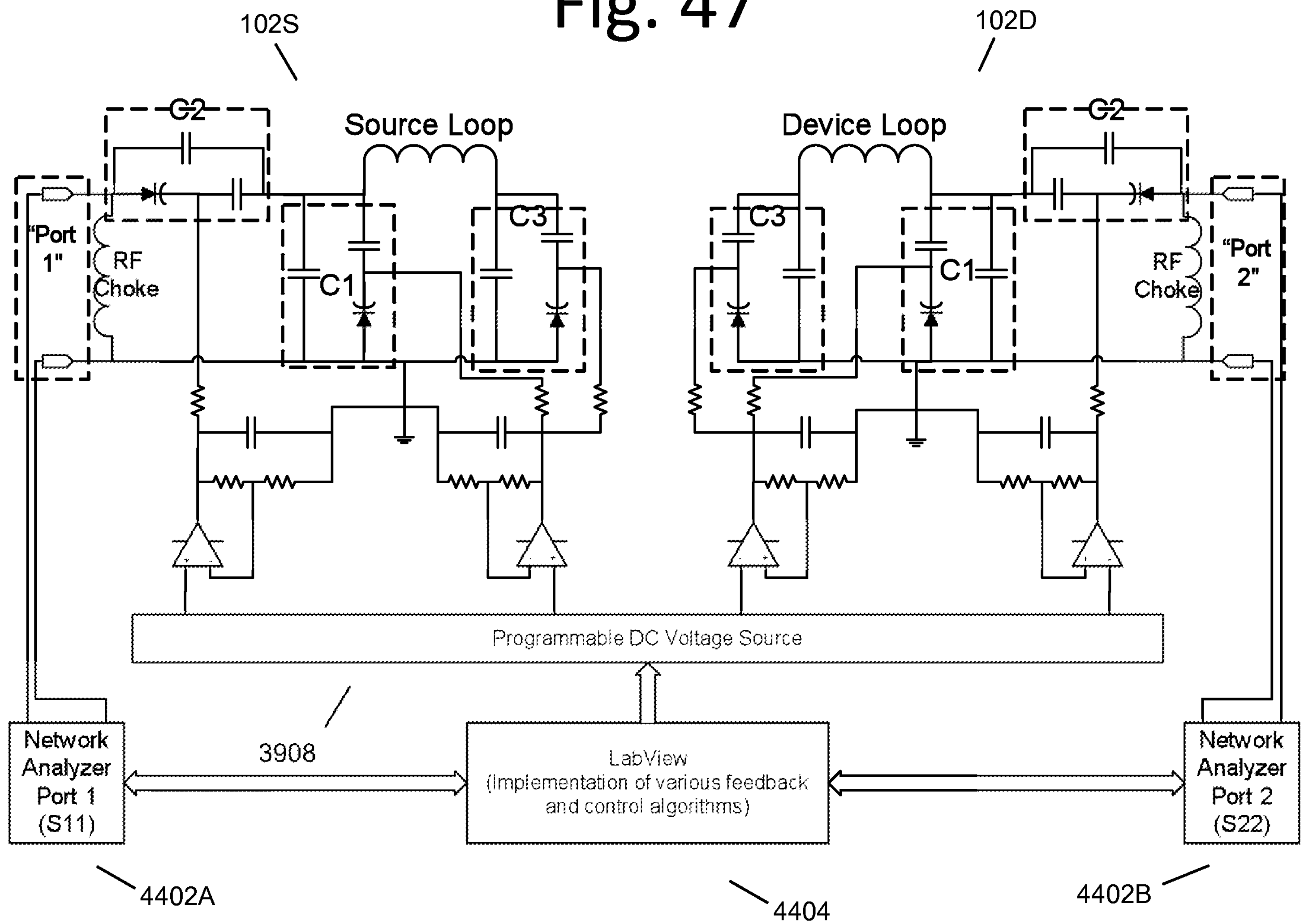

102S
102D
Source Loop
Device Loop
C2
C1
C3
"Port 1"
"Port 2"
RF Choke
Programmable DC Voltage Source
3908
Network Analyzer Port 1 (S11)
LabView (Implementation of various feedback and control algorithms)
Network Analyzer Port 2 (S22)
4402A
4404
4402B

Fig. 48

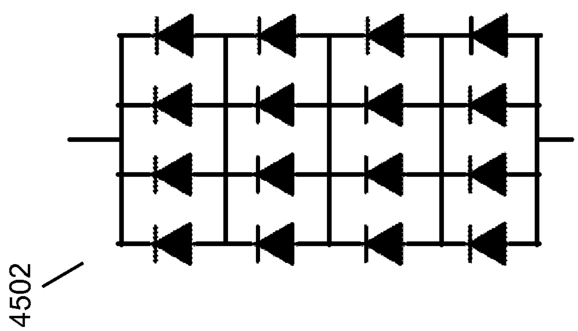


Fig. 49

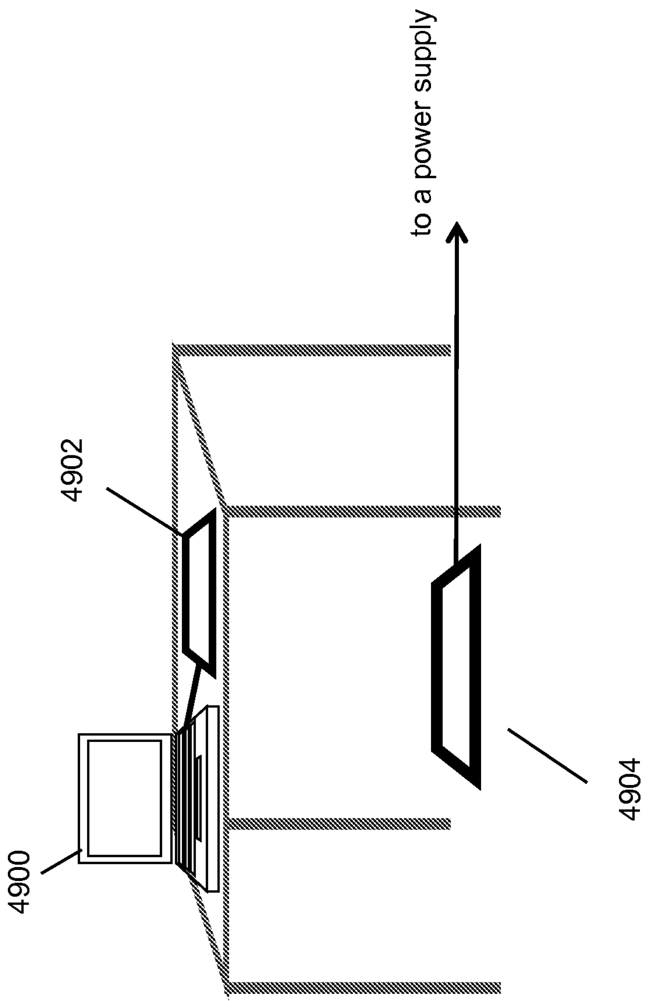
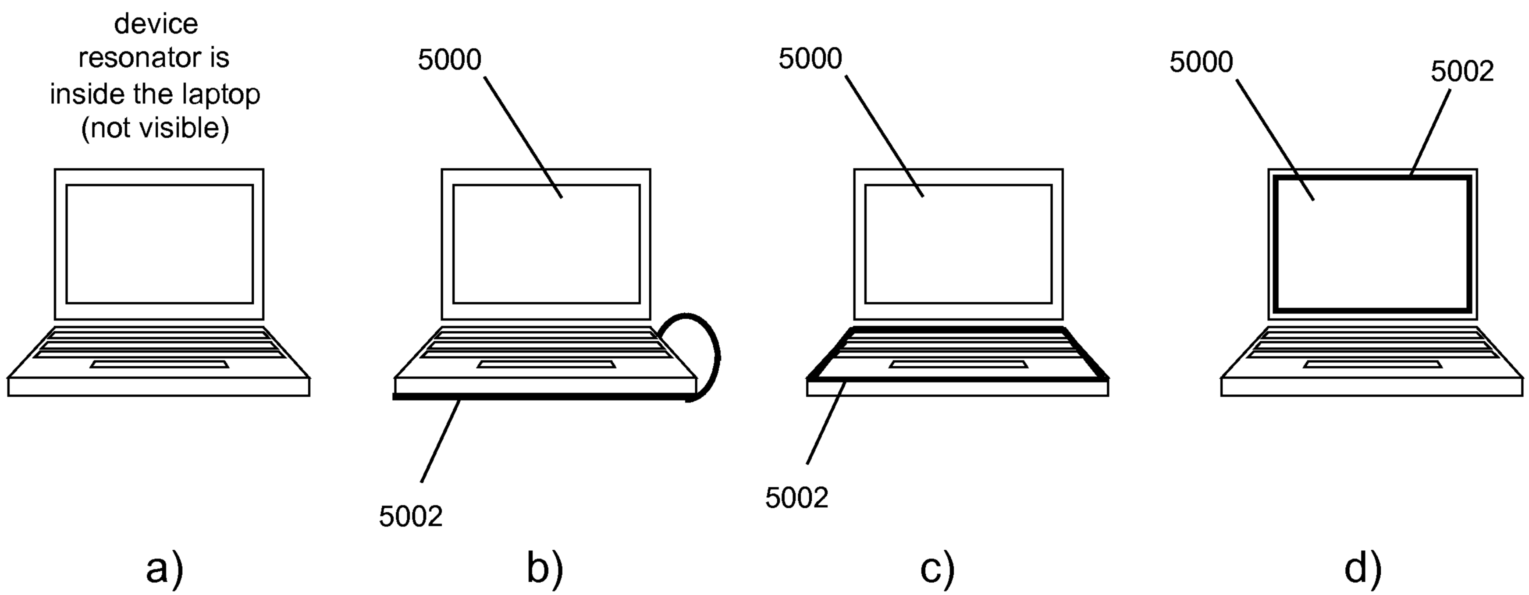


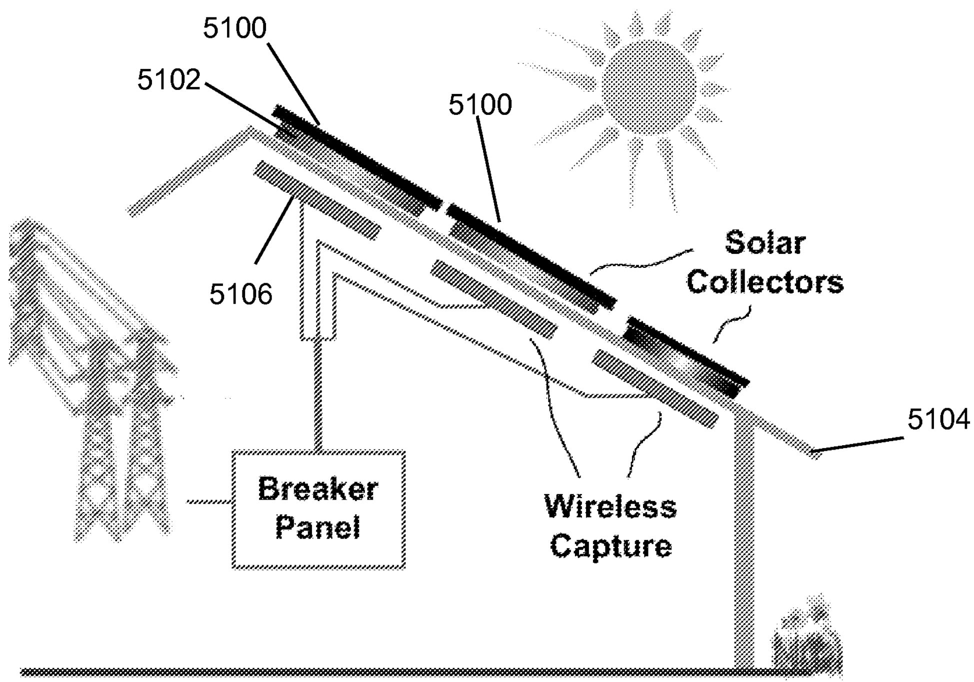
Fig. 50



device
resonator is
inside the laptop
(not visible)
5000
5002
5000
5002
5000
5002
a)
b)
c)
d)

52

Fig. 51


5100
5102
5100
5106
Solar Collectors
5104
Breaker Panel
Wireless Capture

WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS

CROSS-REFERENCE TO RELATED APPLICATIONS

[0001] This application is a continuation-in-part of the following U.S. patent application, U.S. 12/567716 filed September 25, 2009 which claims the benefit of the following U.S. provisional applications, U.S. App. No. 61/100,721 filed September 27,2008; U.S. App. No. 61/108,743 filed October 27, 2008; U.S. App. No. 61/147,386 filed January 26, 2009; U.S. App. No. 61/152,086 filed February 12, 2009; U.S. App. No. 61/178,508 filed May 15, 2009; U.S. App. No. 61/182,768 filed June 1, 2009; U.S. App. No. 61/121,159 filed December 9, 2008; U.S. App. No. 61/142,977 filed January 7, 2009; U.S. App. No. 61/142,885 filed January 6, 2009; U.S. App. No. 61/142,796 filed January 6, 2009; U.S. App. No. 61/142,889 filed January 6, 2009; U.S. App. No. 61/142,880 filed January 6, 2009; U.S. App. No. 61/142,818 filed January 6, 2009; U.S. App. No. 61/142,887 filed January 6, 2009; U.S. App. No. 61/156,764 filed March 2, 2009; U.S. App. No. 61/143,058 filed January 7, 2009; U.S. App. No. 61/152,390 filed February 13, 2009; U.S. App. No. 61/163,695 filed March 26, 2009; U.S. App. No. 61/172,633 filed April 24, 2009; U.S. App. No. 61/169,240 filed April 14, 2009, U.S. App. No. 61/173,747 filed April 29, 2009.

[0002] Each of the foregoing applications is incorporated herein by reference in its entirety.

BACKGROUND

[0003] Field:

[0004] This disclosure relates to wireless energy transfer, also referred to as wireless power transmission.

[0005] Description of the Related Art:

[0006] Energy or power may be transferred wirelessly using a variety of known radiative, or far-field, and non-radiative, or near-field, techniques. For example, radiative wireless information transfer using low-directionality antennas, such as those used in radio and cellular communications systems and home computer networks, may be considered wireless energy transfer. However, this type of radiative transfer is very inefficient because only a tiny

portion of the supplied or radiated power, namely, that portion in the direction of, and overlapping with, the receiver is picked up. The vast majority of the power is radiated away in all the other directions and lost in free space. Such inefficient power transfer may be acceptable for data transmission, but is not practical for transferring useful amounts of electrical energy for the purpose of doing work, such as for powering or charging electrical devices. One way to improve the transfer efficiency of some radiative energy transfer schemes is to use directional antennas to confine and preferentially direct the radiated energy towards a receiver. However, these directed radiation schemes may require an uninterruptible line-of-sight and potentially complicated tracking and steering mechanisms in the case of mobile transmitters and/or receivers. In addition, such schemes may pose hazards to objects or people that cross or intersect the beam when modest to high amounts of power are being transmitted. A known non-radiative, or near-field, wireless energy transfer scheme, often referred to as either induction or traditional induction, does not (intentionally) radiate power, but uses an oscillating current passing through a primary coil, to generate an oscillating magnetic near-field that induces currents in a near-by receiving or secondary coil. Traditional induction schemes have demonstrated the transmission of modest to large amounts of power, however only over very short distances, and with very small offset tolerances between the primary power supply unit and the secondary receiver unit. Electric transformers and proximity chargers are examples of devices that utilize this known short range, near-field energy transfer scheme.

[0007] Therefore a need exists for a wireless power transfer scheme that is capable of transferring useful amounts of electrical power over mid-range distances or alignment offsets. Such a wireless power transfer scheme should enable useful energy transfer over greater distances and alignment offsets than those realized with traditional induction schemes, but without the limitations and risks inherent in radiative transmission schemes.

SUMMARY

[0008] There is disclosed herein a non-radiative or near-field wireless energy transfer scheme that is capable of transmitting useful amounts of power over mid-range distances and alignment offsets. This inventive technique uses coupled electromagnetic resonators with long-lived oscillatory resonant modes to transfer power from a power supply to a power drain. The technique is general and may be applied to a wide range of resonators, even where the specific

examples disclosed herein relate to electromagnetic resonators. If the resonators are designed such that the energy stored by the electric field is primarily confined within the structure and that the energy stored by the magnetic field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant magnetic near-field. These types of resonators may be referred to as magnetic resonators. If the resonators are designed such that the energy stored by the magnetic field is primarily confined within the structure and that the energy stored by the electric field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant electric near-field. These types of resonators may be referred to as electric resonators. Either type of resonator may also be referred to as an electromagnetic resonator. Both types of resonators are disclosed herein.

[0009] The omni-directional but stationary (non-lossy) nature of the near-fields of the resonators we disclose enables efficient wireless energy transfer over mid-range distances, over a wide range of directions and resonator orientations, suitable for charging, powering, or simultaneously powering and charging a variety of electronic devices. As a result, a system may have a wide variety of possible applications where a first resonator, connected to a power source, is in one location, and a second resonator, potentially connected to electrical/electronic devices, batteries, powering or charging circuits, and the like, is at a second location, and where the distance from the first resonator to the second resonator is on the order of centimeters to meters. For example, a first resonator connected to the wired electricity grid could be placed on the ceiling of a room, while other resonators connected to devices, such as robots, vehicles, computers, communication devices, medical devices, and the like, move about within the room, and where these devices are constantly or intermittently receiving power wirelessly from the source resonator. From this one example, one can imagine many applications where the systems and methods disclosed herein could provide wireless power across mid-range distances, including consumer electronics, industrial applications, infrastructure power and lighting, transportation vehicles, electronic games, military applications, and the like.

[0010] Energy exchange between two electromagnetic resonators can be optimized when the resonators are tuned to substantially the same frequency and when the losses in the system are minimal. Wireless energy transfer systems may be designed so that the "coupling-time" between resonators is much shorter than the resonators' "loss-times". Therefore, the systems and methods described herein may utilize high quality factor (high-Q) resonators with

low intrinsic-loss rates. In addition, the systems and methods described herein may use sub-wavelength resonators with near-fields that extend significantly longer than the characteristic sizes of the resonators, so that the near-fields of the resonators that exchange energy overlap at mid-range distances. This is a regime of operation that has not been practiced before and that differs significantly from traditional induction designs.

[0011] It is important to appreciate the difference between the high-Q magnetic resonator scheme disclosed here and the known close-range or proximity inductive schemes, namely, that those known schemes do not conventionally utilize high-Q resonators. Using coupled-mode theory (CMT), (see, for example, *Waves and Fields in Optoelectronics*, H.A. Haus, Prentice Hall, 1984), one may show that a high-Q resonator-coupling mechanism can enable orders of magnitude more efficient power delivery between resonators spaced by mid-range distances than is enabled by traditional inductive schemes. Coupled high-Q resonators have demonstrated efficient energy transfer over mid-range distances and improved efficiencies and offset tolerances in short range energy transfer applications.

[0012] The systems and methods described herein may provide for near-field wireless energy transfer via strongly coupled high-Q resonators, a technique with the potential to transfer power levels from picowatts to kilowatts, safely, and over distances much larger than have been achieved using traditional induction techniques. Efficient energy transfer may be realized for a variety of general systems of strongly coupled resonators, such as systems of strongly coupled acoustic resonators, nuclear resonators, mechanical resonators, and the like, as originally described by researchers at M.I.T. in their publications, "Efficient wireless non-radiative mid-range energy transfer", *Annals of Physics*, vol. 323, Issue 1, p. 34 (2008) and "Wireless Power Transfer via Strongly Coupled Magnetic Resonances", *Science*, vol. 317, no. 5834, p. 83, (2007). Disclosed herein are electromagnetic resonators and systems of coupled electromagnetic resonators, also referred to more specifically as coupled magnetic resonators and coupled electric resonators, with operating frequencies below 10 GHz.

[0013] This disclosure describes wireless energy transfer technologies, also referred to as wireless power transmission technologies. Throughout this disclosure, we may use the terms wireless energy transfer, wireless power transfer, wireless power transmission, and the like, interchangeably. We may refer to supplying energy or power from a source, an AC or DC source, a battery, a source resonator, a power supply, a generator, a solar panel, and thermal

collector, and the like, to a device, a remote device, to multiple remote devices, to a device resonator or resonators, and the like. We may describe intermediate resonators that extend the range of the wireless energy transfer system by allowing energy to hop, transfer through, be temporarily stored, be partially dissipated, or for the transfer to be mediated in any way, from a source resonator to any combination of other device and intermediate resonators, so that energy transfer networks, or strings, or extended paths may be realized. Device resonators may receive energy from a source resonator, convert a portion of that energy to electric power for powering or charging a device, and simultaneously pass a portion of the received energy onto other device or mobile device resonators. Energy may be transferred from a source resonator to multiple device resonators, significantly extending the distance over which energy may be wirelessly transferred. The wireless power transmission systems may be implemented using a variety of system architectures and resonator designs. The systems may include a single source or multiple sources transmitting power to a single device or multiple devices. The resonators may be designed to be source or device resonators, or they may be designed to be repeaters. In some cases, a resonator may be a device and source resonator simultaneously, or it may be switched from operating as a source to operating as a device or a repeater. One skilled in the art will understand that a variety of system architectures may be supported by the wide range of resonator designs and functionalities described in this application.

[0014] In the wireless energy transfer systems we describe, remote devices may be powered directly, using the wirelessly supplied power or energy, or the devices may be coupled to an energy storage unit such as a battery, a super-capacitor, an ultra-capacitor, or the like (or other kind of power drain), where the energy storage unit may be charged or re-charged wirelessly, and/or where the wireless power transfer mechanism is simply supplementary to the main power source of the device. The devices may be powered by hybrid battery/energy storage devices such as batteries with integrated storage capacitors and the like. Furthermore, novel battery and energy storage devices may be designed to take advantage of the operational improvements enabled by wireless power transmission systems.

[0015] Other power management scenarios include using wirelessly supplied power to recharge batteries or charge energy storage units while the devices they power are turned off, in an idle state, in a sleep mode, and the like. Batteries or energy storage units may be charged or recharged at high (fast) or low (slow) rates. Batteries or energy storage units may be trickle

charged or float charged. Multiple devices may be charged or powered simultaneously in parallel or power delivery to multiple devices may be serialized such that one or more devices receive power for a period of time after which other power delivery is switched to other devices. Multiple devices may share power from one or more sources with one or more other devices either simultaneously, or in a time multiplexed manner, or in a frequency multiplexed manner, or in a spatially multiplexed manner, or in an orientation multiplexed manner, or in any combination of time and frequency and spatial and orientation multiplexing. Multiple devices may share power with each other, with at least one device being reconfigured continuously, intermittently, periodically, occasionally, or temporarily, to operate as wireless power sources. It would be understood by one of ordinary skill in the art that there are a variety of ways to power and/or charge devices, and the variety of ways could be applied to the technologies and applications described herein.

[0016] Wireless energy transfer has a variety of possible applications including for example, placing a source (e.g. one connected to the wired electricity grid) on the ceiling, under the floor, or in the walls of a room, while devices such as robots, vehicles, computers, PDAs or similar are placed or move freely within the room. Other applications may include powering or recharging electric-engine vehicles, such as buses and/or hybrid cars and medical devices, such as wearable or implantable devices. Additional example applications include the ability to power or recharge autonomous electronics (e.g. laptops, cell-phones, portable music players, house-hold robots, GPS navigation systems, displays, etc), sensors, industrial and manufacturing equipment, medical devices and monitors, home appliances and tools (e.g. lights, fans, drills, saws, heaters, displays, televisions, counter-top appliances, etc.), military devices, heated or illuminated clothing, communications and navigation equipment, including equipment built into vehicles, clothing and protective-wear such as helmets, body armor and vests, and the like, and the ability to transmit power to physically isolated devices such as to implanted medical devices, to hidden, buried, implanted or embedded sensors or tags, to and/or from roof-top solar panels to indoor distribution panels, and the like.

[0017] In one aspect, disclosed herein is a system including a source resonator having a Q-factor Q_1 and a characteristic size $_{x1}$, coupled to a power generator with direct electrical connections; and a second resonator having a Q-factor Q_2 and a characteristic size x_2, coupled to a load with direct electrical connections, and located a distance D from the source resonator,

wherein the source resonator and the second resonator are coupled to exchange energy wirelessly among the source resonator and the second resonator in order to transmit power from the power generator to the load, and wherein $\sqrt{Q_1 Q_2}$ is greater than 100.

[0018] Q_1 may be greater than 100 and Q_2 may be less than 100. Q_1 may be greater than 100 and Q_2 may be greater than 100. A useful energy exchange may be maintained over an operating distance from 0 to D, where D is larger than the smaller of x_1 and x_2. At least one of the source resonator and the second resonator may be a coil of at least one turn of a conducting material connected to a first network of capacitors. The first network of capacitors may include at least one tunable capacitor. The direct electrical connections of at least one of the source resonator to the ground terminal of the power generator and the second resonator to the ground terminal of the load may be made at a point on an axis of electrical symmetry of the first network of capacitors. The first network of capacitors may include at least one tunable butterfly-type capacitor, wherein the direct electrical connection to the ground terminal is made on a center terminal of the at least one tunable butterfly-type capacitor. The direct electrical connection of at least one of the source resonator to the power generator and the second resonator to the load may be made via a second network of capacitors, wherein the first network of capacitors and the second network of capacitors form an impedance matching network. The impedance matching network may be designed to match the coil to a characteristic impedance of the power generator or the load at a driving frequency of the power generator.

[0019] At least one of the first network of capacitors and the second network of capacitors may include at least one tunable capacitor. The first network of capacitors and the second network of capacitors may be adjustable to change an impedance of the impedance matching network at a driving frequency of the power generator. The first network of capacitors and the second network of capacitors may be adjustable to match the coil to the characteristic impedance of the power generator or the load at a driving frequency of the power generator. At least one of the first network of capacitors and the second network of capacitors may include at least one fixed capacitor that reduces a voltage across the at least one tunable capacitor. The direct electrical connections of at least one of the source resonator to the power generator and the second resonator to the load may be configured to substantially preserve a resonant mode. At least one of the source resonator and the second resonator may be a tunable resonator. The source resonator may be physically separated from the power generator and the second resonator

may be physically separated from the load. The second resonator may be coupled to a power conversion circuit to deliver DC power to the load. The second resonator may be coupled to a power conversion circuit to deliver AC power to the load. The second resonator may be coupled to a power conversion circuit to deliver both AC and DC power to the load. The second resonator may be coupled to a power conversion circuit to deliver power to a plurality of loads.

[0020] In another aspect, a system disclosed herein includes a source resonator having a Q-factor Q_1 and a characteristic size x_1, and a second resonator having a Q-factor Q_2 and a characteristic size x_2, and located a distance D from the source resonator; wherein the source resonator and the second resonator are coupled to exchange energy wirelessly among the source resonator and the second resonator; and wherein $\sqrt{Q_1 Q_2}$ is greater than 100, and wherein at least one of the resonators is enclosed in a low loss tangent material.

[0021] In another aspect, a system disclosed herein includes a source resonator having a Q-factor Q_1 and a characteristic size x_1, and a second resonator having a Q-factor Q_2 and a characteristic size x_2, and located a distance D from the source resonator; wherein the source resonator and the second resonator are coupled to exchange energy wirelessly among the source resonator and the second resonator, and wherein $\sqrt{Q_1 Q_2}$ is greater than 100; and wherein at least one of the resonators includes a coil of a plurality of turns of a conducting material connected to a network of capacitors, wherein the plurality of turns are in a common plane, and wherein a characteristic thickness of the at least one of the resonators is much less than a characteristic size of the at least one of the resonators.

[0022] In embodiments, the present invention may provide for a method and system comprising a source resonator optionally coupled to an energy source and a second resonator located a distance from the source resonator, where the source resonator and the second resonator are coupled to provide near-field wireless energy transfer among the source resonator and the second resonator and where the field of at least one of the source resonator and the second resonator is shaped to avoid a loss-inducing object. In embodiments, at least one of the source resonator and the second resonator may have a quality factor, Q>100. The source resonator Q may be greater than 100 and the second resonator Q may be greater than 100. The square root of the source resonator Q times the second resonator Q may be greater than 100. In embodiments, there may be more than one source resonator, more than one second resonator, more than three resonators, and the like.

[0023] Throughout this disclosure we may refer to the certain circuit components such as capacitors, inductors, resistors, diodes, switches and the like as circuit components or elements. We may also refer to series and parallel combinations of these components as elements, networks, topologies, circuits, and the like. We may describe combinations of capacitors, diodes, varactors, transistors, and/or switches as adjustable impedance networks, tuning networks, matching networks, adjusting elements, and the like. We may also refer to "self-resonant" objects that have both capacitance, and inductance distributed (or partially distributed, as opposed to solely lumped) throughout the entire object. It would be understood by one of ordinary skill in the art that adjusting and controlling variable components within a circuit or network may adjust the performance of that circuit or network and that those adjustments may be described generally as tuning, adjusting, matching, correcting, and the like. Other methods to tune or adjust the operating point of the wireless power transfer system may be used alone, or in addition to adjusting tunable components such as inductors and capacitors, or banks of inductors and capacitors.

[0024] Unless otherwise defined, all technical and scientific terms used herein have the same meaning as commonly understood by one of ordinary skill in the art to which this disclosure belongs. In case of conflict with publications, patent applications, patents, and other references mentioned or incorporated herein by reference, the present specification, including definitions, will control.

[0025] Any of the features described above may be used, alone or in combination, without departing from the scope of this disclosure. Other features, objects, and advantages of the systems and methods disclosed herein will be apparent from the following detailed description and figures.

BRIEF DESCRIPTION OF FIGURES

[0026] Fig. 1 (a) and (b) depict exemplary wireless power systems containing a source resonator 1 and device resonator 2 separated by a distance D.

[0027] Fig. 2 shows an exemplary resonator labeled according to the labeling convention described in this disclosure. Note that there are no extraneous objects or additional resonators shown in the vicinity of resonator 1.

[0028] Fig. 3 shows an exemplary resonator in the presence of a "loading" object, labeled according to the labeling convention described in this disclosure.

[0029] Fig. 4 shows an exemplary resonator in the presence of a "perturbing" object, labeled according to the labeling convention described in this disclosure.

[0030] Fig. 5 shows a plot of efficiency, η, vs. strong coupling factor, $U = \kappa/\sqrt{\Gamma_s \Gamma_d} = k\sqrt{Q_s Q_d}$.

[0031] Fig. 6 (a) shows a circuit diagram of one example of a resonator (b) shows a diagram of one example of a capacitively-loaded inductor loop magnetic resonator, (c) shows a drawing of a self-resonant coil with distributed capacitance and inductance, (d) shows a simplified drawing of the electric and magnetic field lines associated with an exemplary magnetic resonator of the current disclosure, and (e) shows a diagram of one example of an electric resonator.

[0032] Fig. 7 shows a plot of the "quality factor", Q (solid line), as a function of frequency, of an exemplary resonator that may be used for wireless power transmission at MHz frequencies. The absorptive Q (dashed line) increases with frequency, while the radiative Q (dotted line) decreases with frequency, thus leading the overall Q to peak at a particular frequency.

[0033] Fig. 8 shows a drawing of a resonator structure with its characteristic size, thickness and width indicated.

[0034] Fig. 9 (a) and (b) show drawings of exemplary inductive loop elements.

[0035] Fig. 10 (a) and (b) show two examples of trace structures formed on printed circuit boards and used to realize the inductive element in magnetic resonator structures.

[0036] Fig. 11 (a) shows a perspective view diagram of a planar magnetic resonator, (b) shows a perspective view diagram of a two planar magnetic resonator with various geometries, and c) shows is a perspective view diagram of a two planar magnetic resonators separated by a distance D.

[0037] Fig. 12 is a perspective view of an example of a planar magnetic resonator.

[0038] Fig. 13 is a perspective view of a planar magnetic resonator arrangement with a circular resonator coil.

[0039] Fig. 14 is a perspective view of an active area of a planar magnetic resonator.

[0040] Fig. 15 is a perspective view of an application of the wireless power transfer system with a source at the center of a table powering several devices placed around the source.

[0041] Fig. 16(a) shows a 3D finite element model of a copper and magnetic material structure driven by a square loop of current around the choke point at its center. In this example, a structure may be composed of two boxes made of a conducting material such as copper, covered by a layer of magnetic material, and connected by a block of magnetic material. The inside of the two conducting boxes in this example would be shielded from AC electromagnetic fields generated outside the boxes and may house lossy objects that might lower the Q of the resonator or sensitive components that might be adversely affected by the AC electromagnetic fields. Also shown are the calculated magnetic field streamlines generated by this structure, indicating that the magnetic field lines tend to follow the lower reluctance path in the magnetic material. Fig. 16(b) shows interaction, as indicated by the calculated magnetic field streamlines, between two identical structures as shown in (a). Because of symmetry, and to reduce computational complexity, only one half of the system is modeled (but the computation assumes the symmetrical arrangement of the other half).

[0042] Fig. 17 shows an equivalent circuit representation of a magnetic resonator including a conducting wire wrapped N times around a structure, possibly containing magnetically permeable material. The inductance is realized using conducting loops wrapped around a structure comprising a magnetic material and the resistors represent loss mechanisms in the system (R_{wire} for resistive losses in the loop, R_{μ} denoting the equivalent series resistance of the structure surrounded by the loop). Losses may be minimized to realize high-Q resonators.

[0043] Fig. 18 shows a Finite Element Method (FEM) simulation of two high conductivity surfaces above and below a disk composed of lossy dielectric material, in an external magnetic field of frequency 6.78 MHz. Note that the magnetic field was uniform before the disk and conducting materials were introduced to the simulated environment. This simulation is performed in cylindrical coordinates. The image is azimuthally symmetric around the $r = 0$ axis. The lossy dielectric disk has $\epsilon_r = 1$ and $\sigma = 10$ S/m.

[0044] Fig. 19 shows a drawing of a magnetic resonator with a lossy object in its vicinity completely covered by a high-conductivity surface.

[0045] Fig. 20 shows a drawing of a magnetic resonator with a lossy object in its vicinity partially covered by a high-conductivity surface.

[0046] Fig. 21 shows a drawing of a magnetic resonator with a lossy object in its vicinity placed on top of a high-conductivity surface.

[0047] Fig. 22 shows a diagram of a completely wireless projector.

[0048] Fig. 23 shows the magnitude of the electric and magnetic fields along a line that contains the diameter of the circular loop inductor and along the axis of the loop inductor.

[0049] Fig. 24 shows a drawing of a magnetic resonator and its enclosure along with a necessary but lossy object placed either (a) in the corner of the enclosure, as far away from the resonator structure as possible or (b) in the center of the surface enclosed by the inductive element in the magnetic resonator.

[0050] Fig. 25 shows a drawing of a magnetic resonator with a high-conductivity surface above it and a lossy object, which may be brought into the vicinity of the resonator, but above the high-conductivity sheet.

[0051] Fig. 26(a) shows an axially symmetric FEM simulation of a thin conducting (copper) cylinder or disk (20 cm in diameter, 2 cm in height) exposed to an initially uniform, externally applied magnetic field (gray flux lines) along the z-axis. The axis of symmetry is at r=0. The magnetic streamlines shown originate at $z = -\infty$, where they are spaced from r=3 cm to r=10 cm in intervals of 1 cm. The axes scales are in meters. Fig. 26 (b) shows the same structure and externally applied field as in (a), except that the conducting cylinder has been modified to include a 0.25 mm layer of magnetic material (not visible) with $\mu_r^{'}$=40, on its outside surface. Note that the magnetic streamlines are deflected away from the cylinder significantly less than in (a).

[0052] Fig. 27 shows an axi-symmetric view of a variation based on the system shown in Fig. 26. Only one surface of the lossy material is covered by a layered structure of copper and magnetic materials. The inductor loop is placed on the side of the copper and magnetic material structure opposite to the lossy material as shown.

[0053] Fig. 28 (a) depicts a general topology of a matching circuit including an indirect coupling to a high-Q inductive element.

[0054] Fig. 28 (b) shows a block diagram of a magnetic resonator that includes a conductor loop inductor and a tunable impedance network. Physical electrical connections to this resonator may be made to the terminal connections.

[0055] Fig. 28 (c) depicts a general topology of a matching circuit directly coupled to a high-Q inductive element.

[0056] Fig. 28 (d) depicts a general topology of a symmetric matching circuit directly coupled to a high-Q inductive element and driven anti-symmetrically (balanced drive).

[0057] Fig. 28 (e) depicts a general topology of a matching circuit directly coupled to a high-Q inductive element and connected to ground at a point of symmetry of the main resonator (unbalanced drive).

[0058] Figs. 29(a) and 29(b) depict two topologies of matching circuits transformer-coupled (i.e. indirectly or inductively) to a high-Q inductive element. The highlighted portion of the Smith chart in (c) depicts the complex impedances (arising from *L* and *R* of the inductive element) that may be matched to an arbitrary real impedance Z_0 by the topology of Fig. 31(b) in the case $\omega L_2 = 1/\omega C_2$.

[0059] Figs. 30(a),(b),(c),(d),(e),(f) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in series with Z_0. The topologies shown in Figs. 30(a),(b),(c) are driven with a common-mode signal at the input terminals, while the topologies shown in Figs 30(d),(e),(f) are symmetric and receive a balanced drive. The highlighted portion of the Smith chart in 30(g) depicts the complex impedances that may be matched by these topologies. Figs. 30(h),(i),(j),(k),(l),(m) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in series with Z_0.

[0060] Figs. 31(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in series with Z_0. They are connected to ground at the center point of a capacitor and receive an unbalanced drive. The highlighted portion of the Smith chart in Fig. 31(d) depicts the complex impedances that may be matched by these topologies. Figs. 31(e),(f),(g) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in series with Z_0.

[0061] Figs. 32(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in series with Z_0. They are connected to ground by tapping at the center point of the inductor loop and receive an unbalanced drive. The highlighted portion of the Smith chart in (d) depicts the complex impedances that may be matched by these topologies, (e),(f),(g) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in series with Z_0.

[0062] Figs. 33(a),(b),(c),(d),(e),(f) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in parallel with Z_o. The topologies shown in Figs. 33(a),(b),(c) are driven with a common-mode signal at the input terminals, while the topologies shown in Figs 33(d),(e),(f) are symmetric and receive a balanced drive. The highlighted portion of the Smith chart in Fig. 33(g) depicts the complex impedances that may be matched by these topologies. Figs. 33(h),(i),(j),(k),(l),(m) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in parallel with Z_o.

[0063] Figs. 34(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in parallel with Z_0. They are connected to ground at the center point of a capacitor and receive an unbalanced drive. The highlighted portion of the Smith chart in (d) depicts the complex impedances that may be matched by these topologies. Figs. 34(e),(f),(g) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in parallel with Z_0.

[0064] Figs. 35(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in parallel with Z_0. They are connected to ground by tapping at the center point of the inductor loop and receive an unbalanced drive. The highlighted portion of the Smith chart in Figs. 35(d),(e), and (f) depict the complex impedances that may be matched by these topologies.

[0065] Figs. 36(a),(b),(c),(d) depict four topologies of networks of fixed and variable capacitors designed to produce an overall variable capacitance with finer tuning resolution and some with reduced voltage on the variable capacitor.

[0066] Figs. 37(a) and 37(b) depict two topologies of networks of fixed capacitors and a variable inductor designed to produce an overall variable capacitance.

[0067] Fig. 38 depicts a high level block diagram of a wireless power transmission system.

[0068] Fig. 39 depicts a block diagram of an exemplary wirelessly powered device.

[0069] Fig. 40 depicts a block diagram of the source of an exemplary wireless power transfer system.

[0070] Fig. 41 shows an equivalent circuit diagram of a magnetic resonator. The slash through the capacitor symbol indicates that the represented capacitor may be fixed or variable.

The port parameter measurement circuitry may be configured to measure certain electrical signals and may measure the magnitude and phase of signals.

[0071] Fig. 42 shows a circuit diagram of a magnetic resonator where the tunable impedance network is realized with voltage controlled capacitors. Such an implementation may be adjusted, tuned or controlled by electrical circuits including programmable or controllable voltage sources and/or computer processors. The voltage controlled capacitors may be adjusted in response to data measured by the port parameter measurement circuitry and processed by measurement analysis and control algorithms and hardware. The voltage controlled capacitors may be a switched bank of capacitors.

[0072] Fig. 43 shows an end-to-end wireless power transmission system. In this example, both the source and the device contain port measurement circuitry and a processor. The box labeled "coupler/switch" indicates that the port measurement circuitry may be connected to the resonator by a directional coupler or a switch, enabling the measurement, adjustment and control of the source and device resonators to take place in conjunction with, or separate from, the power transfer functionality.

[0073] Fig. 44 shows an end-to-end wireless power transmission system. In this example, only the source contains port measurement circuitry and a processor. In this case, the device resonator operating characteristics may be fixed or may be adjusted by analog control circuitry and without the need for control signals generated by a processor.

[0074] Fig. 45 shows an end-to-end wireless power transmission system. In this example, both the source and the device contain port measurement circuitry but only the source contains a processor. Data from the device is transmitted through a wireless communication channel, which could be implemented either with a separate antenna, or through some modulation of the source drive signal.

[0075] Fig. 46 shows an end-to-end wireless power transmission system. In this example, only the source contains port measurement circuitry and a processor. Data from the device is transmitted through a wireless communication channel, which could be implemented either with a separate antenna, or through some modulation of the source drive signal.

[0076] Fig. 47 shows coupled magnetic resonators whose frequency and impedance may be automatically adjusted using algorithms implemented using a processor or a computer.

[0077] Fig. 48 shows a varactor array.

[0078] Fig. 49 shows a device (laptop computer) being wirelessly powered or charged by a source, where both the source and device resonator are physically separated from, but electrically connected to, the source and device.

[0079] Fig. 50 (a) is an illustration of a wirelessly powered or charged laptop application where the device resonator is inside the laptop case and is not visible.

[0080] Fig. 50 (b) is an illustration of a wirelessly powered or charged laptop application where the resonator is underneath the laptop base and is electrically connected to the laptop power input by an electrical cable.

[0081] Fig. 50 (c) is an illustration of a wirelessly powered or charged laptop application where the resonator is attached to the laptop base.

[0082] Fig. 50 (d) is an illustration of a wirelessly powered or charged laptop application where the resonator is attached to the laptop display.

[0083] Fig. 51 is a diagram of rooftop PV panels with wireless power transfer.

DETAILED DESCRIPTION

[0084] As described above, this disclosure relates to coupled electromagnetic resonators with long-lived oscillatory resonant modes that may wirelessly transfer power from a power supply to a power drain. However, the technique is not restricted to electromagnetic resonators, but is general and may be applied to a wide variety of resonators and resonant objects. Therefore, we first describe the general technique, and then disclose electromagnetic examples for wireless energy transfer.

[0085] **Resonators**

[0086] A resonator may be defined as a system that can store energy in at least two different forms, and where the stored energy is oscillating between the two forms. The resonance has a specific oscillation mode with a resonant (modal) frequency, *f*, and a resonant (modal) field. The angular resonant frequency, *ω,* may be defined as $\omega = 2\pi f$, the resonant wavelength, λ, may be defined as $\lambda = c/f$, where *c* is the speed of light, and the resonant period, *T*, may be defined as $T = 1/f = 2\pi/\omega$. In the absence of loss mechanisms, coupling mechanisms or external energy supplying or draining mechanisms, the total resonator stored energy, *W*, would stay fixed

and the two forms of energy would oscillate, wherein one would be maximum when the other is minimum and vice versa.

[0087] In the absence of extraneous materials or objects, the energy in the resonator 102 shown in Fig. 1 may decay or be lost by intrinsic losses. The resonator fields then obey the following linear equation:

$$\frac{da(t)}{dt} = -i(\omega - i\Gamma)a(t),$$

where the variable *a(t)* is the resonant field amplitude, defined so that the energy contained within the resonator is given by $|a(t)|^2$. Γ is the intrinsic energy decay or loss rate (e.g. due to absorption and radiation losses).

[0088] The Quality Factor, or *Q*-factor, or *Q*, of the resonator, which characterizes the energy decay, is inversely proportional to these energy losses. It may be defined as $Q=\omega*W/P$, where *P* is the time-averaged power lost at steady state. That is, a resonator 102 with a high-*Q* has relatively low intrinsic losses and can store energy for a relatively long time. Since the resonator loses energy at its intrinsic decay rate, 2Γ, its *Q*, also referred to as its intrinsic *Q*, is given by $Q=\omega/2\Gamma$. The quality factor also represents the number of oscillation periods, *T*, it takes for the energy in the resonator to decay by a factor of *e*.

[0089] As described above, we define the quality factor or *Q* of the resonator as that due only to intrinsic loss mechanisms. A subscript index such as Q_1, indicates the resonator (resonator 1 in this case) to which the *Q* refers. Fig. 2 shows an electromagnetic resonator 102 labeled according to this convention. Note that in this figure, there are no extraneous objects or additional resonators in the vicinity of resonator 1.

[0090] Extraneous objects and/or additional resonators in the vicinity of a first resonator may perturb or load the first resonator, thereby perturbing or loading the *Q* of the first resonator, depending on a variety of factors such as the distance between the resonator and object or other resonator, the material composition of the object or other resonator, the structure of the first resonator, the power in the first resonator, and the like. Unintended external energy losses or coupling mechanisms to extraneous materials and objects in the vicinity of the resonators may be referred to as "perturbing" the *Q* of a resonator, and may be indicated by a subscript within rounded parentheses, (). Intended external energy losses, associated with energy transfer via coupling to other resonators and to generators and loads in the wireless energy transfer system

may be referred to as "loading" the Q of the resonator, and may be indicated by a subscript within square brackets, [].

[0091] The Q of a resonator 102 connected or coupled to a power generator, g, or load 302 , l, may be called the "loaded quality factor" or the "loaded Q" and may be denoted by $Q_{[g]}$ or $Q_{[l]}$, as illustrated in Fig. 3. In general, there may be more than one generator or load 302 connected to a resonator 102. However, we do not list those generators or loads separately but rather use "g" and "l" to refer to the equivalent circuit loading imposed by the combinations of generators and loads. In general descriptions, we may use the subscript "l" to refer to either generators or loads connected to the resonators.

[0092] In some of the discussion herein, we define the "loading quality factor" or the "loading Q" due to a power generator or load connected to the resonator, as $\delta Q_{[l]}$, where, $1/\delta Q_{[l]} \equiv 1/Q_{[l]} - 1/Q$. Note that the larger the loading Q, $\delta Q_{[l]}$, of a generator or load, the less the loaded Q, $Q_{[l]}$, deviates from the unloaded Q of the resonator.

[0093] The Q of a resonator in the presence of an extraneous object 402, p, that is not intended to be part of the energy transfer system may be called the "perturbed quality factor" or the "perturbed Q" and may be denoted by $Q_{(p)}$, as illustrated in Fig. 4. In general, there may be many extraneous objects, denoted as $p1$, $p2$, etc., or a set of extraneous objects $\{p\}$, that perturb the Q of the resonator 102. In this case, the perturbed Q may be denoted $Q_{(p1+p2+...)}$ or $Q_{(\{p\})}$. For example, $Q_{1(brick+wood)}$ may denote the perturbed quality factor of a first resonator in a system for wireless power exchange in the presence of a brick and a piece of wood, and $Q_{2(\{office\})}$ may denote the perturbed quality factor of a second resonator in a system for wireless power exchange in an office environment.

[0094] In some of the discussion herein, we define the "perturbing quality factor" or the "perturbing Q" due to an extraneous object, p, as $\delta Q_{(p)}$, where $1/\delta Q_{(p)} \equiv 1/Q_{(p)} - 1/Q$. As stated before, the perturbing quality factor may be due to multiple extraneous objects, $p1$, $p2$, etc. or a set of extraneous objects, $\{p\}$. The larger the perturbing Q, $\delta Q_{(p)}$, of an object, the less the perturbed Q, $Q_{(p)}$, deviates from the unperturbed Q of the resonator.

[0095] In some of the discussion herein, we also define $\Theta_{(p)} \equiv Q_{(p)} / Q$ and call it the "quality factor insensitivity" or the "Q-insensitivity" of the resonator in the presence of an

extraneous object. A subscript index, such as $\Theta_{1(p)}$, indicates the resonator to which the perturbed and unperturbed quality factors are referring, namely, $\Theta_{1(p)} \equiv Q_{1(p)} / Q_1$.

[0096] Note that the quality factor, *Q,* may also be characterized as "unperturbed", when necessary to distinguish it from the perturbed quality factor, $Q_{(p)}$, and "unloaded", when necessary to distinguish it from the loaded quality factor, $Q_{[l]}$. Similarly, the perturbed quality factor, $Q_{(p)}$, may also be characterized as "unloaded", when necessary to distinguish them from the loaded perturbed quality factor, $Q_{(p)[l]}$.

[0097] Coupled Resonators

[0098] Resonators having substantially the same resonant frequency, coupled through any portion of their near-fields may interact and exchange energy. There are a variety of physical pictures and models that may be employed to understand, design, optimize and characterize this energy exchange. One way to describe and model the energy exchange between two coupled resonators is using coupled mode theory (CMT).

[0099] In coupled mode theory, the resonator fields obey the following set of linear equations:

$$\frac{da_m(t)}{dt} = -i\left(\omega_m - i\Gamma_m\right)a_m(t) + i\sum_{n \neq m} \kappa_{mn} a_n(t)$$

where the indices denote different resonators and and κ_{mn} are the coupling coefficients between the resonators. For a reciprocal system, the coupling coefficients may obey the relation $\kappa_{mn} = \kappa_{nm}$. Note that, for the purposes of the present specification, far-field radiation interference effects will be ignored and thus the coupling coefficients will be considered real. Furthermore, since in all subsequent calculations of system performance in this specification the coupling coefficients appear only with their square, κ_{mn}^2, we use κ_{mn} to denote the absolute value of the real coupling coefficients.

[00100] Note that the coupling coefficient, κ_{mn}, from the CMT described above is related to the so-called coupling factor, k_{mn}, between resonators *m* and *n* by $k_{mn} = 2\kappa_{mn} \big/ \sqrt{\omega_m \omega_n}$. We define a "strong-coupling factor", U_{mn}, as the ratio of the coupling and loss rates between resonators *m* and *n*, by $U_{mn} = \kappa_{mn} \big/ \sqrt{\Gamma_m \Gamma_n} = k_{mn}\sqrt{Q_m Q_n}$.

[00101] The quality factor of a resonator *m,* in the presence of a similar frequency resonator *n* or additional resonators, may be loaded by that resonator *n* or additional resonators, in a fashion similar to the resonator being loaded by a connected power generating or consuming device. The fact that resonator *m* may be loaded by resonator *n* and vice versa is simply a different way to see that the resonators are coupled.

[00102] The loaded *Q*'s of the resonators in these cases may be denoted as $Q_{m[n]}$ and $Q_{n[m]}$. For multiple resonators or loading supplies or devices, the total loading of a resonator may be determined by modeling each load as a resistive loss, and adding the multiple loads in the appropriate parallel and/or series combination to determine the equivalent load of the ensemble.

[00103] In some of the discussion herein, we define the "loading quality factor" or the "loading Q_m" of resonator *m* due to resonator *n* as $\delta Q_{m[n]}$, where $1/\delta Q_{m[n]} \equiv 1/Q_{m[n]} - 1/Q_m$. Note that resonator *n* is also loaded by resonator *m* and its "loading Q_n" is given by $1/\delta Q_{n[m]} \equiv 1/Q_{n[m]} - 1/Q_n$.

[00104] When one or more of the resonators are connected to power generators or loads, the set of linear equations is modified to:

$$\frac{da_m(t)}{dt} = -i(\omega_m - i\Gamma_m)a_m(t) + i\sum_{n\neq m}\kappa_{mn}a_n(t) - \kappa_m a_m(t) + \sqrt{2\kappa_m}s_{+m}(t)$$
$$s_{-m}(t) = \sqrt{2\kappa_m}a_m(t) - s_{+m}(t),$$

where $s_{+m}(t)$ and $s_{-m}(t)$ are respectively the amplitudes of the fields coming from a generator into the resonator *m* and going out of the resonator *m* either back towards the generator or into a load, defined so that the power they carry is given by $|s_{+m}(t)|^2$ and $|s_{-m}(t)|^2$. The loading coefficients κ_m relate to the rate at which energy is exchanged between the resonator *m* and the generator or load connected to it.

[00105] Note that the loading coefficient, κ_m, from the CMT described above is related to the loading quality factor, $\delta Q_{m[l]}$, defined earlier, by $\delta Q_{m[l]} = \omega_m/2\kappa_m$.

[00106] We define a "strong-loading factor", $U_{m[l]}$, as the ratio of the loading and loss rates of resonator *m*, $U_{m[l]} = \kappa_m/\Gamma_m = Q_m/\delta Q_{m[l]}$.

[00107] Fig. 1(a) shows an example of two coupled resonators 1000, a first resonator 102S, configured as a source resonator and a second resonator 102D, configured as a device resonator. Energy may be transferred over a distance D between the resonators. The source resonator 102S may be driven by a power supply or generator (not shown). Work may be extracted from the device resonator 102D by a power consuming drain or load (e.g. a load resistor, not shown). Let us use the subscripts "s" for the source, "d" for the device, "g" for the generator, and "l" for the load, and, since in this example there are only two resonators and $\kappa_{sd}=\kappa_{ds}$, let us drop the indices on κ_{sd}, k_{sd}, and U_{sd}, and denote them as κ, k, and U, respectively.

[00108] The power generator may be constantly driving the source resonator at a constant driving frequency, f, corresponding to an angular driving frequency, ω, where $\omega = 2\pi f$.

[00109] In this case, the efficiency, $\eta = |s_{-d}|^2 / |s_{+s}|^2$, of the power transmission from the generator to the load (via the source and device resonators) is maximized under the following conditions: The source resonant frequency, the device resonant frequency and the generator driving frequency have to be matched, namely

$$\omega_s = \omega_d = \omega.$$

Furthermore, the loading Q of the source resonator due to the generator, $\delta Q_{s[g]}$, has to be matched (equal) to the loaded Q of the source resonator due to the device resonator and the load, $Q_{s[dl]}$, and inversely the loading Q of the device resonator due to the load, $\delta Q_{d[l]}$, has to be matched (equal) to the loaded Q of the device resonator due to the source resonator and the generator, $Q_{d[sg]}$, namely

$$\delta Q_{s[g]} = Q_{s[dl]} \text{ and } \delta Q_{d[l]} = Q_{d[sg]}.$$

These equations determine the optimal loading rates of the source resonator by the generator and of the device resonator by the load as

$$U_{d[l]} = \kappa_d / \Gamma_d = Q_d / \delta Q_{d[l]} = \sqrt{1+U^2} = \sqrt{1+\left(\kappa / \sqrt{\Gamma_s \Gamma_d}\right)^2} = Q_s / \delta Q_{s[g]} = \kappa_s / \Gamma_s = U_{s[g]}.$$

Note that the above frequency matching and Q matching conditions are together known as "impedance matching" in electrical engineering.

[00110] Under the above conditions, the maximized efficiency is a monotonically increasing function of only the strong-coupling factor, $U = \kappa / \sqrt{\Gamma_s \Gamma_d} = k\sqrt{Q_s Q_d}$, between the source and device resonators and is given by, $\eta = U^2 / \left(1+\sqrt{1+U^2}\right)^2$, as shown in Fig. 5. Note that the coupling efficiency, η, is greater than 1% when U is greater than 0.2, is greater than 10% when

U is greater than 0.7, is greater than 17% when U is greater than 1, is greater than 52% when U is greater than 3, is greater than 80% when U is greater than 9, is greater than 90% when U is greater than 19, and is greater than 95% when U is greater than 45. In some applications, the regime of operation where $U>1$ may be referred to as the "strong-coupling" regime.

[00111] Since a large $U=\kappa/\sqrt{\Gamma_s\Gamma_d}=\left(2\kappa/\sqrt{\omega_s\omega_d}\right)\sqrt{Q_sQ_d}$ is desired in certain circumstances, resonators may be used that are high-Q. The Q of each resonator may be high. The geometric mean of the resonator Q's, $\sqrt{Q_sQ_d}$ may also or instead be high.

[00112] The coupling factor, k, is a number between $0\le k\le 1$, and it may be independent (or nearly independent) of the resonant frequencies of the source and device resonators, rather it may determined mostly by their relative geometry and the physical decay-law of the field mediating their coupling. In contrast, the coupling coefficient, $\kappa=k\sqrt{\omega_s\omega_d}/2$, may be a strong function of the resonant frequencies. The resonant frequencies of the resonators may be chosen preferably to achieve a high Q rather than to achieve a low Γ, as these two goals may be achievable at two separate resonant frequency regimes.

[00113] A high-Q resonator may be defined as one with $Q>100$. Two coupled resonators may be referred to as a system of high-Q resonators when each resonator has a Q greater than 100, $Q_s>100$ and $Q_d>100$. In other implementationss, two coupled resonators may be referred to as a system of high-Q resonators when the geometric mean of the resonator Q's is greater than 100, $\sqrt{Q_sQ_d}>100$.

[00114] The resonators may be named or numbered. They may be referred to as source resonators, device resonators, first resonators, second resonators, repeater resonators, and the like. It is to be understood that while two resonators are shown in Fig. 1, and in many of the examples below, other implementations may include three (3) or more resonators. For example, a single source resonator 102S may transfer energy to multiple device resonators 102D or multiple devices. Energy may be transferred from a first device to a second, and then from the second device to the third, and so forth. Multiple sources may transfer energy to a single device or to multiple devices connected to a single device resonator or to multiple devices connected to multiple device resonators. Resonators 102 may serve alternately or simultaneously as sources, devices, or they may be used to relay power from a source in one location to a device in another location. Intermediate electromagnetic resonators 102 may be used to extend the distance range of wireless energy transfer systems. Multiple resonators 102 may be daisy chained together, exchanging energy over extended distances and with a wide range of sources and devices. High power levels may be split between multiple sources 102S, transferred to multiple devices and recombined at a distant location.

[00115] The analysis of a single source and a single device resonator may be extended to multiple source resonators and/or multiple device resonators and/or multiple intermediate resonators. In such an analysis, the conclusion may be that large strong-coupling factors, U_{mn}, between at least some or all of the multiple resonators is preferred for a high system efficiency in the wireless energy transfer. Again, implementations may use source, device and intermediate resonators that have a high *Q*. The *Q* of each resonator may be high. The geometric mean $\sqrt{Q_m Q_n}$ of the *Q*'s for pairs of resonators *m* and *n*, for which a large U_{mn} is desired, may also or instead be high.

[00116] Note that since the strong-coupling factor of two resonators may be determined by the relative magnitudes of the loss mechanisms of each resonator and the coupling mechanism between the two resonators, the strength of any or all of these mechanisms may be perturbed in the presence of extraneous objects in the vicinity of the resonators as described above.

[00117] Continuing the conventions for labeling from the previous sections, we describe *k* as the coupling factor in the absence of extraneous objects or materials. We denote the coupling factor in the presence of an extraneous object, *p*, as $k_{(p)}$, and call it the "perturbed coupling factor" or the "perturbed *k*". Note that the coupling factor, *k*, may also be characterized as "unperturbed", when necessary to distinguish from the perturbed coupling factor $k_{(p)}$.

[00118] We define $\delta k_{(p)} \equiv k_{(p)} - k$ and we call it the "perturbation on the coupling factor" or the "perturbation on *k*" due to an extraneous object, *p*.

[00119] We also define $\beta_{(p)} \equiv k_{(p)}/k$ and we call it the "coupling factor insensitivity" or the "k-insensitivity". Lower indices, such as $\beta_{12(p)}$, indicate the resonators to which the perturbed and unperturbed coupling factor is referred to, namely $\beta_{12(p)} \equiv k_{12(p)}/k_{12}$.

[00120] Similarly, we describe *U* as the strong-coupling factor in the absence of extraneous objects. We denote the strong-coupling factor in the presence of an extraneous object, *p*, as $U_{(p)}$, $U_{(p)} = k_{(p)}\sqrt{Q_{1(p)}Q_{2(p)}}$, and call it the "perturbed strong-coupling factor" or the "perturbed U". Note that the strong-coupling factor U may also be characterized as "unperturbed", when necessary to distinguish from the perturbed strong-coupling factor $U_{(p)}$. Note that the strong-coupling factor U may also be characterized as "unperturbed", when necessary to distinguish from the perturbed strong-coupling factor $U_{(p)}$.

[00121] We define $\delta U_{(p)} \equiv U_{(p)} - U$ and call it the "perturbation on the strong-coupling factor" or the "perturbation on U" due to an extraneous object, p.

[00122] We also define $\Xi_{(p)} \equiv U_{(p)}/U$ and call it the "strong-coupling factor insensitivity" or the "U-insensitivity". Lower indices, such as $\Xi_{12(p)}$, indicate the resonators to which the perturbed and unperturbed coupling factor refers, namely $\Xi_{12(p)} \equiv U_{12(p)}/U_{12}$.

[00123] The efficiency of the energy exchange in a perturbed system may be given by the same formula giving the efficiency of the unperturbed system, where all parameters such as strong-coupling factors, coupling factors, and quality factors are replaced by their perturbed equivalents. For example, in a system of wireless energy transfer including one source and one device resonator, the optimal efficiency may calculated as $\eta_{(p)} = \left[U_{(p)} \Big/ \left(1 + \sqrt{1 + U_{(p)}^{\;2}}\right) \right]^2$. Therefore, in a system of wireless energy exchange which is perturbed by extraneous objects, large perturbed strong-coupling factors, $U_{mn(p)}$, between at least some or all of the multiple resonators may be desired for a high system efficiency in the wireless energy transfer. Source, device and/or intermediate resonators may have a high $Q_{(p)}$.

[00124] Some extraneous perturbations may sometimes be detrimental for the perturbed strong-coupling factors (via large perturbations on the coupling factors or the quality factors). Therefore, techniques may be used to reduce the effect of extraneous perturbations on the system and preserve large strong-coupling factor insensitivites.

[00125] **<u>Efficiency of Energy Exchange</u>**

[00126] The so-called "useful" energy in a useful energy exchange is the energy or power that must be delivered to a device (or devices) in order to power or charge the device. The transfer efficiency that corresponds to a useful energy exchange may be system or application dependent. For example, high power vehicle charging applications that transfer kilowatts of power may need to be at least 80% efficient in order to supply useful amounts of power resulting in a useful energy exchange sufficient to recharge a vehicle battery, without significantly heating up various components of the transfer system. In some consumer electronics applications, a useful energy exchange may include any energy transfer efficiencies greater than 10%, or any other amount acceptable to keep rechargeable batteries "topped off" and running for long periods of time. For some wireless sensor applications, transfer efficiencies that are much less than 1% may be adequate for powering multiple low power sensors from a single source located a significant distance from the sensors. For still other applications, where wired power transfer is either impossible or impractical, a wide range of transfer efficiencies may be acceptable for a useful energy exchange and may be said to supply useful power to devices in those applications. In general, an operating distance is any distance over which a useful energy exchange is or can be maintained according to the principles disclosed herein.

[00127] A useful energy exchange for a wireless energy transfer in a powering or recharging application may be efficient, highly efficient, or efficient enough, as long as the wasted energy levels, heat dissipation, and associated field strengths are within tolerable limits. The tolerable limits may depend on the application, the environment and the system location. Wireless energy transfer for powering or recharging applications may be efficient, highly efficient, or efficient enough, as long as the desired system performance may be attained for the reasonable cost restrictions, weight restrictions, size restrictions, and the like. Efficient energy transfer may be determined relative to that which could be achieved using traditional inductive techniques that are not high-Q systems. Then, the energy transfer may be defined as being efficient, highly efficient, or efficient enough, if more energy is delivered than could be delivered by similarly sized coil structures in traditional inductive schemes over similar distances or alignment offsets.

[00128] Note that, even though certain frequency and Q matching conditions may optimize the system efficiency of energy transfer, these conditions may not need to be exactly met in order to have efficient enough energy transfer for a useful energy exchange. Efficient energy exchange may be realized so long as the relative offset of the resonant frequencies ($|\omega_m - \omega_n| / \sqrt{\omega_m \omega_n}$) is less than approximately the maximum among $1/Q_{m(p)}$, $1/Q_{n(p)}$ and $k_{mn(p)}$. The Q matching condition may be less critical than the frequency matching condition for efficient energy exchange. The degree by which the strong-loading factors, $U_{m[l]}$, of the resonators due to generators and/or loads may be away from their optimal values and still have efficient enough energy exchange depends on the particular system, whether all or some of the generators and/or loads are Q-mismatched and so on.

[00129] Therefore, the resonant frequencies of the resonators may not be exactly matched, but may be matched within the above tolerances. The strong-loading factors of at least some of the resonators due to generators and/or loads may not be exactly matched to their optimal value. The voltage levels, current levels, impedance values, material parameters, and the like may not be at the exact values described in the disclosure but will be within some acceptable tolerance of those values. The system optimization may include cost, size, weight, complexity, and the like, considerations, in addition to efficiency, Q, frequency, strong coupling factor, and the like, considerations. Some system performance parameters, specifications, and designs may be far from optimal in order to optimize other system performance parameters, specifications and designs.

[00130] In some applications, at least some of the system parameters may be varying in time, for example because components, such as sources or devices, may be mobile or aging or because the loads may be variable or because the perturbations or the environmental conditions are changing etc. In these cases, in order to achieve acceptable matching conditions, at least some of the system parameters may need to be dynamically adjustable or tunable. All the system parameters may be dynamically adjustable or tunable to achieve approximately the optimal operating conditions. However, based on the discussion above, efficient enough energy exchange may be realized even if some system parameters are not variable. In some examples, at least some of the devices may not be dynamically adjusted. In some examples, at least some of the sources may not be dynamically adjusted. In some examples, at least some of the intermediate resonators may not be dynamically adjusted. In some examples, none of the system parameters may be dynamically adjusted.

[00131] **Electromagnetic Resonators**

[00132] The resonators used to exchange energy may be electromagnetic resonators. In such resonators, the intrinsic energy decay rates, Γ_m, are given by the absorption (or resistive) losses and the radiation losses of the resonator.

[00133] The resonator may be constructed such that the energy stored by the electric field is primarily confined within the structure and that the energy stored by the magnetic field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant magnetic near-field. These types of resonators may be referred to as magnetic resonators.

[00134] The resonator may be constructed such that the energy stored by the magnetic field is primarily confined within the structure and that the energy stored by the electric field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant electric near-field. These types of resonators may be referred to as electric resonators.

[00135] Note that the total electric and magnetic energies stored by the resonator have to be equal, but their localizations may be quite different. In some cases, the ratio of the average electric field energy to the average magnetic field energy specified at a distance from a resonator may be used to characterize or describe the resonator.

[00136] Electromagnetic resonators may include an inductive element, a distributed inductance, or a combination of inductances with inductance, *L*, and a capacitive element, a distributed capacitance, or a combination of capacitances, with capacitance, *C*. A minimal circuit model of an electromagnetic resonator 102 is shown in Fig. 6a. The resonator may include an inductive element 108 and a capacitive element 104. Provided with initial energy, such as electric field energy stored in the capacitor 104, the system will oscillate as the capacitor discharges transferring energy into magnetic field energy stored in the inductor 108 which in turn transfers energy back into electric field energy stored in the capacitor 104.

[00137] The resonators 102 shown in Figs. 6(b)(c)(d) may be referred to as magnetic resonators. Magnetic resonators may be preferred for wireless energy transfer applications in populated environments because most everyday materials including animals, plants, and humans are non-magnetic (i.e., $\mu_r \approx 1$), so their interaction with magnetic fields is minimal and due primarily to eddy currents induced by the time-variation of the magnetic fields, which is a second-order effect. This characteristic is important both for safety reasons and because it reduces the potential for interactions with extraneous environmental objects and materials that could alter system performance.

[00138] Fig. 6d shows a simplified drawing of some of the electric and magnetic field lines associated with an exemplary magnetic resonator 102B. The magnetic resonator 102B may include a loop of conductor acting as an inductive element 108 and a capacitive element 104 at the ends of the conductor loop. Note that this drawing depicts most of the energy in the region surrounding the resonator being stored in the magnetic field, and most of the energy in the resonator (between the capacitor plates) stored in the electric field. Some electric field, owing to fringing fields, free charges, and the time varying magnetic field, may be stored in the region around the resonator, but the magnetic resonator may be designed to confine the electric fields to be close to or within the resonator itself, as much as possible.

[00139] The inductor 108 and capacitor 104 of an electromagnetic resonator 102 may be bulk circuit elements, or the inductance and capacitance may be distributed and may result from the way the conductors are formed, shaped, or positioned, in the structure. For example, the inductor 108 may be realized by shaping a conductor to enclose a surface area, as shown in Figs. 6(b)(c)(d). This type of resonator 102 may be referred to as a capacitively-loaded loop inductor. Note that we may use the terms “loop” or “coil” to indicate generally a conducting structure (wire, tube, strip, etc.), enclosing a surface of any shape and dimension, with any number of turns. In Fig. 6b, the enclosed surface area is circular, but the surface may be any of a wide variety of other shapes and sizes and may be designed to achieve certain system performance specifications. As an example to indicate how inductance scales with physical dimensions, the inductance for a length of circular conductor arranged to form a circular single-turn loop is approximately,

$$L = \mu_0 x(\ln\frac{8x}{a} - 2),$$

where μ_0 is the magnetic permeability of free space, $x,$ is the radius of the enclosed circular surface area and, $a,$ is the radius of the conductor used to form the inductor loop. A more precise value of the inductance of the loop may be calculated analytically or numerically.

[00140] The inductance for other cross-section conductors, arranged to form other enclosed surface shapes, areas, sizes, and the like, and of any number of wire turns, may be calculated analytically, numerically or it may be determined by measurement. The inductance may be realized using inductor elements, distributed inductance, networks, arrays, series and parallel combinations of inductors and inductances, and the like. The inductance may be fixed or variable and may be used to vary impedance matching as well as resonant frequency operating conditions.

[00141] There are a variety of ways to realize the capacitance required to achieve the desired resonant frequency for a resonator structure. Capacitor plates 110 may be formed and utilized as shown in Fig. 6b, or the capacitance may be distributed and be realized between adjacent windings of a multi-loop conductor 114, as shown in Fig. 6c. The capacitance may be realized using capacitor elements, distributed capacitance, networks, arrays, series and parallel combinations of capacitances, and the like. The capacitance may be fixed or variable and may be used to vary impedance matching as well as resonant frequency operating conditions.

[00142] It is to be understood that the inductance and capacitance in an electromagnetic resonator 102 may be lumped, distributed, or a combination of lumped and distributed inductance and capacitance and that electromagnetic resonators may be realized by combinations of the various elements, techniques and effects described herein.

[00143] Electromagnetic resonators 102 may be include inductors, inductances, capacitors, capacitances, as well as additional circuit elements such as resistors, diodes, switches, amplifiers, diodes, transistors, transformers, conductors, connectors and the like.

[00144] **<u>Resonant Frequency of an Electromagnetic Resonator</u>**

[00145] An electromagnetic resonator 102 may have a characteristic, natural, or resonant frequency determined by its physical properties. This resonant frequency is the frequency at which the energy stored by the resonator oscillates between that stored by the electric field, W_E, ($W_E=q^2/2C$, where q is the charge on the capacitor, C) and that stored by the magnetic field, W_B, ($W_B=Li^2/2$, where i is the current through the inductor, L) of the resonator. In the absence of any losses in the system, energy would continually be exchanged between the electric field in the capacitor 104 and the magnetic field in the inductor 108. The frequency at which this energy is exchanged may be called the characteristic frequency, the natural frequency, or the resonant frequency of the resonator, and is given by ω,

$$\omega = 2\pi f = \sqrt{\frac{1}{LC}}.$$

[00146] The resonant frequency of the resonator may be changed by tuning the inductance, L, and/or the capacitance, C, of the resonator. The resonator frequency may be design to operate at the so-called ISM (Industrial, Scientific and Medical) frequencies as specified by the FCC. The resonator frequency may be chosen to meet certain field limit specifications, specific absorption rate (SAR) limit specifications, electromagnetic compatibility (EMC) specifications, electromagnetic interference (EMI) specifications, component size, cost or performance specifications, and the like.

[00147] <u>Quality Factor of an Electromagnetic Resonator</u>

[00148] The energy in the resonators 102 shown in Fig. 6 may decay or be lost by intrinsic losses including absorptive losses (also called ohmic or resistive losses) and/or radiative losses. The Quality Factor, or Q, of the resonator, which characterizes the energy decay, is inversely proportional to these losses. Absorptive losses may be caused by the finite conductivity of the conductor used to form the inductor as well as by losses in other elements, components, connectors, and the like, in the resonator. An inductor formed from low loss materials may be referred to as a "high-Q inductive element" and elements, components, connectors and the like with low losses may be referred to as having "high resistive Q's". In general, the total absorptive loss for a resonator may be calculated as the appropriate series and/or parallel combination of resistive losses for the various elements and components that make up the resonator. That is, in the absence of any significant radiative or component/connection losses, the Q of the resonator may be given by, Q_{abs},

$$Q_{abs} = \frac{\omega L}{R_{abs}},$$

where ω, is the resonant frequency, L, is the total inductance of the resonator and the resistance for the conductor used to form the inductor, for example, may be given by $R_{abs} = l\rho/A$, (l is the length of the wire, ρ is the resistivity of the conductor material, and A is the cross-sectional area over which current flows in the wire). For alternating currents, the cross-sectional area over which current flows may be less than the physical cross-sectional area of the conductor owing to the skin effect. Therefore, high-Q magnetic resonators may be composed of conductors with high conductivity, relatively large surface areas and/or with specially designed profiles (e.g. Litz wire) to minimize proximity effects and reduce the AC resistance.

[00149] The magnetic resonator structures may include high-Q inductive elements composed of high conductivity wire, coated wire, Litz wire, ribbon, strapping or plates, tubing, paint, gels, traces, and the like. The magnetic resonators may be self-resonant, or they may include external coupled elements such as capacitors, inductors, switches, diodes, transistors, transformers, and the like. The magnetic resonators may include distributed and lumped capacitance and inductance. In general, the Q of the resonators will be determined by the Q's of all the individual components of the resonator.

[00150] Because Q is proportional to inductance, L, resonators may be designed to increase L, within certain other constraints. One way to increase L, for example, is to use more than one turn of the conductor to form the inductor in the resonator. Design techniques and trade-offs may depend on the application, and a wide variety of structures, conductors, components, and resonant frequencies may be chosen in the design of high-Q magnetic resonators.

[00151] In the absence of significant absorption losses, the Q of the resonator may be determined primarily by the radiation losses, and given by, $Q_{rad} = \omega L / R_{rad}$, where R_{rad} is the radiative loss of the resonator and may depend on the size of the resonator relative to the frequency, ω, or wavelength, λ, of operation. For the magnetic resonators discussed above, radiative losses may scale as $R_{rad} \sim (x/\lambda)^4$ (characteristic of magnetic dipole radiation), where x is a characteristic dimension of the resonator, such as the radius of the inductive element shown in Fig. 6b, and where $\lambda = c / f$, where c is the speed of light and f is as defined above. The size of the magnetic resonator may be much less than the wavelength of operation so radiation losses may be very small. Such structures may be referred to as sub-wavelength resonators. Radiation may be a loss mechanism for non-radiative wireless energy transfer systems and designs may be chosen to reduce or minimize R_{rad}. Note that a high-Q_{rad} may be desirable for non-radiative wireless energy transfer schemes.

[00152] Note too that the design of resonators for non-radiative wireless energy transfer differs from antennas designed for communication or far-field energy transmission purposes. Specifically, capacitively-loaded conductive loops may be used as resonant antennas (for example in cell phones), but those operate in the far-field regime where the radiation Q's are intentionally designed to be small to make the antenna efficient at radiating energy. Such designs are not appropriate for the efficient near-field wireless energy transfer technique disclosed in this application.

[00153] The quality factor of a resonator including both radiative and absorption losses is $Q = \omega L / (R_{abs} + R_{rad})$. Note that there may be a maximum Q value for a particular resonator and that resonators may be designed with special consideration given to the size of the resonator, the materials and elements used to construct the resonator, the operating frequency, the connection mechanisms, and the like, in order to achieve a high-Q resonator. Fig. 7 shows a plot of Q of an exemplary magnetic resonator (in this case a coil with a diameter of 60 cm made of copper pipe with an outside diameter (OD) of 4 cm) that may be used for wireless power transmission at MHz frequencies. The absorptive Q (dashed line) 702 increases with frequency, while the radiative Q (dotted line) 704 decreases with frequency, thus leading the overall Q to peak 708 at a particular frequency. Note that the Q of this exemplary resonator is greater than 100 over a wide frequency range. Magnetic resonators may be designed to have high-Q over a range of frequencies and system operating frequency may set to any frequency in that range.

[00154] When the resonator is being described in terms of loss rates, the Q may be defined using the intrinsic decay rate, 2Γ, as described previously. The intrinsic decay rate is the rate at which an uncoupled and undriven resonator loses energy. For the magnetic resonators described above, the intrinsic loss rate may be given by $\Gamma = (R_{abs} + R_{rad})/2L$, and the quality factor, Q, of the resonator is given by $Q = \omega/2\Gamma$.

[00155] Note that a quality factor related only to a specific loss mechanism may be denoted as $Q_{mechanism}$, if the resonator is not specified, or as $Q_{1,mechanism}$, if the resonator is specified (e.g. resonator 1). For example, $Q_{1,rad}$ is the quality factor for resonator 1 related to its radiation losses.

[00156] **Electromagnetic Resonator Near-Fields**

[00157] The high-Q electromagnetic resonators used in the near-field wireless energy transfer system disclosed here may be sub-wavelength objects. That is, the physical dimensions of the resonator may be much smaller than the wavelength corresponding to the resonant frequency. Sub-wavelength magnetic resonators may have most of the energy in the region surrounding the resonator stored in their magnetic near-fields, and these fields may also be described as stationary or non-propagating because they do not radiate away from the resonator. The extent of the near-field in the area surrounding the resonator is typically set by the wavelength, so it may extend well beyond the resonator itself for a sub-wavelength resonator. The limiting surface, where the field behavior changes from near-field behavior to far-field behavior may be called the “radiation caustic”.

[00158] The strength of the near-field is reduced the farther one gets away from the resonator. While the field strength of the resonator near-fields decays away from the resonator, the fields may still interact with objects brought into the general vicinity of the resonator. The degree to which the fields interact depends on a variety of factors, some of which may be controlled and designed, and some of which may not. The wireless energy transfer schemes described herein may be realized when the distance between coupled resonators is such that one resonator lies within the radiation caustic of the other.

[00159] The near-field profiles of the electromagnetic resonators may be similar to those commonly associated with dipole resonators or oscillators. Such field profiles may be described as omni-directional, meaning the magnitudes of the fields are non-zero in all directions away from the object.

[00160] Characteristic Size of An Electromagnetic Resonator

[00161] Spatially separated and/or offset magnetic resonators of sufficient Q may achieve efficient wireless energy transfer over distances that are much larger than have been seen in the prior art, even if the sizes and shapes of the resonator structures are different. Such resonators may also be operated to achieve more efficient energy transfer than was achievable with previous techniques over shorter range distances. We describe such resonators as being capable of mid-range energy transfer.

[00162] Mid-range distances may be defined as distances that are larger than the characteristic dimension of the smallest of the resonators involved in the transfer, where the distance is measured from the center of one resonator structure to the center of a spatially separated second resonator structure. In this definition, two-dimensional resonators are spatially separated when the areas circumscribed by their inductive elements do not intersect and three-dimensional resonators are spatially separated when their volumes do not intersect. A two-dimensional resonator is spatially separated from a three-dimensional resonator when the area circumscribed by the former is outside the volume of the latter.

[00163] Fig. 8 shows some example resonators with their characteristic dimensions labeled. It is to be understood that the characteristic sizes 802 of resonators 102 may be defined in terms of the size of the conductor and the area circumscribed or enclosed by the inductive element in a magnetic resonator and the length of the conductor forming the capacitive element of an electric resonator. Then, the characteristic size 802 of a resonator 102, x_{char}, may be equal to the radius of the smallest sphere that can fit around the inductive or capacitive element of the magnetic or electric resonator respectively, and the center of the resonator structure is the center of the sphere. The characteristic thickness 804, t_{char}, of a resonator 102 may be the smallest possible height of the highest point of the inductive or capacitive element in the magnetic or capacitive resonator respectively, measured from a flat surface on which it is placed. The characteristic width 808 of a resonator 102, w_{char}, may be the radius of the smallest possible circle through which the inductive or capacitive element of the magnetic or electric resonator respectively, may pass while traveling in a straight line. For example, the characteristic width 808 of a cylindrical resonator may be the radius of the cylinder.

[00164] In this inventive wireless energy transfer technique, energy may be exchanged efficiently over a wide range of distances, but the technique is distinguished by the ability to exchange useful energy for powering or recharging devices over mid-range distances and between resonators with different physical dimensions, components and orientations. Note that while k may be small in these circumstances, strong coupling and efficient energy transfer may be realized by using high-Q resonators to achieve a high U, $U = k\sqrt{Q_s Q_d}$. That is, increases in Q may be used to at least partially overcome decreases in k, to maintain useful energy transfer efficiencies.

[00165] Note too that while the near-field of a single resonator may be described as omni-directional, the efficiency of the energy exchange between two resonators may depend on the relative position and orientation of the resonators. That is, the efficiency of the energy exchange may be maximized for particular relative orientations of the resonators. The sensitivity of the transfer efficiency to the relative position and orientation of two uncompensated resonators may be captured in the calculation of either k or κ . While coupling may be achieved between resonators that are offset and/or rotated relative to each other, the efficiency of the exchange may depend on the details of the positioning and on any feedback, tuning, and compensation techniques implemented during operation.

[00166] **High-*Q* Magnetic Resonators**

[00167] In the near-field regime of a sub-wavelength capacitively-loaded loop magnetic resonator ($x \ll \lambda$), the resistances associated with a circular conducting loop inductor composed of N turns of wire whose radius is larger than the skin depth, are approximately $R_{abs} = \sqrt{\mu_o \rho \omega / 2} \cdot Nx / a$ and $R_{rad} = \pi / 6 \cdot \eta_o N^2 (\omega x / c)^4$, where ρ is the resistivity of the conductor material and $\eta_o \approx 120\pi\ \Omega$ is the impedance of free space. . The inductance, L, for such a N-turn loop is approximately N^2 times the inductance of a single-turn loop given previously. The quality factor of such a resonator, $Q = \omega L / (R_{abs} + R_{rad})$, is highest for a particular frequency determined by the system parameters (Fig. 4). As described previously, at lower frequencies the Q is determined primarily by absorption losses and at higher frequencies the Q is determined primarily by radiation losses.

[00168] Note that the formulas given above are approximate and intended to illustrate the functional dependence of R_{abs}, R_{rad} and L on the physical parameters of the structure. More accurate numerical calculations of these parameters that take into account deviations from the strict quasi-static limit, for example a non-uniform current/charge distribution along the conductor, may be useful for the precise design of a resonator structure.

[00169] Note that the absorptive losses may be minimized by using low loss conductors to form the inductive elements. The loss of the conductors may be minimized by using large surface area conductors such as conductive tubing, strapping, strips, machined objects, plates, and the like, by using specially designed conductors such as Litz wire, braided wires, wires of any cross-section, and other conductors with low proximity losses, in which case the frequency scaled behavior described above may be different, and by using low resistivity materials such as high-purity copper and silver, for example. One advantage of using conductive tubing as the conductor at higher operating frequencies is that it may be cheaper and lighter than a similar diameter solid conductor, and may have similar resistance because most of the current is traveling along the outer surface of the conductor owing to the skin effect.

[00170] To get a rough estimate of achievable resonator designs made from copper wire or copper tubing and appropriate for operation in the microwave regime, one may calculate the optimum Q and resonant frequency for a resonator composed of one circular inductive element ($N=1$) of copper wire ($\rho=1.69\cdot10^{-8}\,\Omega m$) with various cross sections. Then for an inductive element with characteristic size $x=1\ cm$ and conductor diameter a=1 mm, appropriate for a cell phone for example, the quality factor peaks at Q=1225 when $f=380$ MHz. For $x=30\ cm$ and $a=2\ mm$, an inductive element size that might be appropriate for a laptop or a household robot, $Q=1103$ at $f=17$ MHz. For a larger source inductive element that might be located in the ceiling for example, x=1 m and a=4 mm, Q may be as high as Q=1315 at $f=5$ MHz. Note that a number of practical examples yield expected quality factors of $Q\approx1000\text{-}1500$ at $\lambda/x\approx50\text{-}80$. Measurements of a wider variety of coil shapes, sizes, materials and operating frequencies than described above show that Q's >100 may be realized for a variety of magnetic resonator structures using commonly available materials.

[00171] As described above, the rate for energy transfer between two resonators of characteristic size x_1 and x_2, and separated by a distance D between their centers, may be given by κ . To give an example of how the defined parameters scale, consider the cell phone, laptop, and ceiling resonator examples from above, at three (3) distances; D/x=10, 8, 6. In the examples considered here, the source and device resonators are the same size, x_1=x_2, and shape, and are oriented as shown in Fig. 1(b). In the cell phone example, $\omega/2\kappa$ =3033, 1553, 655 respectively. In the laptop example, $\omega/2\kappa$ =7131, 3651, 1540 respectively and for the ceiling resonator example, $\omega/2\kappa$ =6481, 3318, 1400. The corresponding coupling-to-loss ratios peak at the frequency where the inductive element Q peaks and are κ/Γ =0.4, 0.79, 1.97 and 0.15, 0.3, 0.72 and 0.2, 0.4, 0.94 for the three inductive element sizes and distances described above. An example using different sized inductive elements is that of an x_1=1 m inductor (e.g. source in the ceiling) and an x_2=30 cm inductor (e.g. household robot on the floor) at a distance D=3 m apart (e.g. room height). In this example, the strong-coupling figure of merit, $U=\kappa/\sqrt{\Gamma_1\Gamma_2}$ =0.88, for an efficiency of approximately 14%, at the optimal operating frequency of f=6.4 MHz. Here, the optimal system operating frequency lies between the peaks of the individual resonator Q's.

[00172] Inductive elements may be formed for use in high-Q magnetic resonators. We have demonstrated a variety of high-Q magnetic resonators based on copper conductors that are formed into inductive elements that enclose a surface. Inductive elements may be formed using a variety of conductors arranged in a variety of shapes, enclosing any size or shaped area, and they may be single turn or multiple turn elements. Drawings of exemplary inductive elements 900A-B are shown in Fig. 9. The inductive elements may be formed to enclose a circle, a rectangle, a square, a triangle, a shape with rounded corners, a shape that follows the contour of a particular structure or device, a shape that follows, fills, or utilizes, a dedicated space within a structure or device, and the like. The designs may be optimized for size, cost, weight, appearance, performance, and the like.

[00173] These conductors may be bent or formed into the desired size, shape, and number of turns. However, it may be difficult to accurately reproduce conductor shapes and sizes using manual techniques. In addition, it may be difficult to maintain uniform or desired center-to-center spacings between the conductor segments in adjacent turns of the inductive elements. Accurate or uniform spacing may be important in determining the self capacitance of the structure as well as any proximity effect induced increases in AC resistance, for example.

[00174] Molds may be used to replicate inductor elements for high-*Q* resonator designs. In addition, molds may be used to accurately shape conductors into any kind of shape without creating kinks, buckles or other potentially deleterious effects in the conductor. Molds may be used to form the inductor elements and then the inductor elements may be removed from the forms. Once removed, these inductive elements may be built into enclosures or devices that may house the high-Q magnetic resonator. The formed elements may also or instead remain in the mold used to form them.

[00175] The molds may be formed using standard CNC (computer numerical control) routing or milling tools or any other known techniques for cutting or forming grooves in blocks. The molds may also or instead be formed using machining techniques, injection molding techniques, casting techniques, pouring techniques, vacuum techniques, thermoforming techniques, cut-in-place techniques, compression forming techniques and the like.

[00176] The formed element may be removed from the mold or it may remain in the mold. The mold may be altered with the inductive element inside. The mold may be covered, machined, attached, painted and the like. The mold and conductor combination may be integrated into another housing, structure or device. The grooves cut into the molds may be any dimension and may be designed to form conducting tubing, wire, strapping, strips, blocks, and the like into the desired inductor shapes and sizes.

[00177] The inductive elements used in magnetic resonators may contain more than one loop and may spiral inward or outward or up or down or in some combination of directions. In general, the magnetic resonators may have a variety of shapes, sizes and number of turns and they may be composed of a variety of conducing materials.

[00178] The magnetic resonators may be free standing or they may be enclosed in an enclosure, container, sleeve or housing. The magnetic resonators may include the form used to make the inductive element. These various forms and enclosures may be composed of almost any kind of material. Low loss materials such as Teflon, REXOLITE, styrene, and the like may be preferable for some applications. These enclosures may contain fixtures that hold the inductive elements.

[00179] Magnetic resonators may be composed of self-resonant coils of copper wire or copper tubing. Magnetic resonators composed of self resonant conductive wire coils may include a wire of length *l,* and cross section radius *a,* wound into a helical coil of radius *x*, height *h,* and number of turns *N*, which may for example be characterized as $N = \sqrt{l^2 - h^2} / 2\pi x$.

[00180] A magnetic resonator structure may be configured so that *x* is about 30 cm, *h* is about 20 cm, *a* is about 3 mm and *N* is about 5.25, and, during operation, a power source coupled to the magnetic resonator may drive the resonator at a resonant frequency, *f*, where *f* is about 10.6 MHz. Where *x* is about 30 cm, *h* is about 20 cm, *a* is about 1 cm and *N* is about 4, the resonator may be driven at a frequency, *f*, where *f* is about 13.4 MHz. Where *x* is about 10 cm, *h* is about 3 cm, *a* is about 2 mm and *N* is about 6, the resonator may be driven at a frequency, *f*, where *f* is about 21.4 MHz.

[00181] High-Q inductive elements may be designed using printed circuit board traces. Printed circuit board traces may have a variety of advantages compared to mechanically formed inductive elements including that they may be accurately reproduced and easily integrated using established printed circuit board fabrication techniques, that their AC resistance may be lowered using custom designed conductor traces, and that the cost of mass-producing them may be significantly reduced.

[00182] High-Q inductive elements may be fabricated using standard PCB techniques on any PCB material such as FR-4 (epoxy E-glass), multi-functional epoxy, high performance epoxy, bismalaimide triazine/epoxy, polyimide, Cyanate Ester, polytetraflouroethylene (Teflon), FR-2, FR-3, CEM-1, CEM-2, Rogers, Resolute, and the like. The conductor traces may be formed on printed circuit board materials with lower loss tangents.

[00183] The conducting traces may be composed of copper, silver, gold, aluminum, nickel and the like, and they may be composed of paints, inks, or other cured materials. The circuit board may be flexible and it may be a flex-circuit. The conducting traces may be formed by chemical deposition, etching, lithography, spray deposition, cutting, and the like. The conducting traces may be applied to form the desired patterns and they may be formed using crystal and structure growth techniques.

[00184] The dimensions of the conducting traces, as well as the number of layers containing conducting traces, the position, size and shape of those traces and the architecture for interconnecting them may be designed to achieve or optimize certain system specifications such as resonator *Q*, $Q_{(p)}$, resonator size, resonator material and fabrication costs, *U,* $U_{(p)}$, and the like.

[00185] As an example, a three-turn high-Q inductive element 1001A was fabricated on a four-layer printed circuit board using the rectangular copper trace pattern as shown in Fig. 10(a). The copper trace is shown in black and the PCB in white. The width and thickness of the copper traces in this example was approximately 1 cm (400 mils) and 43 μm (1.7 mils) respectively. The edge-to-edge spacing between turns of the conducting trace on a single layer was approximately 0.75 cm (300 mils) and each board layer thickness was approximately 100 μm (4 mils). The pattern shown in Fig. 10(a) was repeated on each layer of the board and the conductors were connected in parallel. The outer dimensions of the 3-loop structure were approximately 30 cm by 20 cm. The measured inductance of this PCB loop was 5.3 μH. A magnetic resonator using this inductor element and tunable capacitors had a quality factor, *Q*, of 550 at its designed resonance frequency of 6.78 MHz. The resonant frequency could be tuned by changing the inductance and capacitance values in the magnetic resonator.

[00186] As another example, a two-turn inductor 1001B was fabricated on a four-layer printed circuit board using the rectangular copper trace pattern shown in Fig. 10(b). The copper trace is shown in black and the PCB in white. The width and height of the copper traces in this example were approximately 0.75 cm (300 mils) and 43 μm (1.7 mils) respectively. The edge-to-edge spacing between turns of the conducting trace on a single layer was approximately 0.635 cm (250 mils) and each board layer thickness was approximately 100 μm (4 mils). The pattern shown in Fig. 10(b) was repeated on each layer of the board and the conductors were connected in parallel. The outer dimensions of the two-loop structure were approximately 7.62 cm by 26.7 cm. The measured inductance of this PCB loop was 1.3 μH. Stacking two boards together with a vertical separation of approximately 0.635 cm (250 mils) and connecting the two boards in series produced a PCB inductor with an inductance of approximately 3.4 μH. A magnetic resonator using this stacked inductor loop and tunable capacitors had a quality factor, *Q*, of 390 at its designed resonance frequency of 6.78 MHz. The resonant frequency could be tuned by changing the inductance and capacitance values in the magnetic resonator.

[00187] The inductive elements may be formed using magnetic materials of any size, shape thickness, and the like, and of materials with a wide range of permeability and loss values. These magnetic materials may be solid blocks, they may enclose hollow volumes, they may be formed from many smaller pieces of magnetic material tiled and or stacked together, and they may be integrated with conducting sheets or enclosures made from highly conducting materials. Wires may be wrapped around the magnetic materials to generate the magnetic near-field. These wires may be wrapped around one or more than one axis of the structure. Multiple wires may be wrapped around the magnetic materials and combined in parallel, or in series, or via a switch to form customized near-field patterns.

[00188] The magnetic resonator may include 15 turns of Litz wire wound around a 19.2 cm x 10 cm x 5 mm tiled block of 3F3 ferrite material. The Litz wire may be wound around the ferrite material in any direction or combination of directions to achieve the desire resonator performance. The number of turns of wire, the spacing between the turns, the type of wire, the size and shape of the magnetic materials and the type of magnetic material are all design parameters that may be varied or optimized for different application scenarios.

[00189] **High-*Q* Magnetic resonators using magnetic material structures**

[00190] It may be possible to use magnetic materials assembled to form an open magnetic circuit, albeit one with an air gap on the order of the size of the whole structure, to realize a magnetic resonator structure. In these structures, high conductivity materials are wound around a structure made from magnetic material to form the inductive element of the magnetic resonator. Capacitive elements may be connected to the high conductivity materials, with the resonant frequency then determined as described above. These magnetic resonators have their dipole moment in the plane of the two dimensional resonator structures, rather than perpendicular to it, as is the case for the capacitively-loaded inductor loop resonators.

[00191] A diagram of a single planar resonator structure is shown in Fig. 11(a). The planar resonator structure is constructed of a core of magnetic material 1121, such as ferrite with a loop or loops of conducting material 1122 wrapped around the core 1121. The structure may be used as the source resonator that transfers power and the device resonator that captures energy. When used as a source, the ends of the conductor may be coupled to a power source. Alternating electrical current flowing through the conductor loops excites alternating magnetic fields. When the structure is being used to receive power, the ends of the conductor may be coupled to a power drain or load. Changing magnetic fields induce an electromotive force in the loop or loops of the conductor wound around the core magnetic material. The dipole moment of these types of structures is in the plane of the structures and is, for example, directed along the Y axis for the structure in Figure 11(a). Two such structures have strong coupling when placed substantially in the same plane (i.e. the X,Y plane of Figure 11). The structures of Figure 11(a) have the most favorable orientation when the resonators are aligned in the same plane along their Y axis.

[00192] The geometry and the coupling orientations of the described planar resonators may be preferable for some applications. The planar or flat resonator shape may be easier to integrate into many electronic devices that are relatively flat and planar. The planar resonators may be integrated into the whole back or side of a device without requiring a change in geometry of the device. Due to the flat shape of many devices, the natural position of the devices when placed on a surface is to lay with their largest dimension being parallel to the surface they are placed on. A planar resonator integrated into a flat device is naturally parallel to the plane of the surface and is in a favorable coupling orientation relative to the resonators of other devices or planar resonator sources placed on a flat surface.

[00193] As mentioned, the geometry of the planar resonators may allow easier integration into devices. Their low profile may allow a resonator to be integrated into or as part of a complete side of a device. When a whole side of a device is covered by the resonator, magnetic flux can flow through the resonator core without being obstructed by lossy material that may be part of the device or device circuitry.

[00194] The core of the planar resonator structure may be of a variety of shapes and thicknesses and may be flat or planar such that the minimum dimension does not exceed 30% of the largest dimension of the structure. The core may have complex geometries and may have indentations, notches, ridges, and the like. Geometric enhancements may be used to reduce the coupling dependence on orientation and they may be used to facilitate integration into devices, packaging, packages, enclosures, covers, skins, and the like. Two exemplary variations of core geometries are shown in Figure 11(b). For example, the planar core 1131 may be shaped such that the ends are substantially wider than the middle of the structure to create an indentation for the conductor winding. The core material may be of varying thickness with ends that are thicker and wider than the middle. The core material 1132 may have any number of notches or cutouts 1133 of various depths, width, and shapes to accommodate conductor loops, housing, packaging, and the like.

[00195] The shape and dimensions of the core may be further dictated by the dimensions and characteristics of the device that they are integrated into. The core material may curve to follow the contours of the device, or may require non-symmetric notches or cutouts to allow clearance for parts of the device. The core structure may be a single monolithic piece of magnetic material or may be composed of a plurality of tiles, blocks, or pieces that are arranged together to form the larger structure. The different layers, tiles, blocks, or pieces of the structure may be of similar or may be of different materials. It may be desirable to use materials with different magnetic permeability in different locations of the structure. Core structures with different magnetic permeability may be useful for guiding the magnetic flux, improving coupling, and affecting the shape or extent of the active area of a system.

[00196] The conductor of the planar resonator structure may be wound at least once around the core. In certain circumstances, it may be preferred to wind at least three loops. The conductor can be any good conductor including conducting wire, Litz wire, conducting tubing, sheets, strips, gels, inks, traces and the like.

[00197] The size, shape, or dimensions of the active area of source may be further enhanced, altered, or modified with the use of materials that block, shield, or guide magnetic fields. To create non-symmetric active area around a source once side of the source may be covered with a magnetic shield to reduce the strength of the magnetic fields in a specific direction. The shield may be a conductor or a layered combination of conductor and magnetic material which can be used to guide magnetic fields away from a specific direction. Structures composed of layers of conductors and magnetic materials may be used to reduce energy losses that may occur due to shielding of the source.

[00198] The plurality of planar resonators may be integrated or combined into one planar resonator structure. A conductor or conductors may be wound around a core structure such that the loops formed by the two conductors are not coaxial. An example of such a structure is shown in Figure 12 where two conductors 1201,1202 are wrapped around a planar rectangular core 1203 at orthogonal angles. The core may be rectangular or it may have various geometries with several extensions or protrusions. The protrusions may be useful for wrapping of a conductor, reducing the weight, size, or mass of the core, or may be used to enhance the directionality or omni-directionality of the resonator. A multi wrapped planar resonator with four protrusions is shown by the inner structure 1310 in Figure 13, where four conductors 1301, 1302, 1303, 1304 are wrapped around the core. The core may have extensions 1305,1306,1307,1308 with one or more conductor loops. A single conductor may be wrapped around a core to form loops that are not coaxial. The four conductor loops of Figure 13, for example, may be formed with one continuous piece of conductor, or using two conductors where a single conductor is used to make all coaxial loops.

[00199] Non-uniform or asymmetric field profiles around the resonator comprising a plurality of conductor loops may be generated by driving some conductor loops with non-identical parameters. Some conductor loops of a source resonator with a plurality of conductor loops may be driven by a power source with a different frequency, voltage, power level, duty cycle, and the like all of which may be used to affect the strength of the magnetic field generated by each conductor.

[00200] The planar resonator structures may be combined with a capacitively-loaded inductor resonator coil to provide an omni-directional active area all around, including above and below the source while maintaining a flat resonator structure. As shown in Figure 13, an additional resonator loop coil 1309 comprising of a loop or loops of a conductor, may be placed in a common plane as the planar resonator structure 1310. The outer resonator coil provides an active area that is substantially above and below the source. The resonator coil can be arranged with any number of planar resonator structures and arrangements described herein.

[00201] The planar resonator structures may be enclosed in magnetically permeable packaging or integrated into other devices. The planar profile of the resonators within a single, common plane allows packaging and integration into flat devices. A diagram illustrating the application of the resonators is shown in Figure 14. A flat source 1411 comprising one or more planar resonators 1414 each with one or more conductor loops may transfer power to devices 1412,1413 that are integrated with other planar resonators 1415,1416 and placed within an active area 1417 of the source. The devices may comprise a plurality of planar resonators such that regardless of the orientation of the device with respect to the source the active area of the source does not change. In addition to invariance to rotational misalignment, a flat device comprising of planar resonators may be turned upside down without substantially affecting the active area since the planar resonator is still in the plane of the source.

[00202] Another diagram illustrating a possible use of a power transfer system using the planar resonator structures is shown in Figure 15. A planar source 1521 placed on top of a surface 1525 may create an active area that covers a substantial surface area creating an "energized surface" area. Devices such as computers 1524, mobile handsets 1522, games, and other electronics 1523 that are coupled to their respective planar device resonators may receive energy from the source when placed within the active area of the source, which may be anywhere on top of the surface. Several devices with different dimensions may be placed in the active area and used normally while charging or being powered from the source without having strict placement or alignment constraints. The source may be placed under the surface of a table, countertop, desk, cabinet, and the like, allowing it to be completely hidden while energizing the top surface of the table, countertop, desk, cabinet and the like, creating an active area on the surface that is much larger than the source.

[00203] The source may include a display or other visual, auditory, or vibration indicators to show the direction of charging devices or what devices are being charged, error or problems with charging, power levels, charging time, and the like.

[00204] The source resonators and circuitry may be integrated into any number of other devices. The source may be integrated into devices such as clocks, keyboards, monitors, picture frames, and the like. For example, a keyboard integrated with the planar resonators and appropriate power and control circuitry may be used as a source for devices placed around the keyboard such as computer mice, webcams, mobile handsets, and the like without occupying any additional desk space.

[00205] While the planar resonator structures have been described in the context of mobile devices it should be clear to those skilled in the art that a flat planar source for wireless power transfer with an active area that extends beyond its physical dimensions has many other consumer and industrial applications. The structures and configuration may be useful for a large number of applications where electronic or electric devices and a power source are typically located, positioned, or manipulated in substantially the same plane and alignment. Some of the possible application scenarios include devices on walls, floor, ceilings or any other substantially planar surfaces.

[00206] Flat source resonators may be integrated into a picture frame or hung on a wall thereby providing an active area within the plane of the wall where other electronic devices such as digital picture frames, televisions, lights, and the like can be mounted and powered without wires. Planar resonators may be integrated into a floor resulting in an energized floor or active area on the floor on which devices can be placed to receive power. Audio speakers, lamps, heaters, and the like can be placed within the active are and receive power wirelessly.

[00207] The planar resonator may have additional components coupled to the conductor. Components such as capacitors, inductors, resistors, diodes, and the like may be coupled to the conductor and may be used to adjust or tune the resonant frequency and the impedance matching for the resonators.

[00208] A planar resonator structure of the type described above and shown in Fig. 11(a), may be created, for example, with a quality factor, Q, of 100 or higher and even Q of 1,000 or higher. Energy may be wirelessly transferred from one planar resonator structure to another over a distance larger than the characteristic size of the resonators, as shown in Fig. 11(c).

[00209] In addition to utilizing magnetic materials to realize a structure with properties similar to the inductive element in the magnetic resonators, it may be possible to use a combination of good conductor materials and magnetic material to realize such inductive structures. Fig. 16(a) shows a magnetic resonator structure 1602 that may include one or more enclosures made of high-conductivity materials (the inside of which would be shielded from AC electromagnetic fields generated outside) surrounded by at least one layer of magnetic material and linked by blocks of magnetic material 1604.

A structure may include a high-conductivity sheet of material covered on one side by a layer of magnetic material. The layered structure may instead be applied conformally to an electronic device, so that parts of the device may be covered by the high-conductivity and magnetic material layers, while other parts that need to be easily accessed (such as buttons or screens) may be left uncovered. The structure may also or instead include only layers or bulk pieces of magnetic material. Thus, a magnetic resonator may be incorporated into an existing device without significantly interfering with its existing functions and with little or no need for extensive redesign. Moreover, the layers of good conductor and/or magnetic material may be made thin enough (of the order of a millimeter or less) that they would add little extra weight and volume to the completed device. An oscillating current applied to a length of conductor wound around the structure, as shown by the square loop in the center of the structure in Figure 16 may be used to excite the electromagnetic fields associated with this structure.

[00210] <u>Quality factor of the structure</u>

[00211] A structure of the type described above may be created with a quality factor, Q, of the order of 1,000 or higher. This high-Q is possible even if the losses in the magnetic material are high, if the fraction of magnetic energy within the magnetic material is small compared to the total magnetic energy associated with the object. For structures composed of layers conducting materials and magnetic materials, the losses in the conducting materials may be reduced by the presence of the magnetic materials as described previously. In structures where the magnetic material layer's thickness is of the order of 1/100 of the largest dimension of the system (e.g., the magnetic material may be of the order of 1 mm thick, while the area of the structure is of the order of 10 cm x 10 cm), and the relative permeability is of the order of 1,000, it is possible to make the fraction of magnetic energy contained within the magnetic material only a few hundredths of the total magnetic energy associated with the object or resonator. To see how that comes about, note that the expression for the magnetic energy contained in a volume is $U_m = \int_V d\mathbf{r}\mathbf{B}(\mathbf{r})^2 / (2\mu_r\mu_0)$, so as long as ***B*** (rather than ***H***) is the main field conserved across the magnetic material-air interface (which is typically the case in open magnetic circuits), the fraction of magnetic energy contained in the high-μ_r region may be significantly reduced compared to what it is in air.

[00212] If the fraction of magnetic energy in the magnetic material is denoted by *frac*, and the loss tangent of the material is *tanδ*, then the *Q* of the resonator, assuming the magnetic material is the only source of losses, is *Q=1/(frac x tanδ)*. Thus, even for loss tangents as high as 0.1, it is possible to achieve *Q's* of the order of 1,000 for these types of resonator structures.

[00213] If the structure is driven with *N* turns of wire wound around it, the losses in the excitation inductor loop can be ignored if *N* is sufficiently high. Fig. 17 shows an equivalent circuit 1700 schematic for these structures and the scaling of the loss mechanisms and inductance with the number of turns, *N,* wound around a structure made of conducting and magnetic material. If proximity effects can be neglected (by using an appropriate winding, or a wire designed to minimize proximity effects, such as Litz wire and the like), the resistance 1702 due to the wire in the looped conductor scales linearly with the length of the loop, which is in turn proportional to the number of turns. On the other hand, both the equivalent resistance 1708 and equivalent inductance 1704 of these special structures are proportional to the square of the magnetic field inside the structure. Since this magnetic field is proportional to *N*, the equivalent resistance 1708 and equivalent inductance 1704 are both proportional to N^2. Thus, for large enough N, the resistance 1702 of the wire is much smaller than the equivalent resistance 1708 of the magnetic structure, and the Q of the resonator asymptotes to $Q_{max} = \omega L_\mu / R_\mu$.

[00214] Fig. 16 (a) shows a drawing of a copper and magnetic material structure 1602 driven by a square loop of current around the narrowed segment at the center of the structure 1604 and the magnetic field streamlines generated by this structure 1608. This exemplary structure includes two 20 cm x 8 cm x 2 cm hollow regions enclosed with copper and then completely covered with a 2 mm layer of magnetic material having the properties $\mu_r' = 1{,}400$, $\mu_r'' = 5$, and $\sigma = 0.5$ S/m. These two parallelepipeds are spaced 4 cm apart and are connected by a 2 cm x 4 cm x 2 cm block of the same magnetic material. The excitation loop is wound around the center of this block. At a frequency of 300 kHz, this structure has a calculated Q of 890. The conductor and magnetic material structure may be shaped to optimize certain system parameters. For example, the size of the structure enclosed by the excitation loop may be small to reduce the resistance of the excitation loop, or it may be large to mitigate losses in the magnetic material associated with large magnetic fields. Note that the magnetic streamlines and *Q's* associated with the same structure composed of magnetic material only would be similar to the layer conductor and magnetic material design shown here.

[00215] **<u>Electromagnetic Resonators Interacting with Other Objects</u>**

[00216] For electromagnetic resonators, extrinsic loss mechanisms that perturb the intrinsic Q may include absorption losses inside the materials of nearby extraneous objects and radiation losses related to scattering of the resonant fields from nearby extraneous objects. Absorption losses may be associated with materials that, over the frequency range of interest, have non-zero, but finite, conductivity, σ, (or equivalently a non-zero and finite imaginary part of the dielectric permittivity), such that electromagnetic fields can penetrate it and induce currents in it, which then dissipate energy through resistive losses. An object may be described as lossy if it at least partly includes lossy materials.

[00217] Consider an object including a homogeneous isotropic material of conductivity, σ and magnetic permeability, μ. The penetration depth of electromagnetic fields inside this object is given by the skin depth, $\delta = \sqrt{2/\omega\mu\sigma}$. The power dissipated inside the object,

P_d, can be determined from $P_d = \int_V d\mathbf{r}\sigma\,|\mathbf{E}|^2 = \int_V d\mathbf{r}\,|\mathbf{J}|^2/\sigma$ where we made use of Ohm's law, $\mathbf{J} = \sigma\mathbf{E}$, and where $\mathbf{E}$ is the electric field and $\mathbf{J}$ is the current density.

[00218] If over the frequency range of interest, the conductivity, σ, of the material that composes the object is low enough that the material's skin depth, δ, may be considered long, (i.e. δ is longer than the objects' characteristic size, or δ is longer than the characteristic size of the portion of the object that is lossy) then the electromagnetic fields, $\mathbf{E}$ and $\mathbf{H}$, where $\mathbf{H}$ is the magnetic field, may penetrate significantly into the object. Then, these finite-valued fields may give rise to a dissipated power that scales as $P_d \sim \sigma V_{ol} \left\langle |\mathbf{E}|^2 \right\rangle$, where V_{ol} is the volume of the object that is lossy and $\left\langle |\mathbf{E}|^2 \right\rangle$ is the spatial average of the electric-field squared, in the volume under consideration. Therefore, in the low-conductivity limit, the dissipated power scales proportionally to the conductivity and goes to zero in the limit of a non-conducting (purely dielectric) material.

[00219] If over the frequency range of interest, the conductivity, σ, of the material that composes the object is high enough that the material's skin depth may be considered short, then the electromagnetic fields, $\mathbf{E}$ and $\mathbf{H}$, may penetrate only a short distance into the object (namely they stay close to the 'skin' of the material, where δ is smaller than the characteristic thickness of the portion of the object that is lossy). In this case, the currents induced inside the material may be concentrated very close to the material surface, approximately within a skin depth, and their magnitude may be approximated by the product of a surface current density (mostly determined by the shape of the incident electromagnetic fields and, as long as the thickness of the conductor is much larger than the skin-depth, independent of frequency and conductivity to first order) $K(x, y)$ (where x and y are coordinates parameterizing the surface) and a function decaying exponentially into the surface: $\exp(-z/\delta)/\delta$ (where z denotes the coordinate locally normal to the surface): $J(x, y, z) = K(x, y)\exp(-z/\delta)/\delta$. Then, the dissipated power, P_d, may be estimated by,

$$P_d = {}^{V} d\mathrm{r}\,|\mathbf{J}(\mathrm{r})|^2/\sigma \simeq \left({}^{S}\mathbf{dxdy}\,|\mathbf{K}(\mathbf{x},\mathbf{y})|^2\right)\left({}^{\infty}_{\mathbf{0}}\mathbf{dz}\exp(\mathbf{2z}/\delta)/(\sigma\delta^2)\right) = \sqrt{\mu\omega/8\sigma}\left({}^{S} dxdy\,|K(x,y)|^2\right)$$

[00220] Therefore, in the high-conductivity limit, the dissipated power scales inverse proportionally to the square-root of the conductivity and goes to zero in the limit of a perfectly-conducting material.

[00221] If over the frequency range of interest, the conductivity, σ, of the material that composes the object is finite, then the material's skin depth, δ, may penetrate some distance into the object and some amount of power may be dissipated inside the object, depending also on the size of the object and the strength of the electromagnetic fields. This description can be generalized to also describe the general case of an object including multiple different materials with different properties and conductivities, such as an object with an arbitrary inhomogeneous and anisotropic distribution of the conductivity inside the object.

[00222] Note that the magnitude of the loss mechanisms described above may depend on the location and orientation of the extraneous objects relative to the resonator fields as well as the material composition of the extraneous objects. For example, high-conductivity materials may shift the resonant frequency of a resonator and detune it from other resonant objects. This frequency shift may be fixed by applying a feedback mechanism to a resonator that corrects its frequency, such as through changes in the inductance and/or capacitance of the resonator. These changes may be realized using variable capacitors and inductors, in some cases achieved by changes in the geometry of components in the resonators. Other novel tuning mechanisms, described below, may also be used to change the resonator frequency.

[00223] Where external losses are high, the perturbed Q may be low and steps may be taken to limit the absorption of resonator energy inside such extraneous objects and materials. Because of the functional dependence of the dissipated power on the strength of the electric and magnetic fields, one might optimize system performance by designing a system so that the desired coupling is achieved with shorter evanescent resonant field tails at the source resonator and longer at the device resonator, so that the perturbed Q of the source in the presence of other objects is optimized (or vice versa if the perturbed Q of the device needs to be optimized).

[00224] Note that many common extraneous materials and objects such as people, animals, plants, building materials, and the like, may have low conductivities and therefore may have little impact on the wireless energy transfer scheme disclosed here. An important fact related to the magnetic resonator designs we describe is that their electric fields may be confined primarily within the resonator structure itself, so it should be possible to operate within the commonly accepted guidelines for human safety while providing wireless power exchange over mid range distances.

[00225] **<u>Electromagnetic Resonators with Reduced Interactions</u>**

[00226] One frequency range of interest for near-field wireless power transmission is between 10 kHz and 100 MHz. In this frequency range, a large variety of ordinary non-metallic materials, such as for example several types of wood and plastic may have relatively low conductivity, such that only small amounts of power may be dissipated inside them. In addition, materials with low loss tangents, $\tan\Delta$, where $\tan\Delta = \varepsilon'' / \varepsilon'$, and ε'' and ε' are the imaginary and real parts of the permittivity respectively, may also have only small amounts of power dissipated inside them. Metallic materials, such as copper, silver, gold, and the like, with relatively high conductivity, may also have little power dissipated in them, because electromagnetic fields are not able to significantly penetrate these materials, as discussed earlier. These very high and very low conductivity materials, and low loss tangent materials and objects may have a negligible impact on the losses of a magnetic resonator.

[00227] However, in the frequency range of interest, there are materials and objects such as some electronic circuits and some lower-conductivity metals, which may have moderate (in general inhomogeneous and anisotropic) conductivity, and/or moderate to high loss tangents, and which may have relatively high dissipative losses. Relatively larger amounts of power may be dissipated inside them. These materials and objects may dissipate enough energy to reduce $Q_{(p)}$ by non-trivial amounts, and may be referred to as "lossy objects".

[00228] One way to reduce the impact of lossy materials on the $Q_{(p)}$ of a resonator is to use high-conductivity materials to shape the resonator fields such that they avoid the lossy objects. The process of using high-conductivity materials to tailor electromagnetic fields so that they avoid lossy objects in their vicinity may be understood by visualizing high-conductivity materials as materials that deflect or reshape the fields. This picture is qualitatively correct as long as the thickness of the conductor is larger than the skin-depth because the boundary conditions for electromagnetic fields at the surface of a good conductor force the electric field to be nearly completely perpendicular to, and the magnetic field to be nearly completely tangential to, the conductor surface. Therefore, a perpendicular magnetic field or a tangential electric field will be "deflected away" from the conducting surface. Furthermore, even a tangential magnetic field or a perpendicular electric field may be forced to decrease in magnitude on one side and/or in particular locations of the conducting surface, depending on the relative position of the sources of the fields and the conductive surface.

[00229] As an example, Fig. 18 shows a finite element method (FEM) simulation of two high conductivity surfaces 1802 above and below a lossy dielectric material 1804 in an external, initially uniform, magnetic field of frequency f= 6.78 MHz. The system is azimuthally symmetric around the r=0 axis. In this simulation, the lossy dielectric material 1804 is sandwiched between two conductors 1802, shown as the white lines at approximately z = ±0.01m. In the absence of the conducting surfaces above and below the dielectric disk, the magnetic field (represented by the drawn magnetic field lines) would have remained essentially uniform (field lines straight and parallel with the z-axis), indicating that the magnetic field would have passed straight through the lossy dielectric material. In this case, power would have been dissipated in the lossy dielectric disk. In the presence of conducting surfaces, however, this simulation shows the magnetic field is reshaped. The magnetic field is forced to be tangential to surface of the conductor and so is deflected around those conducting surfaces 1802, minimizing the amount of power that may be dissipated in the lossy dielectric material 1804 behind or between the conducting surfaces. As used herein, an axis of electrical symmetry refers to any axis about which a fixed or time-varying electrical or magnetic field is substantially symmetric during an exchange of energy as disclosed herein.

[00230] A similar effect is observed even if only one conducting surface, above or below, the dielectric disk, is used. If the dielectric disk is thin, the fact that the electric field is essentially zero at the surface, and continuous and smooth close to it, means that the electric field is very low everywhere close to the surface (i.e. within the dielectric disk). A single surface implementation for deflecting resonator fields away from lossy objects may be preferred for applications where one is not allowed to cover both sides of the lossy material or object (e.g. an LCD screen). Note that even a very thin surface of conducting material, on the order of a few skin-depths, may be sufficient (the skin depth in pure copper at 6.78 MHz is ~20 μm, and at 250 kHz is ~100 μm) to significantly improve the $Q_{(p)}$ of a resonator in the presence of lossy materials.

[00231] Lossy extraneous materials and objects may be parts of an apparatus, in which a high-Q resonator is to be integrated. The dissipation of energy in these lossy materials and objects may be reduced by a number of techniques including:

- by positioning the lossy materials and objects away from the resonator, or, in special positions and orientations relative to the resonator.
- by using a high conductivity material or structure to partly or entirely cover lossy materials and objects in the vicinity of a resonator
- by placing a closed surface (such as a sheet or a mesh) of high-conductivity material around a lossy object to completely cover the lossy object and shape the resonator fields such that they avoid the lossy object.
- by placing a surface (such as a sheet or a mesh) of a high-conductivity material around only a portion of a lossy object, such as along the top, the bottom, along the side, and the like, of an object or material.
- by placing even a single surface (such as a sheet or a mesh) of high-conductivity material above or below or on one side of a lossy object to reduce the strength of the fields at the location of the lossy object.

[00232] Fig. 19 shows a capacitively-loaded loop inductor forming a magnetic resonator 102 and a disk-shaped surface of high-conductivity material 1802 that completely surrounds a lossy object 1804 placed inside the loop inductor. Note that some lossy objects may be components, such as electronic circuits, that may need to interact with, communicate with, or be connected to the outside environment and thus cannot be completely electromagnetically isolated. Partially covering a lossy material with high conductivity materials may still reduce extraneous losses while enabling the lossy material or object to function properly.

[00233] Fig. 20 shows a capacitively-loaded loop inductor that is used as the resonator 102 and a surface of high-conductivity material 1802, surrounding only a portion of a lossy object 1804, that is placed inside the inductor loop.

[00234] Extraneous losses may be reduced, but may not be completely eliminated, by placing a single surface of high-conductivity material above, below, on the side, and the like, of a lossy object or material. An example is shown in Fig. 21, where a capacitively-loaded loop inductor is used as the resonator 102 and a surface of high-conductivity material 1802 is placed inside the inductor loop under a lossy object 1804 to reduce the strength of the fields at the location of the lossy object. It may be preferable to cover only one side of a material or object because of considerations of cost, weight, assembly complications, air flow, visual access, physical access, and the like.

[00235] A single surface of high-conductivity material may be used to avoid objects that cannot or should not be covered from both sides (e.g. LCD or plasma screens). Such lossy objects may be avoided using optically transparent conductors. High-conductivity optically opaque materials may instead be placed on only a portion of the lossy object, instead of, or in addition to, optically transparent conductors. The adequacy of single-sided vs. multi-sided covering implementations, and the design trade-offs inherent therein may depend on the details of the wireless energy transfer scenario and the properties of the lossy materials and objects.

[00236] Below we describe an example using high-conductivity surfaces to improve the Q-insensitivity, $\Theta_{(p)}$, of an integrated magnetic resonator used in a wireless energy-transfer system. Fig. 22 shows a wireless projector 2200. The wireless projector may include a device resonator 102C, a projector 2202, a wireless network/video adapter 2204, and power conversion circuits 2208, arranged as shown. The device resonator 102C may include a three-turn conductor loop, arranged to enclose a surface, and a capacitor network 2210. The conductor loop may be designed so that the device resonator 102C has a high Q (e.g., >100) at its operating resonant frequency. Prior to integration in the completely wireless projector 2200, this device resonator 102C has a Q of approximately 477 at the designed operating resonant frequency of 6.78 MHz. Upon integration, and placing the wireless network/video adapter card 2204 in the center of the resonator loop inductor, the resonator $Q_{(integrated)}$ was decreased to approximately 347. At least some of the reduction from Q to $Q_{(integrated)}$ was attributed to losses in the perturbing wireless network/video adapter card. As described above, electromagnetic fields associated with the magnetic resonator 102C may induce currents in and on the wireless network/video adapter card 2204, which may be dissipated in resistive losses in the lossy materials that compose the card. We observed that $Q_{(integrated)}$ of the resonator may be impacted differently depending on the composition, position, and orientation, of objects and materials placed in its vicinity.

[00237] In a completely wireless projector example, covering the network/video adapter card with a thin copper pocket (a folded sheet of copper that covered the top and the bottom of the wireless network/video adapter card, but not the communication antenna) improved the $Q_{(integrated)}$ of the magnetic resonator to a $Q_{(integrated\ +\ copper\ pocket)}$ of approximately 444. In other words, most of the reduction in $Q_{(integrated)}$ due to the perturbation caused by the extraneous network/video adapter card could be eliminated using a copper pocket to deflect the resonator fields away from the lossy materials.

[00238] In another completely wireless projector example, covering the network/video adapter card with a single copper sheet placed beneath the card provided a $Q_{(integrated + copper\ sheet)}$ approximately equal to $Q_{(integrated + copper\ pocket)}$. In that example, the high perturbed Q of the system could be maintained with a single high-conductivity sheet used to deflect the resonator fields away from the lossy adapter card.

[00239] It may be advantageous to position or orient lossy materials or objects, which are part of an apparatus including a high-Q electromagnetic resonator, in places where the fields produced by the resonator are relatively weak, so that little or no power may be dissipated in these objects and so that the Q-insensitivity, $\Theta_{(p)}$, may be large. As was shown earlier, materials of different conductivity may respond differently to electric versus magnetic fields. Therefore, according to the conductivity of the extraneous object, the positioning technique may be specialized to one or the other field.

[00240] Fig. 23 shows the magnitude of the electric 2312 and magnetic fields 2314 along a line that contains the diameter of the circular loop inductor and the electric 2318 and magnetic fields 2320 along the axis of the loop inductor for a capacitively-loaded circular loop inductor of wire of radius 30 cm, resonant at 10 MHz. It can be seen that the amplitude of the resonant near-fields reach their maxima close to the wire and decay away from the loop, 2312 , 2314 . In the plane of the loop inductor 2318, 2320, the fields reach a local minimum at the center of the loop. Therefore, given the finite size of the apparatus, it may be that the fields are weakest at the extrema of the apparatus or it may be that the field magnitudes have local minima somewhere within the apparatus. This argument holds for any other type of electromagnetic resonator 102 and any type of apparatus. Examples are shown in Figs. 24a and 24b, where a capacitively-loaded inductor loop forms a magnetic resonator 102 and an extraneous lossy object 1804 is positioned where the electromagnetic fields have minimum magnitude.

[00241] In a demonstration example, a magnetic resonator was formed using a three-turn conductor loop, arranged to enclose a square surface (with rounded corners), and a capacitor network. The Q of the resonator was approximately 619 at the designed operating resonant frequency of 6.78 MHz. The perturbed Q of this resonator depended on the placement of the perturbing object, in this case a pocket projector, relative to the resonator. When the perturbing projector was located inside the inductor loop and at its center or on top of the inductor wire turns, $Q_{(projector)}$ was approximately 96, lower than when the perturbing projector was placed outside of the resonator, in which case $Q_{(projector)}$ was approximately 513. These measurements support the analysis that shows the fields inside the inductor loop may be larger than those outside it, so lossy objects placed inside such a loop inductor may yield lower perturbed *Q's* for the system than when the lossy object is placed outside the loop inductor. Depending on the resonator designs and the material composition and orientation of the lossy object, the arrangement shown in Fig. 24b may yield a higher Q-insensitivity, $\Theta_{(projector)}$, than the arrangement shown in Fig. 24a.

[00242] High-Q resonators may be integrated inside an apparatus. Extraneous materials and objects of high dielectric permittivity, magnetic permeability, or electric conductivity may be part of the apparatus into which a high-Q resonator is to be integrated. For these extraneous materials and objects in the vicinity of a high-Q electromagnetic resonator, depending on their size, position and orientation relative to the resonator, the resonator field-profile may be distorted and deviate significantly from the original unperturbed field-profile of the resonator. Such a distortion of the unperturbed fields of the resonator may significantly decrease the Q to a lower $Q_{(p)}$, even if the extraneous objects and materials are lossless.

[00243] It may be advantageous to position high-conductivity objects, which are part of an apparatus including a high-Q electromagnetic resonator, at orientations such that the surfaces of these objects are, as much as possible, perpendicular to the electric field lines produced by the unperturbed resonator and parallel to the magnetic field lines produced by the unperturbed resonator, thus distorting the resonant field profiles by the smallest amount possible. Other common objects that may be positioned perpendicular to the plane of a magnetic resonator loop include screens (LCD, plasma, etc), batteries, cases, connectors, radiative antennas, and the like. The Q-insensitivity, $\Theta_{(p)}$, of the resonator may be much larger than if the objects were positioned at a different orientation with respect to the resonator fields.

[00244] Lossy extraneous materials and objects, which are not part of the integrated apparatus including a high-Q resonator, may be located or brought in the vicinity of the resonator, for example, during the use of the apparatus. It may be advantageous in certain circumstances to use high conductivity materials to tailor the resonator fields so that they avoid the regions where lossy extraneous objects may be located or introduced to reduce power dissipation in these materials and objects and to increase Q-insensitivity, $\Theta_{(p)}$. An example is shown in Fig. 25, where a capacitively-loaded loop inductor and capacitor are used as the resonator 102 and a surface of high-conductivity material 1802 is placed above the inductor loop to reduce the magnitude of the fields in the region above the resonator, where lossy extraneous objects 1804 may be located or introduced.

[00245] Note that a high-conductivity surface brought in the vicinity of a resonator to reshape the fields may also lead to $Q_{(cond.\ surface)}<Q$. The reduction in the perturbed Q may be due to the dissipation of energy inside the lossy conductor or to the distortion of the unperturbed resonator field profiles associated with matching the field boundary conditions at the surface of the conductor. Therefore, while a high-conductivity surface may be used to reduce the extraneous losses due to dissipation inside an extraneous lossy object, in some cases, especially in some of those where this is achieved by significantly reshaping the electromagnetic fields, using such a high-conductivity surface so that the fields avoid the lossy object may result effectively in $Q_{(p\ +\ cond.\ surface)}<Q_{(p)}$ rather than the desired result $Q_{(p\ +\ cond.\ surface)}>Q_{(p)}$.

[00246] As described above, in the presence of loss inducing objects, the perturbed quality factor of a magnetic resonator may be improved if the electromagnetic fields associated with the magnetic resonator are reshaped to avoid the loss inducing objects. Another way to reshape the unperturbed resonator fields is to use high permeability materials to completely or partially enclose or cover the loss inducing objects, thereby reducing the interaction of the magnetic field with the loss inducing objects.

[00247] Magnetic field shielding has been described previously, for example in *Electrodynamics 3rd Ed.*, Jackson, pp. 201-203. There, a spherical shell of magnetically permeable material was shown to shield its interior from external magnetic fields. For example, if a shell of inner radius a, outer radius b, and relative permeability μ_r, is placed in an initially uniform magnetic field H_0, then the field inside the shell will have a constant magnitude, $9\mu_r H_0 \big/ \left[(2\mu_r+1)(\mu_r+2)-2(a/b)^3(\mu_r-1)^2\right]$, which tends to $9H_0 \big/ 2\mu_r\left(1-(a/b)^3\right)$ if $\mu_r>>1$. This result shows that an incident magnetic field (but not necessarily an incident electric field) may be greatly attenuated inside the shell, even if the shell is quite thin, provided the magnetic permeability is high enough. It may be advantageous in certain circumstances to use high permeability materials to partly or entirely cover lossy materials and objects so that they are avoided by the resonator magnetic fields and so that little or no power is dissipated in these materials and objects. In such an approach, the Q-insensitivity, $\Theta_{(p)}$, may be larger than if the materials and objects were not covered, possibly larger than 1.

[00248] It may be desirable to keep both the electric and magnetic fields away from loss inducing objects. As described above, one way to shape the fields in such a manner is to use high-conductivity surfaces to either completely or partially enclose or cover the loss inducing objects. A layer of magnetically permeable material, also referred to as magnetic material, (any material or meta-material having a non-trivial magnetic permeability), may be placed on or around the high-conductivity surfaces. The additional layer of magnetic material may present a lower reluctance path (compared to free space) for the deflected magnetic field to follow and may partially shield the electric conductor underneath it from the incident magnetic flux. This arrangement may reduce the losses due to induced currents in the high-conductivity surface. Under some circumstances the lower reluctance path presented by the magnetic material may improve the perturbed Q of the structure.

[00249] Fig. 26a shows an axially symmetric FEM simulation of a thin conducting 2604 (copper) disk (20 cm in diameter, 2 cm in height) exposed to an initially uniform, externally applied magnetic field (gray flux lines) along the z-axis. The axis of symmetry is at r=0. The magnetic streamlines shown originate at $z=-\infty$, where they are spaced from r=3 cm to r=10 cm in intervals of 1 cm. The axes scales are in meters. Imagine, for example, that this conducing cylinder encloses loss-inducing objects within an area circumscribed by a magnetic resonator in a wireless energy transfer system such as shown in Fig. 19.

[00250] This high-conductivity enclosure may increase the perturbing Q of the lossy objects and therefore the overall perturbed Q of the system, but the perturbed Q may still be less than the unperturbed Q because of induced losses in the conducting surface and changes to the profile of the electromagnetic fields. Decreases in the perturbed Q associated with the high-conductivity enclosure may be at least partially recovered by including a layer of magnetic material along the outer surface or surfaces of the high-conductivity enclosure. Fig. 26b shows an axially symmetric FEM simulation of the thin conducting 2604A (copper) disk (20 cm in diameter, 2 cm in height) from Fig. 26a, but with an additional layer of magnetic material placed directly on the outer surface of the high-conductivity enclosure. Note that the presence of the magnetic material may provide a lower reluctance path for the magnetic field, thereby at least partially shielding the underlying conductor and reducing losses due to induced eddy currents in the conductor.

[00251] Fig. 27 depicts a variation (in axi-symmetric view) to the system shown in Fig. 26 where not all of the lossy material 2708 may be covered by a high-conductivity surface 2706. In certain circumstances it may be useful to cover only one side of a material or object, such as due to considerations of cost, weight, assembly complications, air flow, visual access, physical access, and the like. In the exemplary arrangement shown in Fig. 27, only one surface of the lossy material 2708 is covered and the resonator inductor loop is placed on the opposite side of the high-conductivity surface.

[00252] Mathematical models were used to simulate a high-conductivity enclosure made of copper and shaped like a 20 cm diameter by 2 cm high cylindrical disk placed within an area circumscribed by a magnetic resonator whose inductive element was a single-turn wire loop with loop radius *r*=11 cm and wire radius *a*= 1 mm. Simulations for an applied 6.78 MHz electromagnetic field suggest that the perturbing quality factor of this high-conductivity enclosure, $\delta Q_{(enclosure)}$, is 1,870. When the high-conductivity enclosure was modified to include a 0.25 cm-thick layer of magnetic material with real relative permeability, $\mu_r' = 40$, and imaginary relative permeability, $\mu_r'' = 10^{-2}$, simulations suggest the perturbing quality factor is increased to $\delta Q_{(enclosure+magnetic\ \ material)}$=5,060.

[00253] The improvement in performance due to the addition of thin layers of magnetic material 2702 may be even more dramatic if the high-conductivity enclosure fills a larger portion of the area circumscribed by the resonator's loop inductor 2704. In the example above, if the radius of the inductor loop 2704 is reduced so that it is only 3 mm away from the surface of the high-conductivity enclosure, the perturbing quality factor may be improved from 670 (conducting enclosure only) to 2,730 (conducting enclosure with a thin layer of magnetic material) by the addition of a thin layer of magnetic material 2702around the outside of the enclosure.

[00254] The resonator structure may be designed to have highly confined electric fields, using shielding, or distributed capacitors, for example, which may yield high, even when the resonator is very close to materials that would typically induce loss.

[00255] **<u>Coupled Electromagnetic Resonators</u>**

[00256] The efficiency of energy transfer between two resonators may be determined by the strong-coupling figure-of-merit, $U = \kappa / \sqrt{\Gamma_s \Gamma_d} = \left(2\kappa / \sqrt{\omega_s \omega_d}\right) \sqrt{Q_s Q_d}$. In magnetic resonator implementations the coupling factor between the two resonators may be related to the inductance of the inductive elements in each of the resonators, L_1 and L_2, and the mutual inductance, M, between them by $\kappa_{12} = \omega M / 2\sqrt{L_1 L_2}$. Note that this expression assumes there is negligible coupling through electric-dipole coupling. For capacitively-loaded inductor loop resonators where the inductor loops are formed by circular conducting loops with N turns, separated by a distance D, and oriented as shown in Fig. 1(b), the mutual inductance is $M = \pi / 4 \cdot \mu_o N_1 N_2 (x_1 x_2)^2 / D^3$ where x_1 , N_1 and x_2 , N_2 are the characteristic size and number of turns of the conductor loop of the first and second resonators respectively. Note that this is a quasi-static result, and so assumes that the resonator's size is much smaller than the wavelength and the resonators' distance is much smaller than the wavelength, but also that their distance is at least a few times their size. For these circular resonators operated in the quasi-static limit and at mid-range distances, as described above, $k = 2\kappa / \sqrt{\omega_1 \omega_2} \sim \left(\sqrt{x_1 x_2} / D\right)^3$. Strong coupling (a large U) between resonators at mid-range distances may be established when the quality factors of the resonators are large enough to compensate for the small k at mid-range distances

[00257] For electromagnetic resonators, if the two resonators include conducting parts, the coupling mechanism may be that currents are induced on one resonator due to electric and magnetic fields generated from the other. The coupling factor may be proportional to the flux of the magnetic field produced from the high-Q inductive element in one resonator crossing a closed area of the high-Q inductive element of the second resonator.

[00258] **Coupled Electromagnetic Resonators with Reduced Interactions**

[00259] As described earlier, a high-conductivity material surface may be used to shape resonator fields such that they avoid lossy objects, p, in the vicinity of a resonator, thereby reducing the overall extraneous losses and maintaining a high Q-insensitivity $\Theta_{(p\,+\,cond.\,surface)}$ of the resonator. However, such a surface may also lead to a perturbed coupling factor, $k_{(p\,+\,cond.\,surface)}$, between resonators that is smaller than the perturbed coupling factor, $k_{(p)}$ and depends on the size, position, and orientation of the high-conductivity material relative to the resonators. For example, if high-conductivity materials are placed in the plane and within the area circumscribed by the inductive element of at least one of the magnetic resonators in a wireless energy transfer system, some of the magnetic flux through the area of the resonator, mediating the coupling, may be blocked and k may be reduced.

[00260] Consider again the example of Fig. 19. In the absence of the high-conductivity disk enclosure, a certain amount of the external magnetic flux may cross the circumscribed area of the loop. In the presence of the high-conductivity disk enclosure, some of this magnetic flux may be deflected or blocked and may no longer cross the area of the loop, thus leading to a smaller perturbed coupling factor $k_{12(p + cond.\ surfaces)}$. However, because the deflected magnetic-field lines may follow the edges of the high-conductivity surfaces closely, the reduction in the flux through the loop circumscribing the disk may be less than the ratio of the areas of the face of the disk to the area of the loop.

[00261] One may use high-conductivity material structures, either alone, or combined with magnetic materials to optimize perturbed quality factors, perturbed coupling factors, or perturbed efficiencies.

[00262] Consider the example of Fig. 21. Let the lossy object have a size equal to the size of the capacitively-loaded inductor loop resonator, thus filling its area A 2102. A high-conductivity surface 1802 may be placed under the lossy object 1804. Let this be resonator 1 in a system of two coupled resonators 1 and 2, and let us consider how $U_{12(object + cond.\ surface)}$ scales compared to U_{12} as the area A_s 2104 of the conducting surface increases. Without the conducting surface 1802 below the lossy object 1804, the k-insensitivity, $\beta_{12(object)}$, may be approximately one, but the Q-insensitivity, $\Theta_{1(object)}$, may be small, so the U-insensitivity $\Xi_{12(object)}$ may be small.

[00263] Where the high-conductivity surface below the lossy object covers the entire area of the inductor loop resonator (A_s=A), $k_{12(object + cond.\ surface)}$ may approach zero, because little flux is allowed to cross the inductor loop, so $U_{12(object + cond.\ surface)}$ may approach zero. For intermediate sizes of the high-conductivity surface, the suppression of extrinsic losses and the associated Q-insensitivity, $\Theta_{1(object + cond.\ surface)}$, may be large enough compared to $\Theta_{1(object)}$, while the reduction in coupling may not be significant and the associated k-insensitivity, $\beta_{12(object + cond.\ surface)}$, may be not much smaller than $\beta_{12(object)}$, so that the overall $U_{12(object + cond.\ surface)}$ may be increased compared to $U_{12(object)}$. The optimal degree of avoiding of extraneous lossy objects via high-conductivity surfaces in a system of wireless energy transfer may depend on the details of the system configuration and the application.

[00264] We describe using high-conductivity materials to either completely or partially enclose or cover loss inducing objects in the vicinity of high-Q resonators as one potential method to achieve high perturbed Q's for a system. However, using a good conductor alone to cover the objects may reduce the coupling of the resonators as described above, thereby reducing the efficiency of wireless power transfer. As the area of the conducting surface approaches the area of the magnetic resonator, for example, the perturbed coupling factor, $k_{(p)}$, may approach zero, making the use of the conducting surface incompatible with efficient wireless power transfer.

[00265] One approach to addressing the aforementioned problem is to place a layer of magnetic material around the high-conductivity materials because the additional layer of permeable material may present a lower reluctance path (compared to free space) for the deflected magnetic field to follow and may partially shield the electric conductor underneath it from incident magnetic flux. Under some circumstances the lower reluctance path presented by the magnetic material may improve the electromagnetic coupling of the resonator to other resonators. Decreases in the perturbed coupling factor associated with using conducting materials to tailor resonator fields so that they avoid lossy objects in and around high-Q magnetic resonators may be at least partially recovered by including a layer of magnetic material along the outer surface or surfaces of the conducting materials. The magnetic materials may increase the perturbed coupling factor relative to its initial unperturbed value.

[00266] Note that the simulation results in Fig. 26 show that an incident magnetic field may be deflected less by a layered magnetic material and conducting structure than by a conducting structure alone. If a magnetic resonator loop with a radius only slightly larger than that of the disks shown in Figs. 26(a) and 26(b) circumscribed the disks, it is clear that more flux lines would be captured in the case illustrated in Fig. 26(b) than in Fig. 26(a), and therefore $k_{(disk)}$ would be larger for the case illustrated in Fig. 26(b). Therefore, including a layer of magnetic material on the conducting material may improve the overall system performance. System analyses may be performed to determine whether these materials should be partially, totally, or minimally integrated into the resonator.

[00267] As described above, Fig. 27 depicts a layered conductor 2706 and magnetic material 2702 structure that may be appropriate for use when not all of a lossy material 2708 may be covered by a conductor and/or magnetic material structure. It was shown earlier that for a copper conductor disk with a 20 cm diameter and a 2 cm height, circumscribed by a resonator with an inductor loop radius of 11 cm and a wire radius a=1 mm, the calculated perturbing Q for the copper cylinder was 1,870. If the resonator and the conducting disk shell are placed in a uniform magnetic field (aligned along the axis of symmetry of the inductor loop), we calculate that the copper conductor has an associated coupling factor insensitivity of 0.34. For comparison, we model the same arrangement but include a 0.25 cm-thick layer of magnetic material with a real relative permeability, $\mu_r' = 40$, and an imaginary relative permeability, $\mu_r'' = 10^{-2}$. Using the same model and parameters described above, we find that the coupling factor insensitivity is improved to 0.64 by the addition of the magnetic material to the surface of the conductor.

[00268] Magnetic materials may be placed within the area circumscribed by the magnetic resonator to increase the coupling in wireless energy transfer systems. Consider a solid sphere of a magnetic material with relative permeability, μ_r, placed in an initially uniform magnetic field. In this example, the lower reluctance path offered by the magnetic material may cause the magnetic field to concentrate in the volume of the sphere. We find that the magnetic flux through the area circumscribed by the equator of the sphere is enhanced by a factor of $3\mu_r/(\mu_r+2)$, by the addition of the magnetic material. If $\mu_r >> 1$, this enhancement factor may be close to 3.

[00269] One can also show that the dipole moment of a system comprising the magnetic sphere circumscribed by the inductive element in a magnetic resonator would have its magnetic dipole enhanced by the same factor. Thus, the magnetic sphere with high permeability practically triples the dipole magnetic coupling of the resonator. It is possible to keep most of this increase in coupling if we use a spherical shell of magnetic material with inner radius *a,* and outer radius *b,* even if this shell is on top of block or enclosure made from highly conducting materials. In this case, the enhancement in the flux through the equator is

$$\frac{3\mu_r\left(1-\left(\frac{a}{b}\right)^3\right)}{\mu_r\left(1-\left(\frac{a}{b}\right)^3\right)+2\left(1+\frac{1}{2}\left(\frac{a}{b}\right)^3\right)}.$$

For μ_r=1,000 and (a/b)=0.99, this enhancement factor is still 2.73, so it possible to significantly improve the coupling even with thin layers of magnetic material.

[00270] As described above, structures containing magnetic materials may be used to realize magnetic resonators. Fig. 16(a) shows a 3 dimensional model of a copper and magnetic material structure 1600 driven by a square loop of current around the choke point at its center. Fig. 16(b) shows the interaction, indicated by magnetic field streamlines , between two identical structures 1600A-B with the same properties as the one shown in Fig. 16(a). Because of symmetry, and to reduce computational complexity, only one half of the system is modeled. If we fix the relative orientation between the two objects and vary their center-to-center distance (the image shown is at a relative separation of 50 cm), we find that, at 300 kHz, the coupling efficiency varies from 87% to 55% as the separation between the structures varies from 30 cm to 60 cm. Each of the example structures shown 1600 A-B includes two 20 cm x 8 cm x 2cm parallelepipeds made of copper joined by a 4 cm x 4 cm x 2 cm block of magnetic material and entirely covered with a 2 mm layer of the same magnetic material (assumed to have μ_r=1,400+j5). Resistive losses in the driving loop are ignored. Each structure has a calculated Q of 815.

[00271] <u>ELECTROMAGNETIC RESONATORS AND IMPEDANCE MATCHING</u>

[00272] <u>Impedance Matching Architectures for Low-Loss Inductive Elements</u>

[00273] For purposes of the present discussion, an inductive element may be any coil or loop structure (the 'loop') of any conducting material, with or without a (gapped or ungapped) core made of magnetic material, which may also be coupled inductively or in any other contactless way to other systems. The element is inductive because its impedance, including both the impedance of the loop and the so-called 'reflected' impedances of any potentially coupled systems, has positive reactance, X, and resistance, R.

[00274] Consider an external circuit, such as a driving circuit or a driven load or a transmission line, to which an inductive element may be connected. The external circuit (e.g. a driving circuit) may be delivering power to the inductive element and the inductive element may be delivering power to the external circuit (e.g. a driven load). The efficiency and amount of power delivered between the inductive element and the external circuit at a desired frequency may depend on the impedance of the inductive element relative to the properties of the external circuit. Impedance-matching networks and external circuit control techniques may be used to regulate the power delivery between the external circuit and the inductive element, at a desired frequency, f.

[00275] The external circuit may be a driving circuit configured to form a amplifier of class A, B, C, D, DE, E, F and the like, and may deliver power at maximum efficiency (namely with minimum losses within the driving circuit) when it is driving a resonant network with specific impedance Z_o^*, where Z_o may be complex and * denotes complex conjugation. The external circuit may be a driven load configured to form a rectifier of class A, B, C, D, DE, E, F and the like, and may receive power at maximum efficiency (namely with minimum losses within the driven load) when it is driven by a resonant network with specific impedance Z_o^*, where Z_o may be complex. The external circuit may be a transmission line with characteristic impedance, Z_o, and may exchange power at maximum efficiency (namely with zero reflections) when connected to an impedance Z_o^*. We will call the characteristic impedance Z_o of an external circuit the complex conjugate of the impedance that may be connected to it for power exchange at maximum efficiency.

[00276] Typically the impedance of an inductive element, $R+jX$, may be much different from Z_o^*. For example, if the inductive element has low loss (a high X/R), its resistance, R, may be much lower than the real part of the characteristic impedance, Z_0, of the external circuit. Furthermore, an inductive element by itself may not be a resonant network. An impedance-matching network connected to an inductive element may typically create a resonant network, whose impedance may be regulated.

[00277] Therefore, an impedance-matching network may be designed to maximize the efficiency of the power delivered between the external circuit and the inductive element (including the reflected impedances of any coupled systems). The efficiency of delivered power may be maximized by matching the impedance of the combination of an impedance-matching network and an inductive element to the characteristic impedance of an external circuit (or transmission line) at the desired frequency.

[00278] An impedance-matching network may be designed to deliver a specified amount of power between the external circuit and the inductive element (including the reflected impedances of any coupled systems). The delivered power may be determined by adjusting the complex ratio of the impedance of the combination of the impedance-matching network and the inductive element to the impedance of the external circuit (or transmission line) at the desired frequency.

[00279] Impedance-matching networks connected to inductive elements may create magnetic resonators. For some applications, such as wireless power transmission using strongly-coupled magnetic resonators, a high Q may be desired for the resonators. Therefore, the inductive element may be chosen to have low losses (high X/R).

[00280] Since the matching circuit may typically include additional sources of loss inside the resonator, the components of the matching circuit may also be chosen to have low losses. Furthermore, in high-power applications and/or due to the high resonator Q, large currents may run in parts of the resonator circuit and large voltages may be present across some circuit elements within the resonator. Such currents and voltages may exceed the specified tolerances for particular circuit elements and may be too high for particular components to withstand. In some cases, it may be difficult to find or implement components, such as tunable capacitors for example, with size, cost and performance (loss and current/voltage-rating) specifications sufficient to realize high-Q and high-power resonator designs for certain applications. We disclose matching circuit designs, methods, implementations and techniques that may preserve the high Q for magnetic resonators, while reducing the component requirements for low loss and/or high current/voltage-rating.

[00281] Matching-circuit topologies may be designed that minimize the loss and current-rating requirements on some of the elements of the matching circuit. The topology of a circuit matching a low-loss inductive element to an impedance, Z_0, may be chosen so that some of its components lie outside the associated high-Q resonator by being in series with the external circuit. The requirements for low series loss or high current-ratings for these components may be reduced. Relieving the low series loss and/or high-current-rating requirement on a circuit element may be particularly useful when the element needs to be variable and/or to have a large voltage-rating and/or low parallel loss.

[00282] Matching-circuit topologies may be designed that minimize the voltage rating requirements on some of the elements of the matching circuit. The topology of a circuit matching a low-loss inductive element to an impedance, Z_0, may be chosen so that some of its components lie outside the associated high-Q resonator by being in parallel with Z_0. The requirements for low parallel loss or high voltage-rating for these components may be reduced. Relieving the low parallel loss and/or high-voltage requirement on a circuit element may be particularly useful when the element needs to be variable and/or to have a large current-rating and/or low series loss.

[00283] The topology of the circuit matching a low-loss inductive element to an external characteristic impedance, Z_0, may be chosen so that the field pattern of the associated resonant mode and thus its high *Q* are preserved upon coupling of the resonator to the external impedance. Otherwise inefficient coupling to the desired resonant mode may occur (potentially due to coupling to other undesired resonant modes), resulting in an effective lowering of the resonator *Q*.

[00284] For applications where the low-loss inductive element or the external circuit, may exhibit variations, the matching circuit may need to be adjusted dynamically to match the inductive element to the external circuit impedance, Z_0, at the desired frequency, *f*. Since there may typically be two tuning objectives, matching or controlling both the real and imaginary part of the impedance level, Z_0, at the desired frequency, *f*, there may be two variable elements in the matching circuit. For inductive elements, the matching circuit may need to include at least one variable capacitive element.

[00285] A low-loss inductive element may be matched by topologies using two variable capacitors, or two networks of variable capacitors. A variable capacitor may, for example, be a tunable butterfly-type capacitor having, e.g., a center terminal for connection to a ground or other lead of a power source or load, and at least one other terminal across which a capacitance of the tunable butterfly-type capacitor can be varied or tuned, or any other capacitor having a user-configurable, variable capacitance.

[00286] A low-loss inductive element may be matched by topologies using one, or a network of, variable capacitor(s) and one, or a network of, variable inductor(s).

[00287] A low-loss inductive element may be matched by topologies using one, or a network of, variable capacitor(s) and one, or a network of, variable mutual inductance(s), which transformer-couple the inductive element either to an external circuit or to other systems.

[00288] In some cases, it may be difficult to find or implement tunable lumped elements with size, cost and performance specifications sufficient to realize high-*Q*, high-power, and potentially high-speed, tunable resonator designs. The topology of the circuit matching a variable inductive element to an external circuit may be designed so that some of the variability is assigned to the external circuit by varying the frequency, amplitude, phase, waveform, duty cycle, and the like, of the drive signals applied to transistors, diodes, switches and the like, in the external circuit.

[00289] The variations in resistance, R, and inductance, L, of an inductive element at the resonant frequency may be only partially compensated or not compensated at all. Adequate system performance may thus be preserved by tolerances designed into other system components or specifications. Partial adjustments, realized using fewer tunable components or less capable tunable components, may be sufficient.

[00290] Matching-circuit architectures may be designed that achieve the desired variability of the impedance matching circuit under high-power conditions, while minimizing the voltage/current rating requirements on its tunable elements and achieving a finer (i.e. more precise, with higher resolution) overall tunability. The topology of the circuit matching a variable inductive element to an impedance, Z_0, may include appropriate combinations and placements of fixed and variable elements, so that the voltage/current requirements for the variable components may be reduced and the desired tuning range may be covered with finer tuning resolution. The voltage/current requirements may be reduced on components that are not variable.

[00291] The disclosed impedance matching architectures and techniques may be used to achieve the following:

- To maximize the power delivered to, or to minimize impedance mismatches between, the source low-loss inductive elements (and any other systems wirelessly coupled to them) from the power driving generators.
- To maximize the power delivered from, or to minimize impedance mismatches between, the device low-loss inductive elements (and any other systems wirelessly coupled to them) to the power driven loads.
- To deliver a controlled amount of power to, or to achieve a certain impedance relationship between, the source low-loss inductive elements (and any other systems wirelessly coupled to them) from the power driving generators.
- To deliver a controlled amount of power from, or to achieve a certain impedance relationship between, the device low-loss inductive elements (and any other systems wirelessly coupled to them) to the power driven loads.

[00292] TOPOLOGIES FOR PRESERVATION OF MODE PROFILE (HIGH-*Q*)

[00293] The resonator structure may be designed to be connected to the generator or the load wirelessly (indirectly) or with a hard-wired connection (directly).

[00294] Consider a general indirectly coupled matching topology such as that shown by the block diagram in Fig. 28(a). There, an inductive element 2802 , labeled as (*R,L*) and represented by the circuit symbol for an inductor, may be any of the inductive elements discussed in this disclosure or in the references provided herein, and where an impedance-matching circuit 2402 includes or consists of parts A and B. B may be the part of the matching circuit that connects the impedance 2804 , Z_0, to the rest of the circuit (the combination of A and the inductive element (A+(*R,L*)) via a wireless connection (an inductive or capacitive coupling mechanism).

[00295] The combination of A and the inductive element 2802 may form a resonator 102, which in isolation may support a high-*Q* resonator electromagnetic mode, with an associated current and charge distribution. The lack of a wired connection between the external circuit, Z_0 and B, and the resonator, A + (*R,L*), may ensure that the high-*Q* resonator electromagnetic mode and its current/charge distributions may take the form of its intrinsic (in-isolation) profile, so long as the degree of wireless coupling is not too large. That is, the electromagnetic mode, current/charge distributions, and thus the high-*Q* of the resonator may be automatically maintained using an indirectly coupled matching topology.

[00296] This matching topology may be referred to as indirectly coupled, or transformer-coupled, or inductively-coupled, in the case where inductive coupling is used between the external circuit and the inductor loop. This type of coupling scenario was used to couple the power supply to the source resonator and the device resonator to the light bulb in the demonstration of wireless energy transfer over mid-range distances described in the referenced *Science* article.

[00297] Next consider examples in which the inductive element may include the inductive element and any indirectly coupled systems. In this case, as disclosed above, and again because of the lack of a wired connection between the external circuit or the coupled systems and the resonator, the coupled systems may not, with good approximation for not-too-large degree of indirect coupling, affect the resonator electromagnetic mode profile and the current/charge distributions of the resonator. Therefore, an indirectly-coupled matching circuit may work equally well for any general inductive element as part of a resonator as well as for inductive elements wirelessly-coupled to other systems, as defined herein. Throughout this disclosure, the matching topologies we disclose refer to matching topologies for a general inductive element of this type, that is, where any additional systems may be indirectly coupled to the low-loss inductive element, and it is to be understood that those additional systems do not greatly affect the resonator electromagnetic mode profile and the current/charge distributions of the resonator.

[00298] Based on the argument above, in a wireless power transmission system of any number of coupled source resonators, device resonators and intermediate resonators the wireless magnetic (inductive) coupling between resonators does not affect the electromagnetic mode profile and the current/charge distributions of each one of the resonators. Therefore, when these resonators have a high (unloaded and unperturbed) Q, their (unloaded and unperturbed) Q may be preserved in the presence of the wireless coupling. (Note that the loaded Q of a resonator may be reduced in the presence of wireless coupling to another resonator, but we may be interested in preserving the unloaded Q, which relates only to loss mechanisms and not to coupling/loading mechanisms.)

[00299] Consider a matching topology such as is shown in Fig. 28(b). The capacitors shown in Fig. 28(b) may represent capacitor circuits or networks. The capacitors shown may be used to form the resonator 102 and to adjust the frequency and/or impedance of the source and device resonators. This resonator 102 may be directly coupled to an impedance, Z_0, using the ports labeled “terminal connections” 2808. Fig. 28(c) shows a generalized directly coupled matching topology, where the impedance-matching circuit 2602 includes or consists of parts A, B and C. Here, circuit elements in A, B and C may be considered part of the resonator 102 as well as part of the impedance matching 2402 (and frequency tuning) topology. B and C may be the parts of the matching circuit 2402 that connect the impedance Z_0 2804 (or the network terminals) to the rest of the circuit (A and the inductive element) via a single wire connection each. Note that B and C could be empty (short-circuits). If we disconnect or open circuit parts B and C (namely those single wire connections), then, the combination of A and the inductive element (R,L) may form the resonator.

[00300] The high-*Q* resonator electromagnetic mode may be such that the profile of the voltage distribution along the inductive element has nodes, namely positions where the voltage is zero. One node may be approximately at the center of the length of the inductive element, such as the center of the conductor used to form the inductive element, (with or without magnetic materials) and at least one other node may be within A. The voltage distribution may be approximately anti-symmetric along the inductive element with respect to its voltage node. A high *Q* may be maintained by designing the matching topology (A, B, C) and/or the terminal voltages (V1, V2) so that this high-*Q* resonator electromagnetic mode distribution may be approximately preserved on the inductive element. This high-*Q* resonator electromagnetic mode distribution may be approximately preserved on the inductive element by preserving the voltage node (approximately at the center) of the inductive element. Examples that achieve these design goals are provided herein.

[00301] A, B, and C may be arbitrary (namely not having any special symmetry), and V1 and V2 may be chosen so that the voltage across the inductive element is symmetric (voltage node at the center inductive). These results may be achieved using simple matching circuits but potentially complicated terminal voltages, because a topology-dependent common-mode signal (V1+V2)/2 may be required on both terminals.

[00302] Consider an 'axis' that connects all the voltage nodes of the resonator, where again one node is approximately at the center of the length of the inductive element and the others within A. (Note that the 'axis' is really a set of points (the voltage nodes) within the electric-circuit topology and may not necessarily correspond to a linear axis of the actual physical structure. The 'axis' may align with a physical axis in cases where the physical structure has symmetry.) Two points of the resonator are electrically symmetric with respect to the 'axis', if the impedances seen between each of the two points and a point on the 'axis', namely a voltage-node point of the resonator, are the same.

[00303] B and C may be the same (C=B), and the two terminals may be connected to any two points of the resonator (A + (R,L)) that are electrically symmetric with respect to the 'axis' defined above and driven with opposite voltages (V2=-V1) as shown in Fig. 28(d). The two electrically symmetric points of the resonator 102 may be two electrically symmetric points on the inductor loop. The two electrically symmetric points of the resonator may be two electrically symmetric points inside A. If the two electrically symmetric points, (to which each of the equal parts B and C is connected), are inside A, A may need to be designed so that these electrically-symmetric points are accessible as connection points within the circuit. This topology may be referred to as a 'balanced drive' topology. These balanced-drive examples may have the advantage that any common-mode signal that may be present on the ground line, due to perturbations at the external circuitry or the power network, for example, may be automatically rejected (and may not reach the resonator). In some balanced-drive examples, this topology may require more components than other topologies.

[00304] In other examples, C may be chosen to be a short-circuit and the corresponding terminal to be connected to ground (V=0) and to any point on the electric-symmetry (zero-voltage) 'axis' of the resonator, and B to be connected to any other point of the resonator not on the electric-symmetry 'axis', as shown in Fig. 28(e). The ground-connected point on the electric-symmetry 'axis' may be the voltage node on the inductive element, approximately at the center of its conductor length. The ground-connected point on the electric-symmetry 'axis' may be inside the circuit A. Where the ground-connected point on the electric-symmetry 'axis' is inside A, A may need to be designed to include one such point on the electrical-symmetric 'axis' that is electrically accessible, namely where connection is possible.

[00305] This topology may be referred to as an 'unbalanced drive' topology. The approximately anti-symmetric voltage distribution of the electromagnetic mode along the inductive element may be approximately preserved, even though the resonator may not be driven exactly symmetrically. The reason is that the high Q and the large associated R-vs.-Z_0 mismatch necessitate that a small current may run through B and ground, compared to the much larger current that may flow inside the resonator, (A+(R,L)). In this scenario, the perturbation on the resonator mode may be weak and the location of the voltage node may stay at approximately the center location of the inductive element. These unbalanced-drive examples may have the advantage that they may be achieved using simple matching circuits and that there is no restriction on the driving voltage at the V1 terminal. In some unbalanced-drive examples, additional designs may be required to reduce common-mode signals that may appear at the ground terminal.

[00306] The directly-coupled impedance-matching circuit, generally including or consisting of parts A, B and C, as shown in Fig. 28(c), may be designed so that the wires and components of the circuit do not perturb the electric and magnetic field profiles of the electromagnetic mode of the inductive element and/or the resonator and thus preserve the high resonator Q. The wires and metallic components of the circuit may be oriented to be perpendicular to the electric field lines of the electromagnetic mode. The wires and components of the circuit may be placed in regions where the electric and magnetic field of the electromagnetic mode are weak.

[00307] TOPOLOGIES FOR ALLEVIATING LOW-SERIES-LOSS AND HIGH-CURRENT-RATING REQUIREMENTS ON ELEMENTS

[00308] If the matching circuit used to match a small resistance, *R*, of a low-loss inductive element to a larger characteristic impedance, Z_0, of an external circuit may be considered lossless, then $I_{Z_o}^2 Z_o = I_R^2 R \leftrightarrow I_{Z_o} / I_R = \sqrt{R / Z_o}$ and the current flowing through the terminals is much smaller than the current flowing through the inductive element. Therefore, elements connected immediately in series with the terminals (such as in directly-coupled B, C (Fig. 28(c))) may not carry high currents. Then, even if the matching circuit has lossy elements, the resistive loss present in the elements in series with the terminals may not result in a significant reduction in the high-*Q* of the resonator. That is, resistive loss in those series elements may not significantly reduce the efficiency of power transmission from Z_0 to the inductive element or vice versa. Therefore, strict requirements for low-series-loss and/or high current-ratings may not be necessary for these components. In general, such reduced requirements may lead to a wider selection of components that may be designed into the high-*Q* and/or high-power impedance matching and resonator topologies. These reduced requirements may be especially helpful in expanding the variety of variable and/or high voltage and/or low-parallel-loss components that may be used in these high-Q and/or high-power impedance-matching circuits.

[00309] TOPOLOGIES FOR ALLEVIATING LOW-PARALLEL-LOSS AND HIGH-VOLTAGE-RATING REQUIREMENTS ON ELEMENTS

[00310] If, as above, the matching circuit used to match a small resistance, *R*, of a low-loss inductive element to a larger characteristic impedance, Z_0, of an external circuit is lossless, then using the previous analysis,

$$|V_{Z_o} / V_{load}| = | I_{Z_o} Z_o / I_R (R + jX) | \approx \sqrt{R / Z_o} \cdot Z_o / X = \sqrt{Z_o / R} / \left(X / R \right),$$

and, for a low-loss (high-*X/R*) inductive element, the voltage across the terminals may be typically much smaller than the voltage across the inductive element. Therefore, elements connected immediately in parallel to the terminals may not need to withstand high voltages. Then, even if the matching circuit has lossy elements, the resistive loss present in the elements in parallel with the terminals may not result in a significant reduction in the high-*Q* of the resonator. That is, resistive loss in those parallel elements may not significantly reduce the efficiency of power transmission from Z_0 to the inductive element or vice versa. Therefore, strict requirements for low-parallel-loss and/or high voltage-ratings may not be necessary for these components. In general, such reduced requirements may lead to a wider selection of components that may be designed into the high-*Q* and/or high-power impedance matching and resonator topologies. These reduced requirements may be especially helpful in expanding the variety of variable and/or high current and/or low-series-loss components that may be used in these high-*Q* and/or high-power impedance-matching and resonator circuits.

[00311] Note that the design principles above may reduce currents and voltages on various elements differently, as they variously suggest the use of networks in series with Z_0 (such as directly-coupled B, C) or the use of networks in parallel with Z_0. The preferred topology for a given application may depend on the availability of low-series-loss/high-current-rating or low-parallel-loss/high-voltage-rating elements.

[00312] COMBINATIONS OF FIXED AND VARIABLE ELEMENTS FOR ACHIEVING FINE TUNABILITY AND ALLEVIATING HIGH-RATING REQUIREMENTS ON VARIABLE ELEMENTS

[00313] **Circuit topologies**

[00314] Variable circuit elements with satisfactory low-loss and high-voltage or current ratings may be difficult or expensive to obtain. In this disclosure, we describe impedance-matching topologies that may incorporate combinations of fixed and variable elements, such that large voltages or currents may be assigned to fixed elements in the circuit, which may be more likely to have adequate voltage and current ratings, and alleviating the voltage and current rating requirements on the variable elements in the circuit.

[00315] Variable circuit elements may have tuning ranges larger than those required by a given impedance-matching application and, in those cases, fine tuning resolution may be difficult to obtain using only such large-range elements. In this disclosure, we describe impedance-matching topologies that incorporate combinations of both fixed and variable elements, such that finer tuning resolution may be accomplished with the same variable elements.

[00316] Therefore, topologies using combinations of both fixed and variable elements may bring two kinds of advantages simultaneously: reduced voltage across, or current through, sensitive tuning components in the circuit and finer tuning resolution. Note that the maximum achievable tuning range may be related to the maximum reduction in voltage across, or current through, the tunable components in the circuit designs.

[00317] **Element topologies**

[00318] A single variable circuit-element (as opposed to the network of elements discussed above) may be implemented by a topology using a combination of fixed and variable components, connected in series or in parallel, to achieve a reduction in the rating requirements of the variable components and a finer tuning resolution. This can be demonstrated mathematically by the fact that:

$$\text{If } x_{|total|} = x_{|fixed|} + x_{|variable|},$$
$$\text{then } \Delta x_{|total|} / x_{|total|} = \Delta x_{|\text{variable}|} / (x_{|fixed|} + x_{|\text{variable}|}),$$
$$\text{and } X_{\text{variable}} / X_{total} = X_{\text{variable}} / (X_{fixed} + X_{\text{variable}}),$$

where $x_{|\text{subscript}|}$ is any element value (e.g. capacitance, inductance), X is voltage or current, and the "+ sign" denotes the appropriate (series-addition or parallel-addition) combination of elements. Note that the subscript format for $x_{|\text{subscript}|}$, is chosen to easily distinguish it from the radius of the area enclosed by a circular inductive element (e.g. x, x_1, etc.).

[00319] Furthermore, this principle may be used to implement a variable electric element of a certain type (e.g. a capacitance or inductance) by using a variable element of a different type, if the latter is combined appropriately with other fixed elements.

[00320] In conclusion, one may apply a topology optimization algorithm that decides on the required number, placement, type and values of fixed and variable elements with the required tunable range as an optimization constraint and the minimization of the currents and/or voltages on the variable elements as the optimization objective.

[00321] **EXAMPLES**

[00322] In the following schematics, we show different specific topology implementations for impedance matching to and resonator designs for a low-loss inductive element. In addition, we indicate for each topology: which of the principles described above are used, the equations giving the values of the variable elements that may be used to achieve the matching, and the range of the complex impedances that may be matched (using both inequalities and a Smith-chart description). For these examples, we assume that Z_0 is real, but an extension to a characteristic impedance with a non-zero imaginary part is straightforward, as it implies only a small adjustment in the required values of the components of the matching network. We will use the convention that the subscript, *n*, on a quantity implies normalization to (division by) Z_0.

[00323] Fig. 29 shows two examples of a transformer-coupled impedance-matching circuit, where the two tunable elements are a capacitor and the mutual inductance between two inductive elements. If we define respectively $X_2=\omega L_2$ for Fig. 29(a) and $X_2=\omega L_2-1/\omega C_2$ for Fig. 29(b), and $X \equiv \omega L$, then the required values of the tunable elements are:

$$\omega C_1 = \frac{1}{X + RX_{2n}}$$
$$\omega M = \sqrt{Z_o R(1+X_{2n}^2)}\ .$$

For the topology of Fig. 29(b), an especially straightforward design may be to choose $X_2=0$. In that case, these topologies may match the impedances satisfying the inequalities:

$$R_n > 0\,,\ X_n > 0\,,$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 29(c).

[00324] Given a well pre-chosen fixed M, one can also use the above matching topologies with a tunable C_2 instead.

[00325] Fig. 30 shows six examples (a)-(f) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and six examples (h)-(m) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 30(a),(b),(c),(h),(i),(j), a common-mode signal may be required at the two terminals to preserve the voltage node of the resonator at the center of the inductive element and thus the high *Q*. Note that these examples may be described as implementations of the general topology shown in Fig. 28(c). For the symmetric topologies of Figs. 30(d),(e),(f),(k),(l),(m), the two terminals may need to be driven anti-symmetrically (balanced drive) to preserve the voltage node of the resonator at the center of the inductive element and thus the high *Q*. Note that these examples may be described as implementations of the general topology shown in Fig. 28(d). It will be appreciated that a network of capacitors, as used herein, may in general refer to any circuit topology including one or more capacitors, including without limitation any of the circuits specifically disclosed herein using capacitors, or any other equivalent or different circuit structure(s), unless another meaning is explicitly provided or otherwise clear from the context.

[00326] Let us define respectively Z=R+jωL for Figs. 30(a),(d),(h),(k), Z=R+jωL+1/jωC_3 for Figs. 30(b),(e),(i),(l), and Z=(R+jωL)||(1/jωC_3) for Figs. 30(c),(f),(j),(m), where the symbol "||" means "the parallel combination of", and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.30(a)-(f) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X - \sqrt{X^2 R_n - R^2(1-R_n)}}{X^2 + R^2},$$
$$\omega C_2 = \frac{R_n \omega C_1}{1 - X\omega C_1 - R_n},$$

and these topologies can match the impedances satisfying the inequalities:

$$R_n \le 1,\ X_n \ge \sqrt{R_n(1-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 30(g).
For Figs.30(h)-(m) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X + \sqrt{X^2 R_n - R^2(1-R_n)}}{X^2 + R^2},$$
$$\omega L_2 = -\frac{1 - X\omega C_1 - R_n}{R_n \omega C_1}.$$

[00327] Fig. 31 shows three examples (a)-(c) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and three examples (e)-(g) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 31(a),(b),(c),(e),(f),(g), the ground terminal is connected between two equal-value capacitors, 2C_1, (namely on the axis of symmetry of the main resonator) to preserve the voltage node of the resonator at the center of the inductive element and thus the high *Q*. Note that these examples may be described as implementations of the general topology shown in Fig. 28(e).

[00328] Let us define respectively Z=R+jωL for Figs. 31(a),(e), Z=R+jωL+1/jωC_3 for Figs. 31(b),(f), and Z=(R+jωL)||(1/jωC_3) for Fig. 31(c),(g), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.31(a)-(c) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X - \frac{1}{2}\sqrt{X^2 R_n - R^2(4-R_n)}}{X^2 + R^2},$$

$$\omega C_2 = \frac{R_n \omega C_1}{1 - X\omega C_1 - \frac{R_n}{2}},$$

and these topologies can match the impedances satisfying the inequalities:

$$R_n \le 1, \quad X_n \ge \sqrt{\frac{R_n}{1-R_n}(2-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 31(d). For Figs.31(e)-(g) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X + \frac{1}{2}\sqrt{X^2 R_n - R^2(4-R_n)}}{X^2 + R^2},$$

$$\omega L_2 = -\frac{1 - X\omega C_1 - \frac{R_n}{2}}{R_n \omega C_1}.$$

[00329] Fig. 32 shows three examples (a)-(c) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and three examples (e)-(g) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 32(a),(b),(c),(e),(f),(g), the ground terminal may be connected at the center of the inductive element to preserve the voltage node of the resonator at that point and thus the high Q. Note that these example may be described as implementations of the general topology shown in Fig. 28(e).

[00330] Let us define respectively Z=R+jωL for Fig. 32(a), Z=R+jωL+1/jωC_3 for Fig. 32(b), and Z=(R+jωL)||(1/jωC_3) for Fig. 32(c), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.32(a)-(c) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X - \sqrt{\frac{X^2 R_n - 2R^2(2-R_n)}{4-R_n}}}{X^2 + R^2},$$
$$\omega C_2 = \frac{R_n \omega C_1}{1 - X\omega C_1 - \frac{R_n}{2} + \frac{R_n X \omega C_1}{2(1+k)}},$$

where k is defined by $M' = -kL'$, where L' is the inductance of each half of the inductor loop and M' is the mutual inductance between the two halves, and these topologies can match the impedances satisfying the inequalities:

$$R_n \le 2, \quad X_n \ge \sqrt{2R_n(2-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 32(d).
For Figs.32(e)-(g) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X + \sqrt{\frac{X^2 R_n - 2R^2(2-R_n)}{4-R_n}}}{X^2 + R^2},$$

[00331] In the circuits of Figs. 30, 31, 32, the capacitor, C_2, or the inductor, L_2, is (or the two capacitors, $2C_2$, or the two inductors, $L_2/2$, are) in series with the terminals and may not need to have very low series-loss or withstand a large current.

[00332] Fig. 33 shows six examples (a)-(f) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and six examples (h)-(m) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 33(a),(b),(c),(h),(i),(j), a common-mode signal may be required at the two terminals to preserve the voltage node of the resonator at the center of the inductive element and thus the high *Q*. Note that these examples may be described as implementations of the general topology shown in Fig. 28(c), where B and C are short-circuits and A is not balanced. For the symmetric topologies of Figs. 33(d),(e),(f),(k),(l),(m), the two terminals may need to be driven anti-symmetrically (balanced drive) to preserve the voltage node of the resonator at the center of the inductive element and thus the high Q. Note that these examples may be described as implementations of the general topology shown in Fig. 28(d), where B and C are short-circuits and A is balanced.

[00333] Let us define respectively Z=R+jωL for Figs. 33(a),(d),(h),(k), Z=R+jωL+1/jωC_3 for Figs. 33(b),(e),(i),(l), and Z=(R+jωL)||(1/jωC_3) for Figs. 33(c),(f),(j),(m), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.33(a)-(f) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{1}{X - Z_o\sqrt{R_n(1-R_n)}},$$

$$\omega C_2 = \frac{1}{Z_o}\sqrt{\frac{1}{R_n}-1},$$

and these topologies can match the impedances satisfying the inequalities:

$$R_n \le 1, \quad X_n \ge \sqrt{R_n(1-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 33(g). For Figs.35(h)-(m) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{1}{X + Z_o\sqrt{R_n(1-R_n)}},$$

$$\omega L_2 = \frac{Z_o}{\sqrt{\frac{1}{R_n}-1}}.$$

[00334] Fig. 34 shows three examples (a)-(c) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and three examples (e)-(g) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 34(a),(b),(c),(e),(f),(g), the ground terminal is connected between two equal-value capacitors, $2C_2$, (namely on the axis of symmetry of the main resonator) to preserve the voltage node of the resonator at the center of the inductive element and thus the high Q. Note that these examples may be described as implementations of the general topology shown in Fig. 28(e).

[00335] Let us define respectively Z=R+jωL for Fig. 34(a),(e), $Z=R+j\omega L+1/j\omega C_3$ for Fig. 34(b),(f), and $Z=(R+j\omega L)\|(1/j\omega C_3)$ for Fig. 34(c),(g), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.34(a)-(c) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{1}{X - Z_o\sqrt{\frac{1-R_n}{R_n}(2-R_n)}},$$

$$\omega C_2 = \frac{1}{2Z_o}\sqrt{\frac{1}{R_n}-1},$$

and these topologies can match the impedances satisfying the inequalities:

$$R_n \le 1,\ \ X_n \ge \sqrt{\frac{R_n}{1-R_n}(2-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 34(d).
For Figs.34(e)-(g) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{1}{X + Z_o\sqrt{\frac{1-R_n}{R_n}(2-R_n)}},$$

$$\omega L_2 = \frac{2Z_o}{\sqrt{\frac{1}{R_n}-1}}.$$

[00336] Fig. 35 shows three examples of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors. For the topologies of Figs. 35, the ground terminal may be connected at the center of the inductive element to preserve the voltage node of the resonator at that point and thus the high Q. Note that these examples may be described as implementations of the general topology shown in Fig. 28(e).

[00337] Let us define respectively Z=R+jωL for Fig. 35(a), Z=R+jωL+1/jωC_3 for Fig. 35(b), and Z=(R+jωL)||(1/jωC_3) for Fig. 35(c), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{2}{X(1+a) - \sqrt{Z_o R(4-R_n)(1+a^2)}},$$
$$\omega C_2 = \frac{2}{X(1+a) + \sqrt{Z_o R(4-R_n)(1+a^2)}},$$

where $a = \frac{R}{2Z_o - R} \cdot \frac{k}{1+k}$ and k is defined by $M' = -kL'$, where L' is the inductance of each half of the inductive element and M' is the mutual inductance between the two halves. These topologies can match the impedances satisfying the inequalities:

$$R_n \le 2 \,\&\, \frac{2}{\gamma} \le R_n \le 4,$$
$$X_n \ge \sqrt{\frac{R_n(4-R_n)(2-R_n)}{2-\gamma R_n}},$$

where

$$\gamma = \frac{1-6k+k^2}{1+2k+k^2} \le 1$$

which are shown by the area enclosed by the bold lines on the three Smith charts shown in Fig. 35(d) for k=0, Fig. 35(e) for k=0.05, and Fig. 35(f) for k=1. Note that for 0<k<1 there are two disconnected regions of the Smith chart that this topology can match.

[00338] In the circuits of Figs. 33, 34, 35, the capacitor, C_2, or the inductor, L_2, is (or one of the two capacitors, $2C_2$, or one of the two inductors, $2L_2$, are) in parallel with the terminals and thus may not need to have a high voltage-rating. In the case of two capacitors, $2C_2$, or two inductors, $2L_2$, both may not need to have a high voltage-rating, since approximately the same current flows through them and thus they experience approximately the same voltage across them.

[00339] For the topologies of Figs. 30-35, where a capacitor, C_3, is used, the use of the capacitor, C_3, may lead to finer tuning of the frequency and the impedance. For the topologies of

Figs. 30-35, the use of the fixed capacitor, C_3, in series with the inductive element may ensure that a large percentage of the high inductive-element voltage will be across this fixed capacitor, C_3, thus potentially alleviating the voltage rating requirements for the other elements of the impedance matching circuit, some of which may be variable. Whether or not such topologies are preferred depends on the availability, cost and specifications of appropriate fixed and tunable components.

[00340] In all the above examples, a pair of equal-value variable capacitors without a common terminal may be implemented using ganged-type capacitors or groups or arrays of varactors or diodes biased and controlled to tune their values as an ensemble. A pair of equal-value variable capacitors with one common terminal can be implemented using a tunable butterfly-type capacitor or any other tunable or variable capacitor or group or array of varactors or diodes biased and controlled to tune their capacitance values as an ensemble.

[00341] Another criterion which may be considered upon the choice of the impedance matching network is the response of the network to different frequencies than the desired operating frequency. The signals generated in the external circuit, to which the inductive element is coupled, may not be monochromatic at the desired frequency but periodic with the desired frequency, as for example the driving signal of a switching amplifier or the reflected signal of a switching rectifier. In some such cases, it may be desirable to suppress the amount of higher-order harmonics that enter the inductive element (for example, to reduce radiation of these harmonics from this element). Then the choice of impedance matching network may be one that sufficiently suppresses the amount of such harmonics that enters the inductive element.

[00342] The impedance matching network may be such that the impedance seen by the external circuit at frequencies higher than the fundamental harmonic is high, when the external periodic signal is a signal that can be considered to behave as a voltage-source signal (such as the driving signal of a class-D amplifier with a series resonant load), so that little current flows through the inductive element at higher frequencies. Among the topologies of Figs. 30-35, those which use an inductor, L_2, may then be preferable, as this inductor presents a high impedance at high frequencies.

[00343] The impedance matching network may be such that the impedance seen by the external circuit at frequencies higher than the fundamental harmonic is low, when the external periodic signal is a signal that can be considered to behave as a current-source signal, so that little voltage is induced across the inductive element at higher frequencies. Among the topologies of Figs. 30-35, those which use a capacitor, C_2, are then preferable, as this capacitor presents a low impedance at high frequencies.

[00344] Fig. 36 shows four examples of a variable capacitance, using networks of one variable capacitor and the rest fixed capacitors. Using these network topologies, fine tunability of the total capacitance value may be achieved. Furthermore, the topologies of Figs. 36(a),(c),(d), may be used to reduce the voltage across the variable capacitor, since most of the voltage may be assigned across the fixed capacitors.

[00345] Fig. 37 shows two examples of a variable capacitance, using networks of one variable inductor and fixed capacitors. In particular, these networks may provide implementations for a variable reactance, and, at the frequency of interest, values for the variable inductor may be used such that each network corresponds to a net negative variable reactance, which may be effectively a variable capacitance.

[00346] Tunable elements such as tunable capacitors and tunable inductors may be mechanically-tunable, electrically-tunable, thermally-tunable and the like. The tunable elements may be variable capacitors or inductors, varactors, diodes, Schottky diodes, reverse-biased PN diodes, varactor arrays, diode arrays, Schottky diode arrays and the like. The diodes may be Si diodes, GaN diodes, SiC diodes, and the like. GaN and SiC diodes may be particularly attractive for high power applications. The tunable elements may be electrically switched capacitor banks, electrically-switched mechanically-tunable capacitor banks, electrically-switched varactor-array banks, electrically-switched transformer-coupled inductor banks, and the like. The tunable elements may be combinations of the elements listed above.

[00347] As described above, the efficiency of the power transmission between coupled high-Q magnetic resonators may be impacted by how closely matched the resonators are in resonant frequency and how well their impedances are matched to the power supplies and power consumers in the system. Because a variety of external factors including the relative position of extraneous objects or other resonators in the system, or the changing of those relative positions, may alter the resonant frequency and/or input impedance of a high-Q magnetic resonator, tunable impedance networks may be required to maintain sufficient levels of power transmission in various environments or operating scenarios.

[00348] The capacitance values of the capacitors shown may be adjusted to adjust the resonant frequency and/or the impedance of the magnetic resonator. The capacitors may be adjusted electrically, mechanically, thermally, or by any other known methods. They may be adjusted manually or automatically, such as in response to a feedback signal. They may be adjusted to achieve certain power transmission efficiencies or other operating characteristics between the power supply and the power consumer.

[00349] The inductance values of the inductors and inductive elements in the resonator may be adjusted to adjust the frequency and/or impedance of the magnetic resonator. The inductance may be adjusted using coupled circuits that include adjustable components such as tunable capacitors, inductors and switches. The inductance may be adjusted using transformer coupled tuning circuits. The inductance may be adjusted by switching in and out different sections of conductor in the inductive elements and/or using ferro-magnetic tuning and/or mu-tuning, and the like.

[00350] The resonant frequency of the resonators may be adjusted to or may be allowed to change to lower or higher frequencies. The input impedance of the resonator may be adjusted to or may be allowed to change to lower or higher impedance values. The amount of power delivered by the source and/or received by the devices may be adjusted to or may be allowed to change to lower or higher levels of power. The amount of power delivered to the source and/or received by the devices from the device resonator may be adjusted to or may be allowed to change to lower or higher levels of power. The resonator input impedances, resonant frequencies, and power levels may be adjusted depending on the power consumer or consumers in the system and depending on the objects or materials in the vicinity of the resonators. The resonator input impedances, frequencies, and power levels may be adjusted manually or automatically, and may be adjusted in response to feedback or control signals or algorithms.

[00351] Circuit elements may be connected directly to the resonator, that is, by physical electrical contact, for example to the ends of the conductor that forms the inductive element and/or the terminal connectors. The circuit elements may be soldered to, welded to, crimped to, glued to, pinched to, or closely position to the conductor or attached using a variety of electrical components, connectors or connection techniques. The power supplies and the power consumers may be connected to magnetic resonators directly or indirectly or inductively.

Electrical signals may be supplied to, or taken from, the resonators through the terminal connections.

[00352] It is to be understood by one of ordinary skill in the art that in real implementations of the principles described herein, there may be an associated tolerance, or acceptable variation, to the values of real components (capacitors, inductors, resistors and the like) from the values calculated via the herein stated equations, to the values of real signals (voltages, currents and the like) from the values suggested by symmetry or anti-symmetry or otherwise, and to the values of real geometric locations of points (such as the point of connection of the ground terminal close to the center of the inductive element or the 'axis' points and the like) from the locations suggested by symmetry or otherwise.

[00353] **Examples**

[00354] SYSTEM BLOCK DIAGRAMS

[00355] We disclose examples of high-Q resonators for wireless power transmission systems that may wirelessly power or charge devices at mid-range distances. High-*Q* resonator wireless power transmission systems also may wirelessly power or charge devices with magnetic resonators that are different in size, shape, composition, arrangement, and the like, from any source resonators in the system.

[00356] Fig. 1(a)(b) shows high level diagrams of two exemplary two-resonator systems. These exemplary systems each have a single source resonator 102S or 104S and a single device resonator 102D or 104D. Fig. 38 shows a high level block diagram of a system with a few more features highlighted. The wirelessly powered or charged device 2310 may include or consist of a device resonator 102D, device power and control circuitry 2304, and the like, along with the device 2308 or devices, to which either DC or AC or both AC and DC power is transferred. The energy or power source for a system may include the source power and control circuitry 2302, a source resonator 102S, and the like. The device 2308 or devices that receive power from the device resonator 102D and power and control circuitry 2304 may be any kind of device 2308 or devices as described previously. The device resonator 102D and circuitry 2304 delivers power to the device/devices 2308 that may be used to recharge the battery of the device/devices, power the device/devices directly, or both when in the vicinity of the source resonator 102S.

[00357] The source and device resonators may be separated by many meters or they may be very close to each other or they may be separated by any distance in between. The source and device resonators may be offset from each other laterally or axially. The source and device resonators may be directly aligned (no lateral offset), or they may be offset by meters, or anything in between. The source and device resonators may be oriented so that the surface areas enclosed by their inductive elements are approximately parallel to each other. The source and device resonators may be oriented so that the surface areas enclosed by their inductive elements are approximately perpendicular to each other, or they may be oriented for any relative angle (0 to 360 degrees) between them.

[00358] The source and device resonators may be free standing or they may be enclosed in an enclosure, container, sleeve or housing. These various enclosures may be composed of almost any kind of material. Low loss tangent materials such as Teflon, REXOLITE, styrene, and the like may be preferable for some applications. The source and device resonators may be integrated in the power supplies and power consumers. For example, the source and device resonators may be integrated into keyboards, computer mice, displays, cell phones, etc. so that they are not visible outside these devices. The source and device resonators may be separate from the power supplies and power consumers in the system and may be connected by a standard or custom wires, cables, connectors or plugs.

[00359] The source 102S may be powered from a number of DC or AC voltage, current or power sources including a USB port of a computer. The source 102S may be powered from the electric grid, from a wall plug, from a battery, from a power supply, from an engine, from a solar cell, from a generator, from another source resonator, and the like. The source power and control circuitry 2302 may include circuits and components to isolate the source electronics from the power source, so that any reflected power or signals are not coupled out through the source input terminals. The source power and control circuits 2302 may include power factor correction circuits and may be configured to monitor power usage for monitoring accounting, billing, control, and like functionalities.

[00360] The system may be operated bi-directionally. That is, energy or power that is generated or stored in a device resonator may be fed back to a power source including the electric grid, a battery, any kind of energy storage unit, and the like. The source power and control circuits may include power factor correction circuits and may be configured to monitor power usage for monitoring accounting, billing, control, and like functionalities for bi-directional energy flow. Wireless energy transfer systems may enable or promote vehicle-to-grid (V2G) applications.

[00361] The source and the device may have tuning capabilities that allow adjustment of operating points to compensate for changing environmental conditions, perturbations, and loading conditions that can affect the operation of the source and device resonators and the efficiency of the energy exchange. The tuning capability may also be used to multiplex power delivery to multiple devices, from multiple sources, to multiple systems, to multiple repeaters or relays, and the like. The tuning capability may be manually controlled, or automatically controlled and may be performed continuously, periodically, intermittently or at scheduled times or intervals.

[00362] The device resonator and the device power and control circuitry may be integrated into any portion of the device, such as a battery compartment, or a device cover or sleeve, or on a mother board, for example, and may be integrated alongside standard rechargeable batteries or other energy storage units. The device resonator may include a device field reshaper which may shield any combination of the device resonator elements and the device power and control electronics from the electromagnetic fields used for the power transfer and which may deflect the resonator fields away from the lossy device resonator elements as well as the device power and control electronics. A magnetic material and/or high-conductivity field reshaper may be used to increase the perturbed quality factor Q of the resonator and increase the perturbed coupling factor of the source and device resonators.

[00363] The source resonator and the source power and control circuitry may be integrated into any type of furniture, structure, mat, rug, picture frame (including digital picture frames, electronic frames), plug-in modules, electronic devices, vehicles, and the like. The source resonator may include a source field reshaper which may shield any combination of the source resonator elements and the source power and control electronics from the electromagnetic fields used for the power transfer and which may deflect the resonator fields away from the lossy source resonator elements as well as the source power and control electronics. A magnetic material and/or high-conductivity field reshaper may be used to increase the perturbed quality factor Q of the resonator and increase the perturbed coupling factor of the source and device resonators.

[00364] A block diagram of the subsystems in an example of a wirelessly powered device is shown in Fig. 39. The power and control circuitry may be designed to transform the alternating current power from the device resonator 102D and convert it to stable direct current power suitable for powering or charging a device. The power and control circuitry may be designed to transform an alternating current power at one frequency from the device resonator to alternating current power at a different frequency suitable for powering or charging a device. The power and control circuitry may include or consist of impedance matching circuitry 2402D, rectification circuitry 2404, voltage limiting circuitry (not shown), current limiting circuitry (not shown), AC-to-DC converter 2408 circuitry, DC-to-DC converter 2408 circuitry, DC-to-AC converter 2408 circuitry, AC-to-AC converter 2408 circuitry, battery charge control circuitry (not shown), and the like.

[00365] The impedance-matching 2402D network may be designed to maximize the power delivered between the device resonator 102D and the device power and control circuitry 2304 at the desired frequency. The impedance matching elements may be chosen and connected such that the high-Q of the resonators is preserved. Depending on the operating conditions, the impedance matching circuitry 2402D may be varied or tuned to control the power delivered from the source to the device, from the source to the device resonator, between the device resonator and the device power and control circuitry, and the like. The power, current and voltage signals may be monitored at any point in the device circuitry and feedback algorithms circuits, and techniques, may be used to control components to achieve desired signal levels and system operation. The feedback algorithms may be implemented using analog or digital circuit techniques and the circuits may include a microprocessor, a digital signal processor, a field programmable gate array processor and the like.

[00366] The third block of Fig. 39 shows a rectifier circuit 2404 that may rectify the AC voltage power from the device resonator into a DC voltage. In this configuration, the output of the rectifier 2404 may be the input to a voltage clamp circuit. The voltage clamp circuit (not shown) may limit the maximum voltage at the input to the DC-to-DC converter 2408D or DC-to-AC converter 2408D. In general, it may be desirable to use a DC-to-DC/AC converter with a large input voltage dynamic range so that large variations in device position and operation may be tolerated while adequate power is delivered to the device. For example, the voltage level at the output of the rectifier may fluctuate and reach high levels as the power input and load characteristics of the device change. As the device performs different tasks it may have varying power demands. The changing power demands can cause high voltages at the output of the rectifier as the load characteristics change. Likewise as the device and the device resonator are brought closer and further away from the source, the power delivered to the device resonator may vary and cause changes in the voltage levels at the output of the rectifier. A voltage clamp circuit may prevent the voltage output from the rectifier circuit from exceeding a predetermined value which is within the operating range of the DC-to-DC/AC converter. The voltage clamp circuitry may be used to extend the operating modes and ranges of a wireless energy transfer system.

[00367] The next block of the power and control circuitry of the device is the DC-to-DC converter 2408D that may produce a stable DC output voltage. The DC-to-DC converter may be a boost converter, buck converter, boost-buck converter, single ended primary inductance converter (SEPIC), or any other DC-DC topology that fits the requirements of the particular application. If the device requires AC power, a DC-to-AC converter may be substituted for the DC-to-DC converter, or the DC-to-DC converter may be followed by a DC-to-AC converter. If the device contains a rechargeable battery, the final block of the device power and control circuitry may be a battery charge control unit which may manage the charging and maintenance of the battery in battery powered devices.

[00368] The device power and control circuitry 2304 may contain a processor 2410D, such as a microcontroller, a digital signal processor, a field programmable gate array processor, a microprocessor, or any other type of processor. The processor may be used to read or detect the state or the operating point of the power and control circuitry and the device resonator. The processor may implement algorithms to interpret and adjust the operating point of the circuits, elements, components, subsystems and resonator. The processor may be used to adjust the impedance matching, the resonator, the DC to DC converters, the DC to AC converters, the battery charging unit, the rectifier, and the like of the wirelessly powered device.

[00369] The processor may have wireless or wired data communication links to other devices or sources and may transmit or receive data that can be used to adjust the operating point of the system. Any combination of power, voltage, and current signals at a single, or over a range of frequencies, may be monitored at any point in the device circuitry. These signals may be monitored using analog or digital or combined analog and digital techniques. These monitored signals may be used in feedback loops or may be reported to the user in a variety of known ways or they may be stored and retrieved at later times. These signals may be used to alert a user of system failures, to indicate performance, or to provide audio, visual, vibrational, and the like, feedback to a user of the system.

[00370] Fig. 40 shows components of source power and control circuitry 2302 of an exemplary wireless power transfer system configured to supply power to a single or multiple devices. The source power and control circuitry 2302 of the exemplary system may be powered from an AC voltage source 2502 such as a home electrical outlet, a DC voltage source such as a battery, a USB port of a computer, a solar cell, another wireless power source, and the like. The source power and control circuitry 2302 may drive the source resonator 102S with alternating current, such as with a frequency greater than 10 kHz and less than 100 MHz. The source power and control circuitry 2302 may drive the source resonator 102S with alternating current of frequency less than less than 10 GHz. The source power and control circuitry 2302 may include a DC-to-DC converter 2408S, an AC-to-DC converter 2408S, or both an AC-to-DC converter 2408S and a DC-to-DC 2408S converter, an oscillator 2508, a power amplifier 2504, an impedance matching network 2402S, and the like.

[00371] The source power and control circuitry 2302 may be powered from multiple AC-or-DC voltage sources 2502 and may contain AC-to-DC and DC-to-DC converters 2408S to provide necessary voltage levels for the circuit components as well as DC voltages for the power amplifiers that may be used to drive the source resonator. The DC voltages may be adjustable and may be used to control the output power level of the power amplifier. The source may contain power factor correction circuitry.

[00372] The oscillator 2508 output may be used as the input to a power amplifier 2504 that drives the source resonator 102S. The oscillator frequency may be tunable and the amplitude of the oscillator signal may be varied as one means to control the output power level from the power amplifier. The frequency, amplitude, phase, waveform, and duty cycle of the oscillator signal may be controlled by analog circuitry, by digital circuitry or by a combination of analog and digital circuitry. The control circuitry may include a processor 2410S, such as a microprocessor, a digital signal processor, a field programmable gate array processor, and the like.

[00373] The impedance matching blocks 2402 of the source and device resonators may be used to tune the power and control circuits and the source and device resonators. For example, tuning of these circuits may adjust for perturbation of the quality factor Q of the source or device resonators due to extraneous objects or changes in distance between the source and device in a system. Tuning of these circuits may also be used to sense the operating environment, control power flow to one or more devices, to control power to a wireless power network, to reduce power when unsafe or failure mode conditions are detected, and the like.

[00374] Any combination of power, voltage, and current signals may be monitored at any point in the source circuitry. These signals may be monitored using analog or digital or combined analog and digital techniques. These monitored signals may be used in feedback circuits or may be reported to the user in a variety of known ways or they may be stored and retrieved at later times. These signals may be used to alert a user to system failures, to alert a user to exceeded safety thresholds, to indicate performance, or to provide audio, visual, vibrational, and the like, feedback to a user of the system.

[00375] The source power and control circuitry may contain a processor. The processor may be used to read the state or the operating point of the power and control circuitry and the source resonator. The processor may implement algorithms to interpret and adjust the operating point of the circuits, elements, components, subsystems and resonator. The processor may be used to adjust the impedance matching, the resonator, the DC-to-DC converters, the AC-to-DC converters, the oscillator, the power amplifier of the source, and the like. The processor and adjustable components of the system may be used to implement frequency and/or time power delivery multiplexing schemes. The processor may have wireless or wired data communication links to devices and other sources and may transmit or receive data that can be used to adjust the operating point of the system.

[00376] Although detailed and specific designs are shown in these block diagrams, it should be clear to those skilled in the art that many different modifications and rearrangements of the components and building blocks are possible within the spirit of the exemplary system. The division of the circuitry was outlined for illustrative purposes and it should be clear to those skilled in the art that the components of each block may be further divided into smaller blocks or merged or shared. In equivalent examples the power and control circuitry may be composed of individual discrete components or larger integrated circuits. For example, the rectifier circuitry may be composed of discrete diodes, or use diodes integrated on a single chip. A multitude of other circuits and integrated devices can be substituted in the design depending on design criteria such as power or size or cost or application. The whole of the power and control circuitry or any portion of the source or device circuitry may be integrated into one chip.

[00377] The impedance matching network of the device and or source may include a capacitor or networks of capacitors, an inductor or networks of inductors, or any combination of capacitors, inductors, diodes, switches, resistors, and the like. The components of the impedance matching network may be adjustable and variable and may be controlled to affect the efficiency and operating point of the system. The impedance matching may be performed by controlling the connection point of the resonator, adjusting the permeability of a magnetic material, controlling a bias field, adjusting the frequency of excitation, and the like. The impedance matching may use or include any number or combination of varactors, varactor arrays, switched elements, capacitor banks, switched and tunable elements, reverse bias diodes, air gap capacitors, compression capacitors, BZT electrically tuned capacitors, MEMS-tunable capacitors, voltage variable dielectrics, transformer coupled tuning circuits, and the like. The variable components may be mechanically tuned, thermally tuned, electrically tuned, piezo-electrically tuned, and the like. Elements of the impedance matching may be silicon devices, gallium nitride devices, silicon carbide devices and the like. The elements may be chosen to withstand high currents, high voltages, high powers, or any combination of current, voltage and power. The elements may be chosen to be high-Q elements.

[00378] The matching and tuning calculations of the source may be performed on an external device through a USB port that powers the device. The device may be a computer a PDA or other computational platform.

[00379] A demonstration system used a source resonator, coupled to a device resonator, to wirelessly power/recharge multiple electronic consumer devices including, but not limited to, a laptop, a DVD player, a projector, a cell-phone, a display, a television, a projector, a digital picture frame, a light, a TV/DVD player, a portable music player, a circuit breaker, a hand-held tool, a personal digital assistant, an external battery charger, a mouse, a keyboard, a camera, an active load, and the like. A variety of devices may be powered simultaneously from a single device resonator. Device resonators may be operated simultaneously as source resonators.

The power supplied to a device resonator may pass through additional resonators before being delivered to its intended device resonator.

[00380] *Monitoring, Feedback and Control*

[00381] So-called port parameter measurement circuitry may measure or monitor certain power, voltage, and current, signals in the system and processors or control circuits may adjust certain settings or operating parameters based on those measurements. In addition to these port parameter measurements, the magnitude and phase of voltage and current signals, and the magnitude of the power signals, throughout the system may be accessed to measure or monitor the system performance. The measured signals referred to throughout this disclosure may be any combination of the port parameter signals, as well as voltage signals, current signals, power signals, and the like. These parameters may be measured using analog or digital signals, they may be sampled and processed, and they may be digitized or converted using a number of known analog and digital processing techniques. Measured or monitored signals may be used in feedback circuits or systems to control the operation of the resonators and/or the system. In general, we refer to these monitored or measured signals as reference signals, or port parameter measurements or signals, although they are sometimes also referred to as error signals, monitor signals, feedback signals, and the like. We will refer to the signals that are used to control circuit elements such as the voltages used to drive voltage controlled capacitors as the control signals.

[00382] In some cases the circuit elements may be adjusted to achieve a specified or predetermined impedance value for the source and device resonators. In other cases the impedance may be adjusted to achieve a desired impedance value for the source and device resonators when the device resonator is connected to a power consumer or consumers. In other cases the impedance may be adjusted to mitigate changes in the resonant frequency, or impedance or power level changes owing to movement of the source and/or device resonators, or changes in the environment (such as the movement of interacting materials or objects) in the vicinity of the resonators. In other cases the impedance of the source and device resonators may be adjusted to different impedance values.

[00383] The coupled resonators may be made of different materials and may include different circuits, components and structural designs or they may be the same. The coupled resonators may include performance monitoring and measurement circuitry, signal processing and control circuitry or a combination of measurement and control circuitry. Some or all of the high-Q magnetic resonators may include tunable impedance circuits. Some or all of the high-Q magnetic resonators may include automatically controlled tunable impedance circuits.

[00384] Fig. 41 shows a magnetic resonator with port parameter measurement circuitry 3802 configured to measure certain parameters of the resonator. The port parameter measurement circuitry may measure the input impedance of the structure, or the reflected power. Port parameter measurement circuits may be included in the source and/or device resonator designs and may be used to measure two port circuit parameters such as S-parameters (scattering parameters), Z-parameters (impedance parameters), Y-parameters (admittance parameters), T-parameters (transmission parameters), H-parameters (hybrid parameters), ABCD-parameters (chain, cascade or transmission parameters), and the like. These parameters may be used to describe the electrical behavior of linear electrical networks when various types of signals are applied.

[00385] Different parameters may be used to characterize the electrical network under different operating or coupling scenarios. For example, S-parameters may be used to measure matched and unmatched loads. In addition, the magnitude and phase of voltage and current signals within the magnetic resonators and/or within the sources and devices themselves may be monitored at a variety of points to yield system performance information. This information may be presented to users of the system via a user interface such as a light, a read-out, a beep, a noise, a vibration or the like, or it may be presented as a digital signal or it may be provided to a processor in the system and used in the automatic control of the system. This information may be logged, stored, or may be used by higher level monitoring and control systems.

[00386] Fig. 42 shows a circuit diagram of a magnetic resonator where the tunable impedance network may be realized with voltage controlled capacitors 3902 or capacitor networks. Such an implementation may be adjusted, tuned or controlled by electrical circuits and/or computer processors, such as a programmable voltage source 3908, and the like. For example, the voltage controlled capacitors may be adjusted in response to data acquired by the port parameter measurement circuitry 3802 and processed by a measurement analysis and control algorithm subsystem 3904. Reference signals may be derived from the port parameter measurement circuitry or other monitoring circuitry designed to measure the degree of deviation from a desired system operating point. The measured reference signals may include voltage, current, complex-impedance, reflection coefficient, power levels and the like, at one or several points in the system and at a single frequency or at multiple frequencies.

[00387] The reference signals may be fed to measurement analysis and control algorithm subsystem modules that may generate control signals to change the values of various components in a tunable impedance matching network. The control signals may vary the resonant frequency and/or the input impedance of the magnetic resonator, or the power level supplied by the source, or the power level drawn by the device, to achieve the desired power exchange between power supplies/generators and power drains/loads.

[00388] Adjustment algorithms may be used to adjust the frequency and/or impedance of the magnetic resonators. The algorithms may take in reference signals related to the degree of deviation from a desired operating point for the system and output correction or control signals related to that deviation that control variable or tunable elements of the system to bring the system back towards the desired operating point or points. The reference signals for the magnetic resonators may be acquired while the resonators are exchanging power in a wireless power transmission system, or they may be switched out of the circuit during system operation. Corrections to the system may be applied or performed continuously, periodically, upon a threshold crossing, digitally, using analog methods, and the like.

[00389] Fig. 43 shows an end-to-end wireless power transmission system. Both the source and the device may include port measurement circuitry 3802 and a processor 2410. The box labeled "coupler/switch" 4002 indicates that the port measurement circuitry 3802 may be connected to the resonator 102 by a directional coupler or a switch, enabling the measurement, adjustment and control of the source and device resonators to take place in conjunction with, or separate from, the power transfer functionality.

[00390] The port parameter measurement and/or processing circuitry may reside with some, any, or all resonators in a system. The port parameter measurement circuitry may utilize portions of the power transmission signal or may utilize excitation signals over a range of frequencies to measure the source/device resonator response (i.e. transmission and reflection between any two ports in the system), and may contain amplitude and/or phase information. Such measurements may be achieved with a swept single frequency signal or a multi-frequency signal. The signals used to measure and monitor the resonators and the wireless power transmission system may be generated by a processor or processors and standard input/output

(I/O) circuitry including digital to analog converters (DACs), analog to digital converters (ADCs), amplifiers, signal generation chips, passive components and the like. Measurements may be achieved using test equipment such as a network analyzer or using customized circuitry. The measured reference signals may be digitized by ADCs and processed using customized algorithms running on a computer, a microprocessor, a DSP chip, an ASIC, and the like. The measured reference signals may be processed in an analog control loop.

[00391] The measurement circuitry may measure any set of two port parameters such as S-parameters, Y-parameters, Z-parameters, H-parameters, G-parameters, T-parameters, ABCD-parameters, and the like. Measurement circuitry may be used to characterize current and voltage signals at various points in the drive and resonator circuitry, the impedance and/or admittance of the source and device resonators at opposite ends of the system, i.e. looking into the source resonator matching network ("port 1" in Fig. 43) towards the device and vice versa.

[00392] The device may measure relevant signals and/or port parameters, interpret the measurement data, and adjust its matching network to optimize the impedance looking into the coupled system independently of the actions of the source. The source may measure relevant port parameters, interpret the measurement data, and adjust its matching network to optimize the impedance looking into the coupled system independently of the actions of the device.

[00393] Fig. 43 shows a block diagram of a source and device in a wireless power transmission system. The system may be configured to execute a control algorithm that actively adjusts the tuning/matching networks in either of or both the source and device resonators to optimize performance in the coupled system. Port measurement circuitry 3802S may measure signals in the source and communicate those signals to a processor 2410. A processor 2410 may use the measured signals in a performance optimization or stabilization algorithm and generate control signals based on the outputs of those algorithms. Control signals may be applied to variable circuit elements in the tuning/impedance matching circuits 2402S to adjust the source's operating characteristics, such as power in the resonator and coupling to devices. Control signals may be applied to the power supply or generator to turn the supply on or off, to increase or decrease the power level, to modulate the supply signal and the like.

[00394] The power exchanged between sources and devices may depend on a variety of factors. These factors may include the effective impedance of the sources and devices, the Q's of the sources and devices, the resonant frequencies of the sources and devices, the distances between sources and devices, the interaction of materials and objects in the vicinity of sources and devices and the like. The port measurement circuitry and processing algorithms may work in concert to adjust the resonator parameters to maximize power transfer, to hold the power transfer constant, to controllably adjust the power transfer, and the like, under both dynamic and steady state operating conditions.

[00395] Some, all or none of the sources and devices in a system implementation may include port measurement circuitry 3802S and processing 2410 capabilities. Fig. 44 shows an end-to-end wireless power transmission system in which only the source 102S contains port measurement circuitry 3802 and a processor 2410S. In this case, the device resonator 102D operating characteristics may be fixed or may be adjusted by analog control circuitry and without the need for control signals generated by a processor.

[00396] Fig. 45 shows an end-to-end wireless power transmission system. Both the source and the device may include port measurement circuitry 3802 but in the system of Fig. 45, only the source contains a processor 2410S. The source and device may be in communication with each other and the adjustment of certain system parameters may be in response to control signals that have been wirelessly communicated, such as though wireless communications circuitry 4202, between the source and the device. The wireless communication channel 4204 may be separate from the wireless power transfer channel 4208, or it may be the same. That is, the resonators 102 used for power exchange may also be used to exchange information. In some cases, information may be exchanged by modulating a component a source or device circuit and sensing that change with port parameter or other monitoring equipment.

[00397] Implementations where only the source contains a processor 2410 may be beneficial for multi-device systems where the source can handle all of the tuning and adjustment "decisions" and simply communicate the control signals back to the device(s). This implementation may make the device smaller and cheaper because it may eliminate the need for, or reduce the required functionality of, a processor in the device. A portion of or an entire data set from each port measurement at each device may be sent back to the source microprocessor for analysis, and the control instructions may be sent back to the devices. These communications may be wireless communications.

[00398] Fig. 46 shows an end-to-end wireless power transmission system. In this example, only the source contains port measurement circuitry 3802 and a processor 2410S. The source and device may be in communication, such as via wireless communication circuitry 4202, with each other and the adjustment of certain system parameters may be in response to control signals that have been wirelessly communicated between the source and the device.

[00399] Fig. 47 shows coupled electromagnetic resonators 102 whose frequency and impedance may be automatically adjusted using a processor or a computer. Resonant frequency tuning and continuous impedance adjustment of the source and device resonators may be implemented with reverse biased diodes, Schottky diodes and/or varactor elements contained within the capacitor networks shown as C1, C2, and C3 in Fig. 47. The circuit topology that was built and demonstrated and is described here is exemplary and is not meant to limit the discussion of automatic system tuning and control in any way. Other circuit topologies could be utilized with the measurement and control architectures discussed in this disclosure.

[00400] Device and source resonator impedances and resonant frequencies may be measured with a network analyzer 4402A-B, or by other means described above, and implemented with a controller, such as with Lab View 4404. The measurement circuitry or equipment may output data to a computer or a processor that implements feedback algorithms and dynamically adjusts the frequencies and impedances via a programmable DC voltage source.

[00401] In one arrangement, the reverse biased diodes (Schottky, semiconductor junction, and the like) used to realize the tunable capacitance drew very little DC current and could be reverse biased by amplifiers having large series output resistances. This implementation may enable DC control signals to be applied directly to the controllable circuit elements in the resonator circuit while maintaining a very high-Q in the magnetic resonator.

[00402] C2 biasing signals may be isolated from C1 and/or C3 biasing signals with a DC blocking capacitor as shown in Fig. 47, if the required DC biasing voltages are different. The output of the biasing amplifiers may be bypassed to circuit ground to isolate RF voltages from the biasing amplifiers, and to keep non-fundamental RF voltages from being injected into the resonator. The reverse bias voltages for some of the capacitors may instead be applied through the inductive element in the resonator itself, because the inductive element acts as a short circuit at DC.

[00403] The port parameter measurement circuitry may exchange signals with a processor (including any required ADCs and DACs) as part of a feedback or control system that is used to automatically adjust the resonant frequency, input impedance, energy stored or captured by the resonator or power delivered by a source or to a device load. The processor may also send control signals to tuning or adjustment circuitry in or attached to the magnetic resonator.

[00404] When utilizing varactors or diodes as tunable capacitors, it may be beneficial to place fixed capacitors in parallel and in series with the tunable capacitors operating at high reverse bias voltages in the tuning/matching circuits. This arrangement may yield improvements in circuit and system stability and in power handling capability by optimizing the operating voltages on the tunable capacitors.

[00405] Varactors or other reverse biased diodes may be used as a voltage controlled capacitor. Arrays of varactors may be used when higher voltage compliance or different capacitance is required than that of a single varactor component. Varactors may be arranged in an N by M array connected serially and in parallel and treated as a single two terminal component with different characteristics than the individual varactors in the array. For example, an N by N array of equal varactors where components in each row are connected in parallel and components in each column are connected in series may be used as a two terminal device with the same capacitance as any single varactor in the array but with a voltage compliance that is N times that of a single varactor in the array. Depending on the variability and differences of parameters of the individual varactors in the array additional biasing circuits composed of resistors, inductors, and the like may be needed. A schematic of a four by four array of unbiased varactors 4502 that may be suitable for magnetic resonator applications is shown in Fig. 48.

[00406] Further improvements in system performance may be realized by careful selection of the fixed value capacitor(s) that are placed in parallel and/or in series with the tunable (varactor/diode/capacitor) elements. Multiple fixed capacitors that are switched in or out of the circuit may be able to compensate for changes in resonator Q's, impedances, resonant frequencies, power levels, coupling strengths, and the like, that might be encountered in test, development and operational wireless power transfer systems. Switched capacitor banks and other switched element banks may be used to assure the convergence to the operating frequencies and impedance values required by the system design.

[00407] An exemplary control algorithm for isolated and coupled magnetic resonators may be described for the circuit and system elements shown in Fig. 47. One control algorithm first adjusts each of the source and device resonator loops "in isolation", that is, with the other resonators in the system “shorted out” or “removed” from the system. For practical purposes, a resonator can be “shorted out” by making it resonant at a much lower frequency such as by maximizing the value of C1 and/or C3. This step effectively reduces the coupling between the resonators, thereby effectively reducing the system to a single resonator at a particular frequency and impedance.

[00408] Tuning a magnetic resonator in isolation includes varying the tunable elements in the tuning and matching circuits until the values measured by the port parameter measurement circuitry are at their predetermined, calculated or measured relative values. The desired values for the quantities measured by the port parameter measurement circuitry may be chosen based on the desired matching impedance, frequency, strong coupling parameter, and the like. For the exemplary algorithms disclosed below, the port parameter measurement circuitry measures S-parameters over a range of frequencies. The range of frequencies used to characterize the resonators may be a compromise between the system performance information obtained and computation/measurement speed. For the algorithms described below the frequency range may be approximately +/- 20% of the operating resonant frequency.

[00409] Each isolated resonator may be tuned as follows. First, short out the resonator not being adjusted. Next minimize C1, C2, and C3, in the resonator that is being characterized and adjusted. In most cases there will be fixed circuit elements in parallel with C1, C2, and C3, so this step does not reduce the capacitance values to zero. Next, start increasing C2 until the resonator impedance is matched to the “target” real impedance at any frequency in the range of measurement frequencies described above. The initial “target” impedance may be less than the expected operating impedance for the coupled system.

[00410] C2 may be adjusted until the initial “target” impedance is realized for a frequency in the measurement range. Then C1 and/or C3 may be adjusted until the loop is resonant at the desired operating frequency.

[00411] Each resonator may be adjusted according to the above algorithm. After tuning each resonator in isolation, a second feedback algorithm may be applied to optimize the resonant frequencies and/or input impedances for wirelessly transferring power in the coupled system.

[00412] The required adjustments to C1 and/or C2 and/or C3 in each resonator in the coupled system may be determined by measuring and processing the values of the real and imaginary parts of the input impedance from either and/or both "port(s)" shown in Fig. 43. For coupled resonators, changing the input impedance of one resonator may change the input impedance of the other resonator. Control and tracking algorithms may adjust one port to a desired operating point based on measurements at that port, and then adjust the other port based on measurements at that other port. These steps may be repeated until both sides converge to the desired operating point.

[00413] S-parameters may be measured at both the source and device ports and the following series of measurements and adjustments may be made. In the description that follows, Z_0 is an input impedance and may be the target impedance. In some cases Z_0 is 50 ohms or is near 50 ohms. Z_1 and Z_2 are intermediate impedance values that may be the same value as Z_0 or may be different than Z_0. Re{value} means the real part of a value and Im{value} means the imaginary part of a value.

[00414] An algorithm that may be used to adjust the input impedance and resonant frequency of two coupled resonators is set forth below:

1) Adjust each resonator "in isolation" as described above.
2) Adjust source C1/C3 until, at ω_o, Re{S11} = (Z_1 +/- ε_{Re}) as follows:
   - If Re{S11 @ ω_o } > ($Z_1 + \varepsilon_{Re}$), decrease C1/C3. If Re{S11 @ ω_o } < (Zo - ε_{Re}), increase C1/C3.
3) Adjust source C2 until, at ω_o, Im{S11} = (+/- ε_{Im}) as follows:
   - If Im{S11 @ ω_o} > ε_{Im}, decrease C2. If Im{S11 @ ω_o} < - ε_{Im}, increase C2.
4) Adjust device C1/C3 until, at ω_0, Re{S22} = (Z_2 +/- ε_{Re}) as follows:
   - If Re{S22 @ ω_o} > ($Z_2 + \varepsilon_{Re}$), decrease C1/C3. If Re{S22 @ ω_o} < (Zo - ε_{Re}), increase C1/C3.
5) Adjust device C2 until, at ω_o, Im{S22} = 0 as follows:
   - If Im{S22 @ ω_o} > ε_{Im}, decrease C2. If Im{S22 @ ω_o} < -ε_{Im}, increase C2.

[00415] We have achieved a working system by repeating steps 1-4 until both (Re{S11}, Im{S11}) and (Re{S22}, Im{S22}) converge to ((Z_0 +/- ε_{Re}), (+/- ε_{Im})) at ω_o, where Z_0 is the desired matching impedance and ω_o is the desired operating frequency. Here, ε_{Im} represents the maximum deviation of the imaginary part, at ω_o, from the desired value of 0, and ε_{Re} represents the maximum deviation of the real part from the desired value of Z_0. It is understood that ε_{Im} and ε_{Re} can be adjusted to increase or decrease the number of steps to convergence at the potential cost of system performance (efficiency). It is also understood that steps 1-4 can be performed in a variety of sequences and a variety of ways other than that outlined above (i.e. first adjust the source imaginary part, then the source real part; or first adjust the device real part, then the device imaginary part, etc.) The intermediate impedances Z_1 and Z_2 may be adjusted during steps 1-4 to reduce the number of steps required for convergence. The desire or target impedance value may be complex, and may vary in time or under different operating scenarios.

[00416] Steps 1-4 may be performed in any order, in any combination and any number of times. Having described the above algorithm, variations to the steps or the described implementation may be apparent to one of ordinary skill in the art. The algorithm outlined above may be implemented with any equivalent linear network port parameter measurements (i.e., Z-parameters, Y-parameters, T-parameters, H-parameters, ABCD-parameters, etc.) or other monitor signals described above, in the same way that impedance or admittance can be alternatively used to analyze a linear circuit to derive the same result.

[00417] The resonators may need to be retuned owing to changes in the “loaded” resistances, Rs and Rd, caused by changes in the mutual inductance M (coupling) between the source and device resonators. Changes in the inductances, Ls and Ld, of the inductive elements themselves may be caused by the influence of external objects, as discussed earlier, and may also require compensation. Such variations may be mitigated by the adjustment algorithm described above.

[00418] A directional coupler or a switch may be used to connect the port parameter measurement circuitry to the source resonator and tuning/adjustment circuitry. The port parameter measurement circuitry may measure properties of the magnetic resonator while it is exchanging power in a wireless power transmission system, or it may be switched out of the circuit during system operation. The port parameter measurement circuitry may measure the parameters and the processor may control certain tunable elements of the magnetic resonator at start-up, or at certain intervals, or in response to changes in certain system operating parameters.

[00419] A wireless power transmission system may include circuitry to vary or tune the impedance and/or resonant frequency of source and device resonators. Note that while tuning circuitry is shown in both the source and device resonators, the circuitry may instead be included in only the source or the device resonators, or the circuitry may be included in only some of the source and/or device resonators. Note too that while we may refer to the circuitry as "tuning" the impedance and or resonant frequency of the resonators, this tuning operation simply means that various electrical parameters such as the inductance or capacitance of the structure are being varied. In some cases, these parameters may be varied to achieve a specific predetermined value, in other cases they may be varied in response to a control algorithm or to stabilize a target performance value that is changing. In some cases, the parameters are varied as a function of temperature, of other sources or devices in the area, of the environment, at the like.

[00420] **Applications**

[00421] For each listed application, it will be understood by one of ordinary skill-in-the-art that there are a variety of ways that the resonator structures used to enable wireless power transmission may be connected or integrated with the objects that are supplying or being powered. The resonator may be physically separate from the source and device objects. The resonator may supply or remove power from an object using traditional inductive techniques or through direct electrical connection, with a wire or cable for example. The electrical connection may be from the resonator output to the AC or DC power input port on the object. The electrical connection may be from the output power port of an object to the resonator input.

[00422] FIG. 49 shows a source resonator 4904 that is physically separated from a power supply and a device resonator 4902 that is physically separated from the device 4900, in this illustration a laptop computer. Power may be supplied to the source resonator, and power may be taken from the device resonator directly, by an electrical connection. One of ordinary skill in the art will understand from the materials incorporated by reference that the shape, size, material composition, arrangement, position and orientation of the resonators above are provided by way of non-limiting example, and that a wide variation in any and all of these parameters could be supported by the disclosed technology for a variety of applications.

[00423] Continuing with the example of the laptop, and without limitation, the device resonator may be physically connected to the device it is powering or charging. For example, as shown in FIG. 50a and FIG. 50b, the device resonator 5002 may be (a) integrated into the housing of the device 5000 or (b) it may be attached by an adapter. The resonator 5002 may (FIG. 50b-d) or may not (FIG. 50a) be visible on the device. The resonator may be affixed to the device, integrated into the device, plugged into the device, and the like.

[00424] The source resonator may be physically connected to the source supplying the power to the system. As described above for the devices and device resonators, there are a variety of ways the resonators may be attached to, connected to or integrated with the power supply. One of ordinary skill in the art will understand that there are a variety of ways the resonators may be integrated in the wireless power transmission system, and that the sources and devices may utilize similar or different integration techniques.

[00425] Continuing again with the example of the laptop computer, and without limitation, the laptop computer may be powered, charged or recharged by a wireless power transmission system. A source resonator may be used to supply wireless power and a device resonator may be used to capture the wireless power. A device resonator 5002 may be integrated into the edge of the screen (display) as illustrated in FIG. 50d, and/or into the base of the laptop as illustrated in FIG. 50c. The source resonator 5002 may be integrated into the base of the laptop and the device resonator may be integrated into the edge of the screen. The resonators may also or instead be affixed to the power source and/or the laptop. The source and device resonators may also or instead be physically separated from the power supply and the laptop and may be electrically connected by a cable. The source and device resonators may also or instead be physically separated from the power supply and the laptop and may be electrically coupled using a traditional inductive technique. One of ordinary skill in the art will understand that, while the preceding examples relate to wireless power transmission to a laptop, that the methods and systems disclosed for this application may be suitably adapted for use with other electrical or electronic devices. In general, the source resonator may be external to the source and supplying power to a device resonator that in turn supplies power the device, or the source resonator may be connected to the source and supplying power to a device resonator that in turn supplies power to a portion of the device, or the source resonator may internal to the source and supplying power to a device resonator that in turn supplies power to a portion of the device, as well as any combination of these.

[00426] A system or method disclosed herein may provide power to an electrical or electronics device, such as, and not limited to, phones, cell phones, cordless phones, smart phones, PDAs, audio devices, music players, MP3 players, radios, portable radios and players, wireless headphones, wireless headsets, computers, laptop computers, wireless keyboards, wireless mouse, televisions, displays, flat screen displays, computer displays, displays embedded in furniture, digital picture frames, electronic books, (e.g. the Kindle, e-ink books, magazines, and the like), remote control units (also referred to as controllers, game controllers, commanders, clickers, and the like, and used for the remote control of a plurality of electronics devices, such as televisions, video games, displays, computers, audio visual equipment, lights, and the like), lighting devices, cooling devices, air circulation devices, purification devices, personal hearing aids, power tools, security systems, alarms, bells, flashing lights, sirens, sensors, loudspeakers, electronic locks, electronic keypads, light switches, other electrical switches, and the like. Here the term electronic lock is used to indicate a door lock which operates electronically (e.g. with electronic combo-key, magnetic card, RFID card, and the like) which is placed on a door instead of a mechanical key-lock. Such locks are often battery operated, risking the possibility that the lock might stop working when a battery dies, leaving the user locked-out. This may be avoided where the battery is either charged or completely replaced by a wireless power transmission implementation as described herein.

[00427] Here, the term light switch (or other electrical switch) is meant to indicate any switch (e.g. on a wall of a room) in one part of the room that turns on/off a device (e.g. light fixture at the center of the ceiling) in another part of the room. To install such a switch by direct connection, one would have to run a wire all the way from the device to the switch. Once such a switch is installed at a particular spot, it may be very difficult to move. Alternately, one can envision a ‘wireless switch’, where “wireless” means the switching (on/off) commands are communicated wirelessly, but such a switch has traditionally required a battery for operation. In general, having too many battery operated switches around a house may be impractical, because those many batteries will need to be replaced periodically. So, a wirelessly communicating switch may be more convenient, provided it is also wirelessly powered. For example, there already exist communications wireless door-bells that are battery powered, but where one still has to replace the battery in them periodically. The remote doorbell button may be made to be completely wireless, where there may be no need to ever replace the battery again. Note that here, the term 'cordless' or 'wireless' or 'communications wireless' is used to indicate that there is a cordless or wireless communications facility between the device and another electrical component, such as the base station for a cordless phone, the computer for a wireless keyboard, and the like. One skilled in the art will recognize that any electrical or electronics device may include a wireless communications facility, and that the systems and methods described herein may be used to add wireless power transmission to the device. As described herein, power to the electrical or electronics device may be delivered from an external or internal source resonator, and to the device or portion of the device. Wireless power transmission may significantly reduce the need to charge and/or replace batteries for devices that enter the near vicinity of the source resonator and thereby may reduce the downtime, cost and disposal issues often associated with batteries.

[00428] The systems and methods described herein may provide power to lights without the need for either wired power or batteries. That is, the systems and methods described herein may provide power to lights without wired connection to any power source, and provide the energy to the light non-radiatively across mid-range distances, such as across a distance of a quarter of a meter, one meter, three meters, and the like. A 'light' as used herein may refer to the light source itself, such as an incandescent light bulb, florescent light bulb lamps, Halogen lamps, gas discharge lamps, fluorescent lamps, neon lamps, high-intensity discharge lamps, sodium vapor lamps, Mercury-vapor lamps, electroluminescent lamps, light emitting diodes (LED) lamps, and the like; the light as part of a light fixture, such as a table lamp, a floor lamp, a ceiling lamp, track lighting, recessed light fixtures, and the like; light fixtures integrated with other functions, such as a light/ceiling fan fixture, and illuminated picture frame, and the like. As such, the systems and methods described herein may reduce the complexity for installing a light, such as by minimizing the installation of electrical wiring, and allowing the user to place or mount the light with minimal regard to sources of wired power. For instance, a light may be placed anywhere in the vicinity of a source resonator, where the source resonator may be mounted in a plurality of different places with respect to the location of the light, such as on the floor of the room above, (e.g. as in the case of a ceiling light and especially when the room above is the attic); on the wall of the next room, on the ceiling of the room below, (e.g. as in the case of a floor lamp); in a component within the room or in the infrastructure of the room as described herein; and the like. For example, a light/ceiling fan combination is often installed in a master bedroom, and the master bedroom often has the attic above it. In this instance a user may more easily install the light/ceiling fan combination in the master bedroom, such as by simply mounting the light/ceiling fan combination to the ceiling, and placing a source coil (plugged into the house wired AC power) in the attic above the mounted fixture. In another example, the light may be an external light, such as a flood light or security light, and the source resonator mounted inside the structure. This way of installing lighting may be particularly beneficial to users who rent their homes, because now they may be able to mount lights and such other electrical components without the need to install new electrical wiring. The control for the light may also be communicated by near-field communications as described herein, or by traditional wireless communications methods.

[00429] The systems and methods described herein may provide power from a source resonator to a device resonator that is either embedded into the device component, or outside the device component, such that the device component may be a traditional electrical component or fixture. For instance, a ceiling lamp may be designed or retrofitted with a device resonator integrated into the fixture, or the ceiling lamp may be a traditional wired fixture, and plugged into a separate electrical facility equipped with the device resonator. In an example, the electrical facility may be a wireless junction box designed to have a device resonator for receiving wireless power, say from a source resonator placed on the floor of the room above (e.g. the attic), and which contains a number of traditional outlets that are powered from the device resonator. The wireless junction box, mounted on the ceiling, may now provide power to traditional wired electrical components on the ceiling (e.g. a ceiling light, track lighting, a ceiling fan). Thus, the ceiling lamp may now be mounted to the ceiling without the need to run wires through the infrastructure of the building. This type of device resonator to traditional outlet junction box may be used in a plurality of applications, including being designed for the interior or exterior of a building, to be made portable, made for a vehicle, and the like. Wireless power may be transferred through common building materials, such as wood, wall board, insulation, glass. brick, stone, concrete, and the like. The benefits of reduced installation cost, re-configurability, and increased application flexibility may provide the user significant benefits over traditional wired installations. The device resonator for a traditional outlet junction box may include a plurality of electrical components for facilitating the transfer of power from the device resonator to the traditional outlets, such as power source electronics which convert the specific frequencies needed to implement efficient power transfer to line voltage, power capture electronics which may convert high frequency AC to usable voltage and frequencies (AC and/or DC), controls which synchronize the capture device and the power output and which ensure consistent, safe, and maximally efficient power transfer, and the like.

[00430] The systems and methods described herein may provide advantages to lights or electrical components that operate in environments that are wet, harsh, controlled, and the like, such has outside and exposed to the rain, in a pool/sauna/shower, in a maritime application, in hermetically sealed components, in an explosive-proof room, on outside signage, a harsh industrial environment in a volatile environment (e.g. from volatile vapors or airborne organics, such as in a grain silo or bakery) , and the like. For example, a light mounted under the water level of a pool is normally difficult to wire up, and is required to be water-sealed despite the need for external wires. But a pool light using the principles disclosed herein may more easily be made water sealed, as there may be no external wires needed. In another example, an explosion proof room, such as containing volatile vapors, may not only need to be hermetically sealed, but may need to have all electrical contacts (that could create a spark) sealed. Again, the principles disclosed herein may provide a convenient way to supply sealed electrical components for such applications.

[00431] The systems and methods disclosed herein may provide power to game controller applications, such as to a remote handheld game controller. These game controllers may have been traditionally powered solely by batteries, where the game controller's use and power profile caused frequent changing of the battery, battery pack, rechargeable batteries, and the like, that may not have been ideal for the consistent use to the game controller, such as during extended game play. A device resonator may be placed into the game controller, and a source resonator, connected to a power source, may be placed in the vicinity. Further, the device resonator in the game controller may provide power directly to the game controller electronics without a battery; provide power to a battery, battery pack, rechargeable battery, and the like, which then provides power to the game controller electronics; and the like. The game controller may utilize multiple battery packs, where each battery pack is equipped with a device resonator, and thus may be constantly recharging while in the vicinity of the source resonator, whether plugged into the game controller or not. The source resonator may be resident in a main game controller facility for the game, where the main game controller facility and source resonator are supplied power from AC 'house' power; resident in an extension facility form AC power, such as in a source resonator integrated into an 'extension cord'; resident in a game chair, which is at least one of plugged into the wall AC, plugged into the main game controller facility, powered by a battery pack in the game chair; and the like. The source resonator may be placed and implemented in any of the configurations described herein.

[00432] The systems and methods disclosed herein may integrate device resonators into battery packs, such as battery packs that are interchangeable with other battery packs. For instance, some portable devices may use up electrical energy at a high rate such that a user may need to have multiple interchangeable battery packs on hand for use, or the user may operate the device out of range of a source resonator and need additional battery packs to continue operation, such as for power tools, portable lights, remote control vehicles, and the like. The use of the principles disclosed herein may not only provide a way for device resonator enabled battery packs to be recharged while in use and in range, but also for the recharging of battery packs not currently in use and placed in range of a source resonator. In this way, battery packs may always be ready to use when a user runs down the charge of a battery pack being used. For example, a user may be working with a wireless power tool, where the current requirements may be greater than can be realized through direct powering from a source resonator. In this case, despite the fact that the systems and methods described herein may be providing charging power to the in-use battery pack while in range, the battery pack may still run down, as the power usage may have exceeded the recharge rate. Further, the user may simply be moving in and out of range, or be completely out of range while using the device. However, the user may have placed additional battery packs in the vicinity of the source resonator, which have been recharged while not in use, and are now charged sufficiently for use. In another example, the user may be working with the power tool away from the vicinity of the source resonator, but leave the supplemental battery packs to charge in the vicinity of the source resonator, such as in a room with a portable source resonator or extension cord source resonator, in the user's vehicle, in user's tool box, and the like. In this way, the user may not have to worry about taking the time to, and/or remembering to plug in their battery packs for future use. The user may only have to change out the used battery pack for the charged battery pack and place the used one in the vicinity of the source resonator for recharging. Device resonators may be built into enclosures with known battery form factors and footprints and may replace traditional chemical batteries in known devices and applications. For example, device resonators may be built into enclosures with mechanical dimensions equivalent to AA batteries, AAA batteries, D batteries, 9V batteries, laptop batteries, cell phone batteries, and the like. The enclosures may include a smaller "button battery" in addition to the device resonator to store charge and provide extended operation, either in terms of time or distance. Other energy storage devices in addition to or instead of button batteries may be integrated with the device resonators and any associated power conversion circuitry. These new energy packs may provide similar voltage and current levels as provided by traditional batteries, but may be composed of device resonators, power conversion electronics, a small battery, and the like. These new energy packs may last longer than traditional batteries because they may be more easily recharged and may be recharging constantly when they are located in a wireless power zone. In addition, such energy packs may be lighter than traditional batteries, may be safer to use and store, may operate over wider temperature and humidity ranges, may be less harmful to the environment when thrown away, and the like. As described herein, these energy packs may last beyond the life of the product when used in wireless power zones as described herein.

[00433] The systems and methods described herein may be used to power visual displays, such as in the case of the laptop screen, but more generally to include the great variety and diversity of displays utilized in today's electrical and electronics components, such as in televisions, computer monitors, desktop monitors, laptop displays, digital photo frames, electronic books, mobile device displays (e.g. on phones, PDAs, games, navigation devices, DVD players), and the like. Displays that may be powered through one or more of the wireless power transmission systems described herein may also include embedded displays, such as embedded in electronic components (e.g. audio equipment, home appliances, automotive displays, entertainment devices, cash registers, remote controls), in furniture, in building infrastructure, in a vehicle, on the surface of an object (e.g. on the surface of a vehicle, building, clothing, signs, transportation), and the like. Displays may be very small with tiny resonant devices, such as in a smart card as described herein, or very large, such as in an advertisement sign. Displays powered using the principles disclosed herein may also be any one of a plurality of imaging technologies, such as liquid crystal display (LCD), thin film transistor LCD, passive LCD, cathode ray tube (CRT), plasma display, projector display (e.g. LCD, DLP, LCOS), surface-conduction electron-emitter display (SED), organic light-emitting diode (OLED), and the like. Source coil configurations may include attaching to a primary power source, such as building power, vehicle power, from a wireless extension cord as described herein, and the like; attached to component power, such as the base of an electrical component (e.g. the base of a computer, a cable box for a TV); an intermediate relay source coil; and the like. For example, hanging a digital display on the wall may be very appealing, such as in the case of a digital photo frame that receives its information signals wirelessly or through a portable memory device, but the need for an unsightly power cord may make it aesthetically unpleasant. However, with a device coil embedded in the digital photo frame, such as wrapped within the frame portion, may allow the digital photo frame to be hung with no wires at all. The source resonator may then be placed in the vicinity of the digital photo frame, such as in the next room on the other side of the wall, plugged directly into a traditional power outlet, from a wireless extension cord as described herein, from a central source resonator for the room, and the like.

[00434] The systems and methods described herein may provide wireless power transmission between different portions of an electronics facility. Continuing with the example of the laptop computer, and without limitation, the screen of the laptop computer may require power from the base of the laptop. In this instance, the electrical power has been traditionally routed via direct electrical connection from the base of the laptop to the screen over a hinged portion of the laptop between the screen and the base. When a wired connection is utilized, the wired connection may tend to wear out and break, the design functionality of the laptop computer may be limited by the required direct electrical connection, the design aesthetics of the laptop computer may be limited by the required direct electrical connection, and the like. However, a wireless connection may be made between the base and the screen. In this instance, the device resonator may be placed in the screen portion to power the display, and the base may be either powered by a second device resonator, by traditional wired connections, by a hybrid of resonator-battery- direct electrical connection, and the like. This may not only improve the reliability of the power connection due to the removal of the physical wired connection, but may also allow designers to improve the functional and/or aesthetic design of the hinge portion of the laptop in light of the absence of physical wires associated with the hinge. Again, the laptop computer has been used here to illustrate how the principles disclosed herein may improve the design of an electric or electronic device, and should not be taken as limiting in any way. For instance, many other electrical devices with separated physical portions could benefit from the systems and methods described herein, such as a refrigerator with electrical functions on the door, including an ice maker, a sensor system, a light, and the like; a robot with movable portions, separated by joints; a car's power system and a component in the car's door; and the like. The ability to provide power to a device via a device resonator from an external source resonator, or to a portion of the device via a device resonator from either external or internal source resonators, will be recognized by someone skilled in the art to be widely applicable across the range of electric and electronic devices.

[00435] The systems and methods disclosed herein may provide for a sharing of electrical power between devices, such as between charged devices and uncharged devices. For instance a charged up device or appliance may act like a source and send a predetermined amount of energy, dialed in amount of energy, requested and approved amount of energy, and the like, to a nearby device or appliance. For example, a user may have a cell phone and a digital camera that are both capable of transmitting and receiving power through embedded source and device resonators, and one of the devices, say the cell phone, is found to be low on charge. The user may then transfer charge from the digital camera to the cell phone. The source and device resonators in these devices may utilize the same physical resonator for both transmission and reception, utilize separate source and device resonators, one device may be designed to receive and transmit while the other is designed to receive only, one device may be designed to transmit only and the other to receive only, and the like.

[00436] To prevent complete draining the battery of a device it may have a setting allowing a user to specify how much of the power resource the receiving device is entitled to. It may be useful, for example, to put a limit on the amount of power available to external devices and to have the ability to shut down power transmission when battery power falls below a threshold.

[00437] The systems and methods described herein may provide wireless power transfer to a nearby electrical or electronics component in association with an electrical facility, where the source resonator is in the electrical facility and the device resonator is in the electronics component. The source resonator may also be connected to, plugged into, attached to the electrical facility, such as through a universal interface (e.g. a USB interface, PC card interface), supplemental electrical outlet, universal attachment point, and the like, of the electrical facility. For example, the source resonator may be inside the structure of a computer on a desk, or be integrated into some object, pad, and the like, that is connected to the computer, such as into one of the computer's USB interfaces. In the example of the source resonator embedded in the object, pad, and the like, and powered through a USB interface, the source resonator may then be easily added to a user's desktop without the need for being integrated into any other electronics device, thus conveniently providing a wireless energy zone around which a plurality of electric and/or electronics devices may be powered. The electrical facility may be a computer, a light fixture, a dedicated source resonator electrical facility, and the like, and the nearby components may be computer peripherals, surrounding electronics components, infrastructure devices, and the like, such as computer keyboards, computer mouse, fax machine, printer, speaker system, cell phone, audio device, intercom, music player, PDA, lights, electric pencil sharpener, fan, digital picture frame, calculator, electronic games, and the like. For example, a computer system may be the electrical facility with an integrated source resonator that utilizes a 'wireless keyboard' and 'wireless mouse', where the use of the term wireless here is meant to indicate that there is wireless communication facility between each device and the computer, and where each device must still contain a separate battery power source. As a result, batteries would need to be replaced periodically, and in a large company, may result in a substantial burden for support personnel for replacement of batteries, cost of batteries, and proper disposal of batteries. Alternatively, the systems and methods described herein may provide wireless power transmission from the main body of the computer to each of these peripheral devices, including not only power to the keyboard and mouse, but to other peripheral components such as a fax, printer, speaker system, and the like, as described herein. A source resonator integrated into the electrical facility may provide wireless power transmission to a plurality of peripheral devices, user devices, and the like, such that there is a significant reduction in the need to charge and/or replace batteries for devices in the near vicinity of the source resonator integrated electrical facility. The electrical facility may also provide tuning or auto-tuning software, algorithms, facilities, and the like, for adjusting the power transfer parameters between the electrical facility and the wirelessly powered device. For example, the electrical facility may be a computer on a user's desktop, and the source resonator may be either integrated into the computer or plugged into the computer (e.g. through a USB connection), where the computer provides a facility for providing the tuning algorithm (e.g. through a software program running on the computer).

[00438] The systems and methods disclosed herein may provide wireless power transfer to a nearby electrical or electronics component in association with a facility infrastructure component, where the source resonator is in, or mounted on, the facility infrastructure component and the device resonator is in the electronics component. For instance, the facility infrastructure component may be a piece of furniture, a fixed wall, a movable wall or partition, the ceiling, the floor, and the source resonator attached or integrated into a table or desk (e.g. just below/above the surface, on the side, integrated into a table top or table leg), a mat placed on the floor (e.g. below a desk, placed on a desk), a mat on the garage floor (e.g. to charge the car and/or devices in the car), in a parking lot/garage (e.g. on a post near where the car is parked), a television (e.g. for charging a remote control), a computer monitor (e.g. to power/charge a wireless keyboard, wireless mouse, cell phone), a chair (e.g. for powering electric blankets, medical devices, personal health monitors), a painting, office furniture, common household appliances, and the like. For example, the facility infrastructure component may be a lighting fixture in an office cubical, where the source resonator and light within the lighting fixture are both directly connected to the facility's wired electrical power. However, with the source resonator now provided in the lighting fixture, there would be no need to have any additional wired connections for those nearby electrical or electronics components that are connected to, or integrated with, a device resonator. In addition, there may be a reduced need for the replacement of batteries for devices with device resonators, as described herein.

[00439] The use of the systems and methods described herein to supply power to electrical and electronic devices from a central location, such as from a source resonator in an electrical facility, from a facility infrastructure component and the like, may minimize the electrical wiring infrastructure of the surrounding work area. For example, in an enterprise office space there are typically a great number of electrical and electronic devices that need to be powered by wired connections. With utilization of the systems and methods described herein, much of this wiring may be eliminated, saving the enterprise the cost of installation, decreasing the physical limitations associated with office walls having electrical wiring, minimizing the need for power outlets and power strips, and the like. The systems and methods described herein may save money for the enterprise through a reduction in electrical infrastructure associated with installation, re-installation (e.g., reconfiguring office space), maintenance, and the like. In another example, the principles disclosed herein may allow the wireless placement of an electrical outlet in the middle of a room. Here, the source could be placed on the ceiling of a basement below the location on the floor above where one desires to put an outlet. The device resonator could be placed on the floor of the room right above it. Installing a new lighting fixture (or any other electric device for that matter, e.g. camera, sensor, etc., in the center of the ceiling may now be substantially easier for the same reason).

[00440] In another example, the systems and methods described herein may provide power "through" walls. For instance, suppose one has an electric outlet in one room (e.g. on a wall), but one would like to have an outlet in the next room, but without the need to call an electrician, or drill through a wall, or drag a wire around the wall, or the like. Then one might put a source resonator on the wall in one room, and a device resonator outlet/pickup on the other side of the wall. This may power a flat-screen TV or stereo system or the like (e.g. one may not want to have an ugly wire climbing up the wall in the living room, but doesn't mind having a similar wire going up the wall in the next room, e.g. storage room or closet, or a room with furniture that blocks view of wires running along the wall). The systems and methods described herein may be used to transfer power from an indoor source to various electric devices outside of homes or buildings without requiring holes to be drilled through, or conduits installed in, these outside walls. In this case, devices could be wirelessly powered outside the building without the aesthetic or structural damage or risks associated with drilling holes through walls and siding. In addition, the systems and methods described herein may provide for a placement sensor to assist in placing an interior source resonator for an exterior device resonator equipped electrical component. For example, a home owner may place a security light on the outside of their home which includes a wireless device resonator, and now needs to adequately or optimally position the source resonator inside the home. A placement sensor acting between the source and device resonators may better enable that placement by indicating when placement is good, or to a degree of good, such as in a visual indication, an audio indication, a display indication, and the like. In another example, and in a similar way, the systems and methods described herein may provide for the installation of equipment on the roof of a home or building, such as radio transmitters and receivers, solar panels and the like. In the case of the solar panel, the source resonator may be associated with the panel, and power may be wirelessly transferred to a distribution panel inside the building without the need for drilling through the roof. The systems and methods described herein may allow for the mounting of electric or electrical components across the walls of vehicles (such as through the roof) without the need to drill holes, such as for automobiles, water craft, planes, trains, and the like. In this way, the vehicle's walls may be left intact without holes being drilled, thus maintaining the value of the vehicle, maintaining watertightness, eliminating the need to route wires, and the like. For example, mounting a siren or light to the roof of a police car decreases the future resale of the car, but with the systems and methods described herein, any light, horn, siren, and the like, may be attached to the roof without the need to drill a hole.

[00441] The systems and methods described herein may be used for wireless transfer of power from solar photovoltaic (PV) panels. PV panels with wireless power transfer capability may have several benefits including simpler installation, more flexible, reliable, and weatherproof design. Wireless power transfer may be used to transfer power from the PV panels to a device, house, vehicle, and the like. Solar PV panels may have a wireless source resonator allowing the PV panel to directly power a device that is enabled to receive the wireless power. For example, a solar PV panel may be mounted directly onto the roof of a vehicle, building, and the like. The energy captured by the PV panel may be wirelessly transferred directly to devices inside the vehicle or under the roof of a building. Devices that have resonators can wirelessly receive power from the PV panel. Wireless power transfer from PV panels may be used to transfer energy to a resonator that is coupled to the wired electrical system of a house, vehicle, and the like allowing traditional power distribution and powering of conventional devices without requiring any direct contact between the exterior PV panels and the internal electrical system.

[00442] With wireless power transfer significantly simpler installation of rooftop PV panels is possible because power may be transmitted wirelessly from the panel to a capture resonator in the house, eliminating all outside wiring, connectors, and conduits, and any holes through the roof or walls of the structure. Wireless power transfer used with solar cells may have a benefit in that it can reduced roof danger since it eliminates the need for electricians to work on the roof to interconnect panels, strings, and junction boxes. Installation of solar panels integrated with wireless power transfer may require less skilled labor since fewer electrical contacts need to be made. Less site specific design may be required with wireless power transfer since the technology gives the installer the ability to individually optimize and position each solar PV panel, significantly reducing the need for expensive engineering and panel layout services. There may not be need to carefully balance the solar load on every panel and no need for specialized DC wiring layout and interconnections.

[00443] For rooftop or on-wall installations of PV panels, the capture resonator may be mounted on the underside of the roof, inside the wall, or in any other easily accessible inside space within a foot or two of the solar PV panel. A diagram showing a possible general rooftop PV panel installation is shown in Figure 51. Various PV solar collectors may be mounted in top of a roof with wireless power capture coils mounted inside the building under the roof. The resonator coils in the PV panels can transfer their energy wirelessly through the roof to the wireless capture coils. The captured energy from the PV cells may be collected and coupled to the electrical system of the house to power electric and electronic devices or coupled to the power grid when more power than needed is generated. Energy is captured from the PV cells without requiring holes or wires that penetrate the roof or the walls of the building. Each PV panel may have a resonator that is coupled to a corresponding resonator on the interior of the vehicle or building. Multiple panels may utilize wireless power transfer between each other to transfer or collect power to one or a couple of designated panels that are coupled to resonators on the interior of the vehicle of house. Panels may have wireless power resonators on their sides or in their perimeter that can couple to resonators located in other like panels allowing transfer of power from panel to panel. An additional bus or connection structure may be provided that wirelessly couples the power from multiple panels on the exterior of a building or vehicle and transfers power to one or a more resonators on the interior of building or vehicle.

[00444] For example, as shown in Fig. 51, a source resonator 5102 may be coupled to a PV cell 5100 mounted on top of roof 5104 of a building. A corresponding capture resonator 5106 is placed inside the building. The solar energy captured by the PV cells can then be transferred between the source resonators 5102 outside to the device resonators 5106 inside the building without having direct holes and connections through the building.

[00445] Each solar PV panel with wireless power transfer may have its own inverter, significantly improving the economics of these solar systems by individually optimizing the power production efficiency of each panel, supporting a mix of panel sizes and types in a single installation, including single panel “pay-as-you-grow” system expansions. Reduction of installation costs may make a single panel economical for installation. Eliminating the need for panel string designs and careful positioning and orienting of multiple panels, and eliminating a single point of failure for the system.

[00446] Wireless power transfer in PV solar panels may enable more solar deployment scenarios because the weather-sealed solar PV panels eliminate the need to drill holes for wiring through sealed surfaces such as car roofs and ship decks, and eliminate the requirement that the panels be installed in fixed locations. With wireless power transfer, PV panels may be deployed temporarily, and then moved or removed, without leaving behind permanent alterations to the surrounding structures. They may be placed out in a yard on sunny days, and moved around to follow the sun, or brought inside for cleaning or storage, for example. For backyard or mobile solar PV applications, an extension cord with a wireless energy capture device may be thrown on the ground or placed near the solar unit. The capture extension cord can be completely sealed from the elements and electrically isolated, so that it may be used in any indoor or outdoor environment.

[00447] With wireless power transfer no wires or external connections may be necessary and the PV solar panels can be completely weather sealed. Significantly improved reliability and lifetime of electrical components in the solar PV power generation and transmission circuitry can be expected since the weather-sealed enclosures can protect components from UV radiation, humidity, weather, and the like. With wireless power transfer and weather-sealed enclosures it may be possible to use less expensive components since they will no longer be directly exposed to external factors and weather elements and it may reduce the cost of PV panels.

[00448] Power transfer between the PV panels and the capture resonators inside a building or a vehicle may be bidirectional. Energy may be transmitted from the house grid to the PV panels to provide power when the panels do not have enough energy to perform certain tasks such. Reverse power flow can be used to melt snow from the panels, or power motors that will position the panels in a more favorable positions with respect to the sun energy. Once the snow is melted or the panels are repositioned and the PV panels can generate their own energy the direction of power transfer can be returned to normal delivering power from the PV panels to buildings, vehicles, or devices.

[00449] PV panels with wireless power transfer may include auto-tuning on installation to ensure maximum and efficient power transfer to the wireless collector. Variations in roofing materials or variations in distances between the PV panels and the wireless power collector in different installations may affect the performance or perturb the properties of the resonators of the wireless power transfer. To reduce the installation complexity the wireless power transfer components may include a tuning capability to automatically adjust their operating point to compensate for any effects due to materials or distance. Frequency, impedance, capacitance, inductance, duty cycle, voltage levels and the like may be adjusted to ensure efficient and safe power transfer

[00450] The systems and methods described herein may be used to provide a wireless power zone on a temporary basis or in extension of traditional electrical outlets to wireless power zones, such as through the use of a wireless power extension cord. For example, a wireless power extension cord may be configured as a plug for connecting into a traditional power outlet, a long wire such as in a traditional power extension cord, and a resonant source coil on the other end (e.g. in place of, or in addition to, the traditional socket end of the extension The wireless extension cord may also be configured where there are source resonators at a plurality of locations along the wireless extension cord. This configuration may then replace any traditional extension cord where there are wireless power configured devices, such as providing wireless power to a location where there is no convenient power outlet (e.g. a location in the living room where there's no outlet), for temporary wireless power where there is no wired power infrastructure (e.g. a construction site), out into the yard where there are no outlets (e.g. for parties or for yard grooming equipment that is wirelessly powered to decrease the chances of cutting the traditional electrical cord), and the like. The wireless extension cord may also be used as a drop within a wall or structure to provide wireless power zones within the vicinity of the drop. For example, a wireless extension cord could be run within a wall of a new or renovated room to provide wireless power zones without the need for the installation of traditional electrical wiring and outlets.

[00451] The systems and methods described herein may be utilized to provide power between moving parts or rotating assemblies of a vehicle, a robot, a mechanical device, a wind turbine, or any other type of rotating device or structure with moving parts such as robot arms, construction vehicles, movable platforms and the like. Traditionally, power in such systems may have been provided by slip rings or by rotary joints for example. Using wireless power transfer as described herein, the design simplicity, reliability and longevity of these devices may be significantly improved because power can be transferred over a range of distances without any physical connections or contact points that may wear down or out with time. In particular, the preferred coaxial and parallel alignment of the source and device coils may provide wireless power transmission that is not severely modulated by the relative rotational motion of the two coils.

[00452] The systems and methods described herein may be utilized to extend power needs beyond the reach of a single source resonator by providing a series of source-device-source-device resonators. For instance, suppose an existing detached garage has no electrical power and the owner now wants to install a new power service. However, the owner may not want to run wires all over the garage, or have to break into the walls to wire electrical outlets throughout the structure. In this instance, the owner may elect to connect a source resonator to the new power service, enabling wireless power to be supplied to device resonator outlets throughout the back of the garage. The owner may then install a device-source 'relay' to supply wireless power to device resonator outlets in the front of the garage. That is, the power relay may now receive wireless power from the primary source resonator, and then supply available power to a second source resonator to supply power to a second set of device resonators in the front of the garage. This configuration may be repeated again and again to extend the effective range of the supplied wireless power.

[00453] Multiple resonators may be used to extend power needs around an energy blocking material. For instance, it may be desirable to integrate a source resonator into a computer or computer monitor such that the resonator may power devices placed around and especially in front of the monitor or computer such as keyboards, computer mice, telephones , and the like. Due to aesthetics, space constraints, and the like an energy source that may be used for the source resonator may only be located or connected to in the back of the monitor or computer. In many designs of computer or monitors metal components and metal containing circuits are used in the design and packaging which may limit and prevent power transfer from source resonator in the back of the monitor or computer to the front of the monitor or computer. An additional repeater resonator may be integrated into the base or pedestal of the monitor or computer that couples to the source resonator in the back of the monitor or computer and allows power transfer to the space in front of the monitor or computer. The intermediate resonator integrated into the base or pedestal of the monitor or computer does not require an additional power source, it captures power from the source resonator and transfers power to the front around the blocking or power shielding metal components of the monitor or computer.

[00454] The systems and methods described herein may be built-into, placed on, hung from, embedded into, integrated into, and the like, the structural portions of a space, such as a vehicle, office, home, room, building, outdoor structure, road infrastructure, and the like. For instance, one or more sources may be built into, placed on, hung from, embedded or integrated into a wall, a ceiling or ceiling panel, a floor, a divider, a doorway, a stairwell, a compartment, a road surface, a sidewalk, a ramp, a fence, an exterior structure, and the like. One or more sources may be built into an entity within or around a structure, for instance a bed, a desk, a chair, a rug, a mirror, a clock, a display, a television, an electronic device, a counter, a table, a piece of furniture, a piece of artwork, an enclosure, a compartment, a ceiling panel, a floor or door panel, a dashboard, a trunk, a wheel well, a post, a beam, a support or any like entity. For example, a source resonator may be integrated into the dashboard of a user's car so that any device that is equipped with or connected to a device resonator may be supplied with power from the dashboard source resonator. In this way, devices brought into or integrated into the car may be constantly charged or powered while in the car.

[00455] The systems and methods described herein may provide power through the walls of vehicles, such as boats, cars, trucks, busses, trains, planes, satellites and the like. For instance, a user may not want to drill through the wall of the vehicle in order to provide power to an electric device on the outside of the vehicle. A source resonator may be placed inside the vehicle and a device resonator may be placed outside the vehicle (e.g. on the opposite side of a window, wall or structure). In this way the user may achieve greater flexibility in optimizing the placement, positioning and attachment of the external device to the vehicle, (such as without regard to supplying or routing electrical connections to the device). In addition, with the electrical power supplied wirelessly, the external device may be sealed such that it is water tight, making it safe if the electric device is exposed to weather (e.g. rain), or even submerged under water. Similar techniques may be employed in a variety of applications, such as in charging or powering hybrid vehicles, navigation and communications equipment, construction equipment, remote controlled or robotic equipment and the like, where electrical risks exist because of exposed conductors. The systems and methods described herein may provide power through the walls of vacuum chambers or other enclosed spaces such as those used in semiconductor growth and processing, material coating systems, aquariums, hazardous materials handling systems and the like. Power may be provided to translation stages, robotic arms, rotating stages, manipulation and collection devices, cleaning devices and the like.

[00456] The systems and methods described herein may provide wireless power to a kitchen environment, such as to counter-top appliances, including mixers, coffee makers, toasters, toaster ovens, grills, griddles, electric skillets, electric pots, electric woks, waffle makers, blenders, food processors, crock pots, warming trays, induction cooktops, lights, computers, displays, and the like. This technology may improve the mobility and/or positioning flexibility of devices, reduce the number of power cords stored on and strewn across the counter-top, improve the washability of the devices, and the like. For example, an electric skillet may traditionally have separate portions, such as one that is submersible for washing and one that is not submersible because it includes an external electrical connection (e.g. a cord or a socket for a removable cord). However, with a device resonator integrated into the unit, all electrical connections may be sealed, and so the entire device may now be submersed for cleaning. In addition, the absence of an external cord may eliminate the need for an available electrical wall outlet, and there is no longer a need for a power cord to be placed across the counter or for the location of the electric griddle to be limited to the location of an available electrical wall outlet.

[00457] The systems and methods described herein may provide continuous power/charging to devices equipped with a device resonator because the device doesn't leave the proximity of a source resonator, such as fixed electrical devices, personal computers, intercom systems, security systems, household robots, lighting, remote control units, televisions, cordless phones, and the like. For example, a household robot (e.g. ROOMBA) could be powered/charged via wireless power, and thus work arbitrarily long without recharging. In this way, the power supply design for the household robot may be changed to take advantage of this continuous source of wireless power, such as to design the robot to only use power from the source resonator without the need for batteries, use power from the source resonator to recharge the robot's batteries, use the power from the source resonator to trickle charge the robot's batteries, use the power from the source resonator to charge a capacitive energy storage unit, and the like. Similar optimizations of the power supplies and power circuits may be enabled, designed, and realized, for any and all of the devices disclosed herein.

[00458] The systems and methods described herein may be able to provide wireless power to electrically heated blankets, heating pads/patches, and the like. These electrically heated devices may find a variety of indoor and outdoor uses. For example, hand and foot warmers supplied to outdoor workers such as guards, policemen, construction workers and the like might be remotely powered from a source resonator associated with or built into a nearby vehicle, building, utility pole, traffic light, portable power unit, and the like.

[00459] The systems and methods described herein may be used to power a portable information device that contains a device resonator and that may be powered up when the information device is near an information source containing a source resonator. For instance, the information device may be a card (e.g. credit card, smart card, electronic card, and the like) carried in a user's pocket, wallet, purse, vehicle, bike, and the like. The portable information device may be powered up when it is in the vicinity of an information source that then transmits information to the portable information device that may contain electronic logic, electronic processors, memory, a display, an LCD display, LEDs, RFID tags, and the like. For example, the portable information device may be a credit card with a display that "turns on" when it is near an information source, and provide the user with some information such as, "You just received a coupon for 50% off your next Coca Cola purchase". The information device may store information such as coupon or discount information that could be used on subsequent purchases. The portable information device may be programmed by the user to contain tasks, calendar appointments, to-do lists, alarms and reminders, and the like. The information device may receive up-to-date price information and inform the user of the location and price of previously selected or identified items.

[00460] The systems and methods described herein may provide wireless power transmission to directly power or recharge the batteries in sensors, such as environmental sensors, security sensors, agriculture sensors, appliance sensors, food spoilage sensors, power sensors, and the like, which may be mounted internal to a structure, external to a structure, buried underground, installed in walls, and the like. For example, this capability may replace the need to dig out old sensors to physically replace the battery, or to bury a new sensor because the old sensor is out of power and no longer operational. These sensors may be charged up periodically through the use of a portable sensor source resonator charging unit. For instance, a truck carrying a source resonator equipped power source, say providing ~kW of power, may provide enough power to a ~mW sensor in a few minutes to extend the duration of operation of the sensor for more than a year. Sensors may also be directly powered, such as powering sensors that are in places where it is difficult to connect to them with a wire but they are still within the vicinity of a source resonator, such as devices outside of a house (security camera), on the other side of a wall, on an electric lock on a door, and the like. In another example, sensors that may need to be otherwise supplied with a wired power connection may be powered through the systems and methods described herein. For example, a ground fault interrupter breaker combines residual current and over-current protection in one device for installation into a service panel. However, the sensor traditionally has to be independently wired for power, and this may complicate the installation. However, with the systems and methods described herein the sensor may be powered with a device resonator, where a single source resonator is provided within the service panel, thus simplifying the installation and wiring configuration within the service panel. In addition, the single source resonator may power device resonators mounted on either side of the source resonator mounted within the service panel, throughout the service panel, to additional nearby service panels, and the like. The systems and methods described herein may be employed to provide wireless power to any electrical component associated with electrical panels, electrical rooms, power distribution and the like, such as in electric switchboards, distribution boards, circuit breakers, transformers, backup batteries, fire alarm control panels, and the like. Through the use of the systems and methods described herein, it may be easier to install, maintain, and modify electrical distribution and protection components and system installations.

[00461] In another example, sensors that are powered by batteries may run continuously, without the need to change the batteries, because wireless power may be supplied to periodically or continuously recharge or trickle charge the battery. In such applications, even low levels of power may adequately recharge or maintain the charge in batteries, significantly extending their lifetime and usefulness. In some cases, the battery life may be extended to be longer than the lifetime of the device it is powering, making it essentially a battery that "lasts forever".

[00462] The systems and methods described herein may be used for charging implanted medical device batteries, such as in an artificial heart, pacemaker, heart pump, insulin pump, implanted coils for nerve or acupressure/acupuncture point stimulation, and the like. For instance, it may not be convenient or safe to have wires sticking out of a patient because the wires may be a constant source of possible infection and may generally be very unpleasant for the patient. The systems and methods described herein may also be used to charge or power medical devices in or on a patient from an external source, such as from a bed or a hospital wall or ceiling with a source resonator. Such medical devices may be easier to attach, read, use and monitor the patient. The systems and methods described herein may ease the need for attaching wires to the patient and the patient's bed or bedside, making it more convenient for the patient to move around and get up out of bed without the risk of inadvertently disconnecting a medical device. This may, for example, be usefully employed with patients that have multiple sensors monitoring them, such as for measuring pulse, blood pressure, glucose, and the like. For medical and monitoring devices that utilize batteries, the batteries may need to be replaced quite often, perhaps multiple times a week. This may present risks associated with people forgetting to replace batteries, not noticing that the devices or monitors are not working because the batteries have died, infection associated with improper cleaning of the battery covers and compartments, and the like.

[00463] The systems and methods described herein may reduce the risk and complexity of medical device implantation procedures. Today many implantable medical devices such as ventricular assist devices, pacemakers, defibrillators and the like, require surgical implantation due to their device form factor, which is heavily influenced by the volume and shape of the long-life battery that is integrated in the device. In one aspect, there is described herein a non-invasive method of recharging the batteries so that the battery size may be dramatically reduced, and the entire device may be implanted, such as via a catheter. A catheter implantable device may include an integrated capture or device coil. A catheter implantable capture or device coil may be designed so that it may be wired internally, such as after implantation. The capture or device coil may be deployed via a catheter as a rolled up flexible coil (e.g. rolled up like two scrolls, easily unrolled internally with a simple spreader mechanism). The power source coil may be worn in a vest or article of clothing that is tailored to fit in such a way that places the source in proper position, may be placed in a chair cushion or bed cushion, may be integrated into a bed or piece of furniture, and the like.

[00464] The systems and methods described herein may enable patients to have a 'sensor vest', sensor patch, and the like, that may include at least one of a plurality of medical sensors and a device resonator that may be powered or charged when it is in the vicinity of a source resonator. Traditionally, this type of medical monitoring facility may have required batteries, thus making the vest, patch, and the like, heavy, and potentially impractical. But using the principles disclosed herein, no batteries (or a lighter rechargeable battery) may be required, thus making such a device more convenient and practical, especially in the case where such a medical device could be held in place without straps, such as by adhesive, in the absence of batteries or with substantially lighter batteries. A medical facility may be able to read the sensor data remotely with the aim of anticipating (e.g. a few minutes ahead of) a stroke, a heart-attack, or the like. When the vest is used by a person in a location remote from the medical facility, such as in their home, the vest may then be integrated with a cell-phone or communications device to call an ambulance in case of an accident or a medical event. The systems and methods described herein may be of particular value in the instance when the vest is to be used by an elderly person, where traditional non-wireless recharging practices (e.g. replacing batteries, plugging in at night, and the like) may not be followed as required. The systems and methods described herein may also be used for charging devices that are used by or that aid handicapped or disabled people who may have difficulty replacing or recharging batteries, or reliably supplying power to devices they enjoy or rely on.

[00465] The systems and methods described herein may be used for the charging and powering of artificial limbs. Artificial limbs have become very capable in terms of replacing the functionality of original limbs, such as arms, legs, hands and feet. However, an electrically powered artificial limb may require substantial power, (such as 10-20W) which may translate into a substantial battery. In that case, the amputee may be left with a choice between a light battery that doesn't last very long, and a heavy battery that lasts much longer, but is more difficult to 'carry' around. The systems and methods described herein may enable the artificial limb to be powered with a device resonator, where the source resonator is either carried by the user and attached to a part of the body that may more easily support the weight (such as on a belt around the waist, for example) or located in an external location where the user will spend an adequate amount of time to keep the device charged or powered, such as at their desk, in their car, in their bed, and the like.

[00466] The systems and methods described herein may be used for charging and powering of electrically powered exo-skeletons, such as those used in industrial and military applications, and for elderly/weak/sick people. An electrically powered exo-skeleton may provide up to a 10-to-20 times increase in "strength" to a person, enabling the person to perform physically strenuous tasks repeatedly without much fatigue. However, exo-skeletons may require more than 100W of power under certain use scenarios, so battery powered operation may be limited to 30 minutes or less. The delivery of wireless power as described herein may provide a user of an exo-skeleton with a continuous supply of power both for powering the structural movements of the exo-skeleton and for powering various monitors and sensors distributed throughout the structure. For instance, an exo-skeleton with an embedded device resonator(s) may be supplied with power from a local source resonator. For an industrial exo-skeleton, the source resonator may be placed in the walls of the facility. For a military exo-skeleton, the source resonator may be carried by an armored vehicle. For an exo-skeleton employed to assist a caretaker of the elderly, the source resonator(s) may be installed or placed in or the room(s) of a person's home.

[00467] The systems and methods described herein may be used for the powering/charging of portable medical equipment, such as oxygen systems, ventilators, defibrillators, medication pumps, monitors, and equipment in ambulances or mobile medical units, and the like. Being able to transport a patient from an accident scene to the hospital, or to move patients in their beds to other rooms or areas, and bring all the equipment that is attached with them and have it powered the whole time offers great benefits to the patients' health and eventual well-being. Certainly one can understand the risks and problems caused by medical devices that stop working because their battery dies or because they must be unplugged while a patient is transported or moved in any way. For example, an emergency medical team on the scene of an automotive accident might need to utilize portable medical equipment in the emergency care of patients in the field. Such portable medical equipment must be properly maintained so that there is sufficient battery life to power the equipment for the duration of the emergency. However, it is too often the case that the equipment is not properly maintained so that batteries are not fully charged and in some cases, necessary equipment is not available to the first responders. The systems and methods described herein may provide for wireless power to portable medical equipment (and associated sensor inputs on the patient) in such a way that the charging and maintaining of batteries and power packs is provided automatically and without human intervention. Such a system also benefits from the improved mobility of a patient unencumbered by a variety of power cords attached to the many medical monitors and devices used in their treatment.

[00468] The systems and methods described herein may be used to for the powering/charging of personal hearing aids. Personal hearing aids need to be small and light to fit into or around the ear of a person. The size and weight restrictions limit the size of batteries that can be used. Likewise, the size and weight restrictions of the device make battery replacement difficult due to the delicacy of the components. The dimensions of the devices and hygiene concerns make it difficult to integrate additional charging ports to allow recharging of the batteries. The systems and methods described herein may be integrated into the hearing aid and may reduce the size of the necessary batteries which may allow even smaller hearing aids. Using the principles disclosed herein, the batteries of the hearing aid may be recharged without requiring external connections or charging ports. Charging and device circuitry and a small rechargeable battery may be integrated into a form factor of a conventional hearing aid battery allowing retrofit into existing hearing aids. The hearing aid may be recharged while it is used and worn by a person. The energy source may be integrated into a pad or a cup allowing recharging when the hearing is placed on such a structure. The charging source may be integrated into a hearing aid dryer box allowing wireless recharging while the hearing aid is drying or being sterilized. The source and device resonator may be used to also heat the device reducing or eliminating the need for an additional heating element. Portable charging cases powered by batteries or AC adaptors may be used as storage and charging stations.

[00469] The source resonator for the medical systems described above may be in the main body of some or all of the medical equipment, with device resonators on the patient's sensors and devices; the source resonator may be in the ambulance with device resonators on the patient's sensors and the main body of some or all of the equipment; a primary source resonator may be in the ambulance for transferring wireless power to a device resonator on the medical equipment while the medical equipment is in the ambulance and a second source resonator is in the main body of the medical equipment and a second device resonator on the patient sensors when the equipment is away from the ambulance; and the like. The systems and methods described herein may significantly improve the ease with which medical personnel are able to transport patients from one location to another, where power wires and the need to replace or manually charge associated batteries may now be reduced.

[00470] The systems and methods described herein may be used for the charging of devices inside a military vehicle or facility, such as a tank, armored carrier, mobile shelter, and the like. For instance, when soldiers come back into a vehicle after "action" or a mission, they may typically start charging their electronic devices. If their electronic devices were equipped with device resonators, and there was a source resonator inside the vehicle, (e.g. integrated in the seats or on the ceiling of the vehicle), their devices would start charging immediately. In fact, the same vehicle could provide power to soldiers/robots (e.g. packbot from iRobot) standing outside or walking beside the vehicle. This capability may be useful in minimizing accidental battery-swapping with someone else (this may be a significant issue, as soldiers tend to trust only their own batteries); in enabling quicker exits from a vehicle under attack; in powering or charging laptops or other electronic devices inside a tank, as too many wires inside the tank may present a hazard in terms of reduced ability to move around fast in case of "trouble" and/or decreased visibility; and the like. The systems and methods described herein may provide a significant improvement in association with powering portable power equipment in a military environment.

[00471] The systems and methods described herein may provide wireless powering or charging capabilities to mobile vehicles such as golf carts or other types of carts, all-terrain vehicles, electric bikes, scooters, cars, mowers, bobcats and other vehicles typically used for construction and landscaping and the like. The systems and methods described herein may provide wireless powering or charging capabilities to miniature mobile vehicles, such as mini-helicopters, airborne drones, remote control planes, remote control boats, remote controlled or robotic rovers, remote controlled or robotic lawn mowers or equipment, bomb detection robots, and the like. For instance, mini-helicopter flying above a military vehicle to increase its field of view can fly for a few minutes on standard batteries. If these mini-helicopters were fitted with a device resonator, and the control vehicle had a source resonator, the mini-helicopter might be able to fly indefinitely. The systems and methods described herein may provide an effective alternative to recharging or replacing the batteries for use in miniature mobile vehicles. In addition, the systems and methods described herein may provide power/charging to even smaller devices, such as microelectromechanical systems (MEMS), nano-robots, nano devices, and the like. In addition, the systems and methods described herein may be implemented by installing a source device in a mobile vehicle or flying device to enable it to serve as an in-field or in-flight re-charger, that may position itself autonomously in proximity to a mobile vehicle that is equipped with a device resonator.

[00472] The systems and methods described herein may be used to provide power networks for temporary facilities, such as military camps, oil drilling setups, remote filming locations, and the like, where electrical power is required, such as for power generators, and where power cables are typically run around the temporary facility. There are many instances when it is necessary to set up temporary facilities that require power. The systems and methods described herein may enable a more efficient way to rapidly set up and tear down these facilities, and may reduce the number of wires that must be run throughout the faculties to supply power. For instance, when Special Forces move into an area, they may erect tents and drag many wires around the camp to provide the required electricity. Instead, the systems and methods described herein may enable an army vehicle, outfitted with a power supply and a source resonator, to park in the center of the camp, and provide all the power to nearby tents where the device resonator may be integrated into the tents, or some other piece of equipment associated with each tent or area. A series of source-device-source-device resonators may be used to extend the power to tents that are farther away. That is, the tents closest to the vehicle could then provide power to tents behind them. The systems and methods described herein may provide a significant improvement to the efficiency with which temporary installations may be set up and torn down, thus improving the mobility of the associated facility.

[00473] The systems and methods described herein may be used in vehicles, such as for replacing wires, installing new equipment, powering devices brought into the vehicle, charging the battery of a vehicle (e.g. for a traditional gas powered engine, for a hybrid car, for an electric car, and the like), powering devices mounted to the interior or exterior of the vehicle, powering devices in the vicinity of the vehicle, and the like. For example, the systems and methods described herein may be used to replace wires such as those are used to power lights, fans and sensors distributed throughout a vehicle. As an example, a typical car may have 50kg of wires associated with it, and the use of the systems and methods described herein may enable the elimination of a substantial amount of this wiring. The performance of larger and more weight sensitive vehicles such as airplanes or satellites could benefit greatly from having the number of cables that must be run throughout the vehicle reduced. The systems and methods described herein may allow the accommodation of removable or supplemental portions of a vehicle with electric and electrical devices without the need for electrical harnessing. For example, a motorcycle may have removable side boxes that act as a temporary trunk space for when the cyclist is going on a long trip. These side boxes may have exterior lights, interior lights, sensors, auto equipment, and the like, and if not for being equipped with the systems and methods described herein might require electrical connections and harnessing.

[00474] An in-vehicle wireless power transmission system may charge or power one or more mobile devices used in a car: mobile phone handset, Bluetooth headset, blue tooth hands free speaker phone, GPS, MP3 player, wireless audio transceiver for streaming MP3 audio through car stereo via FM, Bluetooth, and the like. The in vehicle wireless power source may utilize source resonators that are arranged in any of several possible configurations including charging pad on dash, charging pad otherwise mounted on floor, or between seat and center console, charging "cup" or receptacle that fits in cup holder or on dash, and the like.

[00475] The wireless power transmission source may utilize a rechargeable battery system such that said supply battery gets charged whenever the vehicle power is on such that when the vehicle is turned off the wireless supply can draw power from the supply battery and can continue to wirelessly charge or power mobile devices that are still in the car.

[00476] The plug-in electric cars, hybrid cars, and the like, of the future need to be charged, and the user may need to plug in to an electrical supply when they get home or to a charging station. Based on a single over-night recharging, the user may be able to drive up to 50 miles the next day. Therefore, in the instance of a hybrid car, if a person drives less than 50 miles on most days, they will be driving mostly on electricity. However, it would be beneficial if they didn't have to remember to plug in the car at night. That is, it would be nice to simply drive into a garage, and have the car take care of its own charging. To this end, a source resonator may be built into the garage floor and/or garage side-wall, and the device resonator may be built into the bottom (or side) of the car. Even a few kW transfer may be sufficient to recharge the car over-night. The in-vehicle device resonator may measure magnetic field properties to provide feedback to assist in vehicle (or any similar device) alignment to a stationary resonating source. The vehicle may use this positional feedback to automatically position itself to achieve optimum alignment, thus optimum power transmission efficiency. Another method may be to use the positional feedback to help the human operator to properly position the vehicle or device, such as by making LED's light up, providing noises, and the like when it is well positioned. In such cases where the amount of power being transmitted could present a safety hazard to a person or animal that intrudes into the active field volume, the source or receiver device may be equipped with an active light curtain or some other external device capable of sensing intrusion into the active field volume, and capable of shutting off the source device and alert a human operator. In addition, the source device may be equipped with self-sensing capability such that it may detect that its expected power transmission rate has been interrupted by an intruding element, and in such case shut off the source device and alert a human operator. Physical or mechanical structures such as hinged doors or inflatable bladder shields may be incorporated as a physical barrier to prevent unwanted intrusions. Sensors such as optical, magnetic, capacitive, inductive, and the like may also be used to detect foreign structures or interference between the source and device resonators. The shape of the source resonator may be shaped such to prevent water or debris accumulation. The source resonator may be placed in a cone shaped enclosure or may have an enclosure with an angled top to allow water and debris to roll off. The source of the system may use battery power of the vehicle or its own battery power to transmit its presence to the source to initiate power transmission.

[00477] The source resonator may be mounted on an embedded or hanging post, on a wall, on a stand, and the like for coupling to a device resonator mounted on the bumper, hood, body panel, and the like, of an electric vehicle. The source resonator may be enclosed or embedded into a flexible enclosure such as a pillow, a pad, a bellows, a spring loaded enclosure and the like so that the electric vehicle may make contact with the structure containing the source coil without damaging the car in any way. The structure containing the source may prevent objects from getting between the source and device resonators. Because the wireless power transfer may be relatively insensitive to misalignments between the source and device coils, a variety of flexible source structures and parking procedures may be appropriate for this application.

[00478] The systems and methods described herein may be used to trickle charge batteries of electric, hybrid or combustion engine vehicles. Vehicles may require small amounts of power to maintain or replenish battery power. The power may be transferred wirelessly from a source to a device resonator that may be incorporated into the front grill, roof, bottom, or other parts of the vehicle. The device resonator may be designed to fit into a shape of a logo on the front of a vehicle or around the grill as not to obstruct air flow through the radiator. The device or source resonator may have additional modes of operation that allow the resonator to be used as a heating element which can be used to melt of snow or ice from the vehicle.

[00479] An electric vehicle or hybrid vehicle may require multiple device resonators, such as to increase the ease with which the vehicle may come in proximity with a source resonator for charging (i.e. the greater the number and varied position of device resonators are, the greater the chances that the vehicle can pull in and interface with a diversity of charging stations), to increase the amount of power that can be delivered in a period of time (e.g. additional device resonators may be required to keep the local heating due to charging currents to acceptable levels), to aid in automatic parking/docking the vehicle with the charging station, and the like. For example, the vehicle may have multiple resonators (or a single resonator) with a feedback system that provides guidance to either the driver or an automated parking/docking facility in the parking of the vehicle for optimized charging conditions (i.e., the optimum positioning of the vehicle's device resonator to the charging station's source resonator may provide greater power transfer efficiency). An automated parking/docking facility may allow for the automatic parking of the vehicle based on how well the vehicle is coupled.

[00480] The power transmission system may be used to power devices and peripherals of a vehicle. Power to peripherals may be provided while a vehicle is charging, or while not charging, or power may be delivered to conventional vehicles that do not need charging. For example, power may be transferred wirelessly to conventional non-electric cars to power air conditioning, refrigeration units, heaters, lights, and the like while parked to avoid running the engine which may be important to avoid exhaust build up in garage parking lots or loading docks. Power may for example be wirelessly transferred to a bus while it is parked to allow powering of lights, peripherals, passenger devices, and the like avoiding the use of onboard engines or power sources. Power may be wirelessly transferred to an airplane while parked on the tarmac or in a hanger to power instrumentation, climate control, de-icing equipment, and the like without having to use onboard engines or power sources.

[00481] Wireless power transmission on vehicles may be used to enable the concept of Vehicle to Grid (V2G). Vehicle to grid is based on utilizing electric vehicles and plug-in hybrid electric vehicles (PHEV) as distributed energy storage devices, charged at night when the electric grid is underutilized, and available to discharge back into the grid during episodes of peak demand that occur during the day. The wireless power transmission system on a vehicle and the respective infrastructure may be implemented in such a way as to enable bidirectional energy flow—so that energy can flow back into the grid from the vehicle—without requiring a plug in connection. Vast fleets of vehicles, parked at factories, offices, parking lots, can be viewed as "peaking power capacity" by the smart grid. Wireless power transmission on vehicles can make such a V2G vision a reality. By simplifying the process of connecting a vehicle to the grid, (i.e. by simply parking it in a wireless charging enabled parking spot), it becomes much more likely that a certain number of vehicles will be "dispatchable" when the grid needs to tap their power. Without wireless charging, electric and PHEV owners will likely charge their vehicles at home, and park them at work in conventional parking spots. Who will want to plug their vehicle in at work, if they do not need charging? With wireless charging systems capable of handling 3 kW, 100,000 vehicles can provide 300 Megawatts back to the grid—using energy generated the night before by cost effective base load generating capacity. It is the streamlined ergonomics of the cordless self charging PHEV and electric vehicles that make it a viable V2G energy source.

[00482] The systems and methods described herein may be used to power sensors on the vehicle, such as sensors in tires to measure air-pressure, or to run peripheral devices in the vehicle, such as cell phones, GPS devices, navigation devices, game players, audio or video players, DVD players, wireless routers, communications equipment, anti-theft devices, radar devices, and the like. For example, source resonators described herein may be built into the main compartment of the car in order to supply power to a variety of devices located both inside and outside of the main compartment of the car. Where the vehicle is a motorcycle or the like, devices described herein may be integrated into the body of the motorcycle, such as under the seat, and device resonators may be provided in a user's helmet, such as for communications, entertainment, signaling, and the like, or device resonators may be provided in the user's jacket, such as for displaying signals to other drivers for safety, and the like.

[00483] The systems and methods described herein may be used in conjunction with transportation infrastructure, such as roads, trains, planes, shipping, and the like. For example, source resonators may be built into roads, parking lots, rail-lines, and the like. Source resonators may be built into traffic lights, signs, and the like. For example, with source resonators embedded into a road, and device resonators built into vehicles, the vehicles may be provided power as they drive along the road or as they are parked in lots or on the side of the road. The systems and methods described herein may provide an effective way for electrical systems in vehicles to be powered and/or charged while the vehicle traverses a road network, or a portion of a road network. In this way, the systems and methods described herein may contribute to the powering/charging of autonomous vehicles, automatic guided vehicles, and the like. The systems and methods described herein may provide power to vehicles in places where they typically idle or stop, such as in the vicinity of traffic lights or signs, on highway ramps, or in parking lots.

[00484] The systems and methods described herein may be used in an industrial environment, such as inside a factory for powering machinery, powering/charging robots, powering and/or charging wireless sensors on robot arms, powering/charging tools and the like. For example, using the systems and methods described herein to supply power to devices on the arms of robots may help eliminate direct wire connections across the joints of the robot arm. In this way, the wearing out of such direct wire connections may be reduced, and the reliability of the robot increased. In this case, the device resonator may be out on the arm of the robot, and the source resonator may be at the base of the robot, in a central location near the robot, integrated into the industrial facility in which the robot is providing service, and the like. The use of the systems and methods described herein may help eliminate wiring otherwise associated with power distribution within the industrial facility, and thus benefit the overall reliability of the facility.

[00485] The systems and methods described herein may be used for underground applications, such as drilling, mining, digging, and the like. For example, electrical components and sensors associated with drilling or excavation may utilize the systems and methods described herein to eliminate cabling associated with a digging mechanism, a drilling bit, and the like, thus eliminating or minimizing cabling near the excavation point. In another example, the systems and methods described herein may be used to provide power to excavation equipment in a mining application where the power requirements for the equipment may be high and the distances large, but where there are no people to be subjected to the associated required fields. For instance, the excavation area may have device resonator powered digging equipment that has high power requirements and may be digging relatively far from the source resonator. As a result the source resonator may need to provide high field intensities to satisfy these requirements, but personnel are far enough away to be outside these high intensity fields. This high power, no personnel, scenario may be applicable to a plurality of industrial applications.

[00486] The systems and methods described herein may also use the near-field non-radiative resonant scheme for information transfer rather than, or in addition to, power transfer. For instance, information being transferred by near-field non-radiative resonance techniques may not be susceptible to eavesdropping and so may provide an increased level of security compared to traditional wireless communication schemes. In addition, information being transferred by near-field non-radiative resonance techniques may not interfere with the EM radiative spectrum and so may not be a source of EM interference, thereby allowing communications in an extended frequency range and well within the limits set by any regulatory bodies. Communication services may be provided between remote, inaccessible or hard-to-reach places such as between remote sensors, between sections of a device or vehicle, in tunnels, caves and wells (e.g. oil wells, other drill sites) and between underwater or underground devices, and the like. Communications services may be provided in places where magnetic fields experience less loss than electric fields.

[00487] The systems and methods described herein may enable the simultaneous transmission of power and communication signals between sources and devices in wireless power transmission systems, or it may enable the transmission of power and communication signals during different time periods or at different frequencies. The performance characteristics of the resonator may be controllably varied to preferentially support or limit the efficiency or range of either energy or information transfer. The performance characteristics of the resonators may be controlled to improve the security by reducing the range of information transfer, for example. The performance characteristics of the resonators may be varied continuously, periodically, or according to a predetermined, computed or automatically adjusted algorithm. For example, the power and information transfer enabled by the systems and methods described herein may be provided in a time multiplexed or frequency multiplexed manner. A source and device may signal each other by tuning, changing, varying, dithering, and the like, the resonator impedance which may affect the reflected impedance of other resonators that can be detected. The information transferred as described herein may include information regarding device identification, device power requirements, handshaking protocols, and the like.

[00488] The source and device may sense, transmit, process and utilize position and location information on any other sources and/or devices in a power network. The source and device may capture or use information such as elevation, tilt, latitude and longitude, and the like from a variety of sensors and sources that may be built into the source and device or may be part of a component the source or device connect. The positioning and orientation information may include sources such as global positioning sensors (GPS), compasses, accelerometers, pressure sensors, atmospheric barometric sensors, positioning systems which use Wi-Fi or cellular network signals, and the like. The source and device may use the position and location information to find nearby wireless power transmission sources. A source may broadcast or communicate with a central station or database identifying its location. A device may obtain the source location information from the central station or database or from the local broadcast and guide a user or an operator to the source with the aid of visual, vibrational, or auditory signals. Sources and devices may be nodes in a power network, in a communications network, in a sensor network, in a navigational network, and the like or in kind of combined functionality network.

[00489] The position and location information may also be used to optimize or coordinate power delivery. Additional information about the relative position of a source and a device may be used to optimize magnetic field direction and resonator alignment. The orientation of a device and a source which may be obtained from accelerometers and magnetic sensors, and the like, for example, may be used to identify the orientation of resonators and the most favorable direction of a magnetic field such that the magnetic flux is not blocked by the device circuitry. With such information a source with the most favorable orientation, or a combination of sources, may be used. Likewise, position and orientation information may be used to move or provide feedback to a user or operator of a device to place a device in a favorable orientation or location to maximize power transmission efficiency, minimize losses, and the like.

[00490] The source and device may include power metering and measuring circuitry and capability. The power metering may be used to track how much power was delivered to a device or how much power was transferred by a source. The power metering and power usage information may be used in fee based power delivery arrangements for billing purposes. Power metering may be also be used to enable power delivery policies to ensure power is distributed to multiple devices according to specific criteria. For example, the power metering may be used to categorize devices based on the amount of power they received and priority in power delivery may be given to those having received the least power. Power metering may be used to provide tiered delivery services such as “guaranteed power” and “best effort power” which may be billed at separate rates. Power metering may be used to institute and enforce hierarchical power delivery structures and may enable priority devices to demand and receive- more power under certain circumstances or use scenarios.

[00491] Power metering may be used to optimize power delivery efficiency and minimize absorption and radiation losses. Information related to the power received by devices may be used by a source in conjunction with information about the power output of the source to identify unfavorable operating environments or frequencies. For example, a source may compare the amount of power which was received by the devices and the amount of power which it transmitted to determine if the transmission losses may be unusually or unacceptably large. Large transmission losses may be due to an unauthorized device receiving power from the source and the source and other devices may initiate frequency hopping of the resonance frequency or other defensive measures to prevent or deter unauthorized use. Large transmission losses may be due to absorption losses for example, and the device and source may tune to alternate resonance frequencies to minimize such losses. Large transmission losses may also indicate the presence of unwanted or unknown objects or materials and the source may turn down or off its power level until the unwanted or unknown object is removed or identified, at which point the source may resume powering remote devices.

[00492] The source and device may include authentication capability. Authentication may be used to ensure that only compatible sources and devices are able to transmit and receive power. Authentication may be used to ensure that only authentic devices that are of a specific manufacturer and not clones or devices and sources from other manufacturers, or only devices that are part of a specific subscription or plan, are able to receive power from a source. Authentication may be based on cryptographic request and respond protocols or it may be based on the unique signatures of perturbations of specific devices allowing them to be used and authenticated based on properties similar to physically unclonable functions. Authentication may be performed locally between each source and device with local communication or it may be used with third person authentication methods where the source and device authenticate with communications to a central authority. Authentication protocols may use position information to alert a local source or sources of a genuine device.

[00493] The source and device may use frequency hopping techniques to prevent unauthorized use of a wireless power source. The source may continuously adjust or change the resonant frequency of power delivery. The changes in frequency may be performed in a pseudorandom or predetermined manner that is known, reproducible, or communicated to authorized device but difficult to predict. The rate of frequency hopping and the number of various frequencies used may be large and frequent enough to ensure that unauthorized use is difficult or impractical. Frequency hopping may be implemented by tuning the impedance network, tuning any of the driving circuits, using a plurality of resonators tuned or tunable to multiple resonant frequencies, and the like.

[00494] The source may have a user notification capability to show the status of the source as to whether it is coupled to a device resonator and transmitting power, if it is in standby mode, or if the source resonator is detuned or perturbed by an external object. The notification capability may include visual, auditory, and vibrational methods. The notification may be as simple as three color lights, one for each state, and optionally a speaker to provide notification in case of an error in operation. Alternatively, the notification capability may involve an interactive display that shows the status of the source and optionally provides instructions on how to fix or solve any errors or problems identified.

[00495] As another example, wireless power transfer may be used to improve the safety of electronic explosive detonators. Explosive devices are detonated with an electronic detonator, electric detonator, or shock tube detonator. The electronic detonator utilizes stored electrical energy (usually in a capacitor) to activate the igniter charge, with a low energy trigger signal transmitted conductively or by radio. The electric detonator utilizes a high energy conductive trigger signal to provide both the signal and the energy required to activate the igniter charge. A shock tube sends a controlled explosion through a hollow tube coated with explosive from the generator to the igniter charge. There are safety issues associated with the electric and electronic detonators, as there are cases of stray electromagnetic energy causing unintended activation. Wireless power transfer via sharply resonant magnetic coupling can improve the safety of such systems.

[00496] Using the wireless power transfer methods disclosed herein, one can build an electronic detonation system that has no locally stored energy, thus reducing the risk of unintended activation. A wireless power source can be placed in proximity (within a few meters) of the detonator. The detonator can be equipped with a resonant capture coil. The activation energy can be transferred when the wireless power source has been triggered. The triggering of the wireless power source can be initiated by any number of mechanisms: radio, magnetic near field radio, conductive signaling, ultrasonics, laser light. Wireless power transfer based on resonant magnetic coupling also has the benefit of being able to transfer power through materials such as rock, soil, concrete, water, and other dense materials. The use of very high Q coils as receivers and sources, having very narrow band response and sharply tuned to proprietary frequencies, further ensure that the detonator circuits cannot capture stray EMI and activate unintentionally.

[00497] The resonator of a wirelessly powered device may be external, or outside of the device, and wired to the battery of the device. The battery of the device may be modified to include appropriate rectification and control circuitry to receive the alternating currents of the device resonator. This can enable configurations with larger external coils, such as might be built into a battery door of a keyboard or mouse, or digital still camera, or even larger coils that are attached to the device but wired back to the battery/converter with ribbon cable. The battery door can be modified to provide interconnection from the external coil to the battery/converter (which will need an exposed contact that can touch the battery door contacts.

[00498] While the invention has been described in connection with certain preferred embodiments, other embodiments will be understood by one of ordinary skill in the art and are intended to fall within the scope of this disclosure, which is to be interpreted in the broadest sense allowable by law.

[00499] All documents referenced herein are hereby incorporated by reference.

CLAIMS

What is claimed is:

1. A system, comprising:

a source resonator optionally coupled to an energy source; and

a second resonator located a distance from the source resonator,

wherein the source resonator and the second resonator are coupled to provide near-field wireless energy transfer among the source resonator and the second resonator and wherein the field of at least one of the source resonator and the second resonator is shaped to avoid a loss-inducing object.

2. The system of claim 1, wherein at least one of the source resonator and the second resonator have a quality factor, Q>100.

3. The system of claim 1, wherein the source resonator Q is greater than 100 and the second resonator Q is greater than 100.

4. The system of claim 1, wherein the square root of the source resonator Q times the second resonator Q is greater than 100.

5. The system of claim 1, comprising more than one source resonator.

6. The system of claim 1, comprising more than one second resonator.

7. The system of claim 1, comprising more than three resonators.

8. A method, comprising:

providing a source resonator optionally coupled to an energy source and a second resonator located a distance from the source resonator,

wherein the source resonator and the second resonator are coupled to provide near-field wireless energy transfer among the source resonator and the second resonator and wherein the field of at least one of the source resonator and the second resonator is shaped to avoid a loss-inducing object.

9. The method of claim 8, wherein at least one of the source resonator and the second resonator have a quality factor, Q>100.

10. The method of claim 8, wherein the source resonator Q is greater than 100 and the second resonator Q is greater than 100.

11. The method of claim 8, wherein the square root of the source resonator Q times the second resonator Q is greater than 100.

12. The method of claim 8, comprising more than one source resonator.

13. The method of claim 8, comprising more than one second resonator.

14. The method of claim 8, comprising more than three resonators.

ABSTRACT

[00500] In embodiments of the present invention improved capabilities are described for a method and system comprising a source resonator optionally coupled to an energy source and a second resonator located a distance from the source resonator, where the source resonator and the second resonator are coupled to provide near-field wireless energy transfer among the source resonator and the second resonator and where the field of at least one of the source resonator and the second resonator is shaped to avoid a loss-inducing object.

Electronic Acknowledgement Receipt

EFS ID:	6712276
Application Number:	12647763
International Application Number:	
Confirmation Number:	2576
Title of Invention:	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS
First Named Inventor/Applicant Name:	Aristeidis Karalis
Customer Number:	43520
Filer:	John H. Nortrop
Filer Authorized By:	
Attorney Docket Number:	WTCY-0026-P07
Receipt Date:	28-DEC-2009
Filing Date:	
Time Stamp:	14:26:04
Application Type:	Utility under 35 USC 111(a)

Payment information:

Submitted with Payment	no

File Listing:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
1	Application Data Sheet	WTCY-0026-P07_122809_ADS.pdf	1358534 8923200a7c3c196b92afba5aa867730956d668d3	no	8

Warnings:

Information:

2	Drawings-only black and white line drawings	WTCY-0026-P07_122809_Drawings.pdf	2796384 499f3c914d7e62ffa5871258abe68cb7951ce13c	no	53
Warnings:					
Information:					
3		WTCY-0026-P07_122809_Specification.pdf	741649 dd5c0132df90a64a90d17ffcb1d84065d990cea1	yes	143

Multipart Description/PDF files in .zip description

Document Description	Start	End
Specification	1	140
Claims	141	142
Abstract	143	143

Warnings:

Information:

Total Files Size (in bytes):	4896567

This Acknowledgement Receipt evidences receipt on the noted date by the USPTO of the indicated documents, characterized by the applicant, and including page counts, where applicable. It serves as evidence of receipt similar to a Post Card, as described in MPEP 503.

New Applications Under 35 U.S.C. 111
If a new application is being filed and the application includes the necessary components for a filing date (see 37 CFR 1.53(b)-(d) and MPEP 506), a Filing Receipt (37 CFR 1.54) will be issued in due course and the date shown on this Acknowledgement Receipt will establish the filing date of the application.

National Stage of an International Application under 35 U.S.C. 371
If a timely submission to enter the national stage of an international application is compliant with the conditions of 35 U.S.C. 371 and other applicable requirements a Form PCT/DO/EO/903 indicating acceptance of the application as a national stage submission under 35 U.S.C. 371 will be issued in addition to the Filing Receipt, in due course.

New International Application Filed with the USPTO as a Receiving Office
If a new international application is being filed and the international application includes the necessary components for an international filing date (see PCT Article 11 and MPEP 1810), a Notification of the International Application Number and of the International Filing Date (Form PCT/RO/105) will be issued in due course, subject to prescriptions concerning national security, and the date shown on this Acknowledgement Receipt will establish the international filing date of the application.

DocCode - SCORE

SCORE Placeholder Sheet for IFW Content

Application Number: 12647763 Document Date: 12/28/2009

The presence of this form in the IFW record indicates that the following document type was received in electronic format on the date identified above. This content is stored in the SCORE database.

- Drawings – Other than Black and White Line Drawings

Since this was an electronic submission, there is no physical artifact folder, no artifact folder is recorded in PALM, and no paper documents or physical media exist. The TIFF images in the IFW record were created from the original documents that are stored in SCORE.

To access the documents in the SCORE database, refer to instructions developed by SIRA.

At the time of document entry (noted above):
- Examiners may access SCORE content via the eDAN interface.
- Other USPTO employees can bookmark the current SCORE URL (http://es/ScoreAccessWeb/).
- External customers may access SCORE content via the Public and Private PAIR interfaces.

Form Revision Date: February 8, 2006

Filing Date: 12/28/09

PTO/SB/06 (12-04)
Approved for use through 7/31/2006. OMB 0651-0032
U.S. Patent and Trademark Office; U.S. DEPARTMENT OF COMMERCE
Under the Paperwork Reduction Act of 1995, no persons are required to respond to a collection of information unless it displays a valid OMB control number.

PATENT APPLICATION FEE DETERMINATION RECORD

Substitute for Form PTO-875

Application or Docket Number: 12/647,763

APPLICATION AS FILED – PART I

FOR	NUMBER FILED (Column 1)	NUMBER EXTRA (Column 2)	SMALL ENTITY RATE ($)	SMALL ENTITY FEE ($)	OR	OTHER THAN SMALL ENTITY RATE ($)	OTHER THAN SMALL ENTITY FEE ($)
BASIC FEE (37 CFR 1.16(a), (b), or (c))	N/A	N/A	N/A			N/A	330
SEARCH FEE (37 CFR 1.16(k), (i), or (m))	N/A	N/A	N/A			N/A	540
EXAMINATION FEE (37 CFR 1.16(o), (p), or (q))	N/A	N/A	N/A			N/A	220
TOTAL CLAIMS (37 CFR 1.16(i))	14 minus 20 =		x$26		OR	x$52	
INDEPENDENT CLAIMS (37 CFR 1.16(h))	2 minus 3 =	*	x$110			x$220	
APPLICATION SIZE FEE (37 CFR 1.16(s))	If the specification and drawings exceed 100 sheets of paper, the application size fee due is $270 ($135 for small entity) for each additional 50 sheets or fraction thereof. See 35 U.S.C. 41(a)(1)(G) and 37 CFR						270
MULTIPLE DEPENDENT CLAIM PRESENT (37 CFR 1.16(j))			195			390	
			TOTAL			TOTAL	1360

* If the difference in column 1 is less than zero, enter "0" in column 2.

APPLICATION AS AMENDED – PART II

AMENDMENT A

	CLAIMS REMAINING AFTER AMENDMENT (Column 1)		HIGHEST NUMBER PREVIOUSLY PAID FOR (Column 2)	PRESENT EXTRA (Column 3)	SMALL ENTITY RATE ($)	SMALL ENTITY ADDITIONAL FEE ($)	OR	OTHER THAN SMALL ENTITY RATE ($)	OTHER THAN SMALL ENTITY ADDITIONAL FEE ($)
Total (37 CFR 1.16(i))	*	Minus	**	=	x =		OR	x =	
Independent (37 CFR 1.16(h))	*	Minus	***	=	x =		OR	x =	
Application Size Fee (37 CFR 1.16(s))									
FIRST PRESENTATION OF MULTIPLE DEPENDENT CLAIM (37 CFR 1.16(j))					N/A		OR	N/A	
					TOTAL ADD'T FEE		OR	TOTAL ADD'T FEE	

AMENDMENT B

	CLAIMS REMAINING AFTER AMENDMENT (Column 1)		HIGHEST NUMBER PREVIOUSLY PAID FOR (Column 2)	PRESENT EXTRA (Column 3)	RATE ($)	ADDITIONAL FEE ($)	OR	RATE ($)	ADDITIONAL FEE ($)
Total (37 CFR 1.16(i))	*	Minus	**	=	x =		OR	x =	
Independent (37 CFR 1.16(h))	*	Minus	***	=	x =		OR	x =	
Application Size Fee (37 CFR 1.16(s))									
FIRST PRESENTATION OF MULTIPLE DEPENDENT CLAIM (37 CFR 1.16(j))					N/A		OR	N/A	
					TOTAL ADD'T FEE		OR	TOTAL ADD'T FEE	

* If the entry in column 1 is less than the entry in column 2, write "0" in column 3.
** If the "Highest Number Previously Paid For" IN THIS SPACE is less than 20, enter "20".
*** If the "Highest Number Previously Paid For" IN THIS SPACE is less than 3, enter "3".
The "Highest Number Previously Paid For" (Total or Independent) is the highest number found in the appropriate box in column 1.

This collection of information is required by 37 CFR 1.16. The information is required to obtain or retain a benefit by the public which is to file (and by the USPTO to process) an application. Confidentiality is governed by 35 U.S.C. 122 and 37 CFR 1.14. This collection is estimated to take 12 minutes to complete, including gathering, preparing, and submitting the completed application form to the USPTO. Time will vary depending upon the individual case. Any comments on the amount of time you require to complete this form and/or suggestions for reducing this burden, should be sent to the Chief Information Officer, U.S. Patent and Trademark Office, U.S. Department of Commerce, P.O. Box 1450, Alexandria, VA 22313-1450. DO NOT SEND FEES OR COMPLETED FORMS TO THIS ADDRESS. **SEND TO: Commissioner for Patents, P.O. Box 1450, Alexandria, VA 22313-1450.**

If you need assistance in completing the form, call 1-800-PTO-9199 and select option 2.

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APPLICATION NUMBER	FILING or 371(c) DATE	GRP ART UNIT	FIL FEE REC'D	ATTY.DOCKET.NO	TOT CLAIMS	IND CLAIMS
12/647,763	12/28/2009	2828	0.00	WTCY-0026-P07	14	2

CONFIRMATION NO. 2576

43520
STRATEGIC PATENTS P.C..
C/O PORTFOLIOIP
P.O. BOX 52050
MINNEAPOLIS, MN 55402

FILING RECEIPT

OC000000039637579

Date Mailed: 01/14/2010

Receipt is acknowledged of this non-provisional patent application. The application will be taken up for examination in due course. Applicant will be notified as to the results of the examination. Any correspondence concerning the application must include the following identification information: the U.S. APPLICATION NUMBER, FILING DATE, NAME OF APPLICANT, and TITLE OF INVENTION. Fees transmitted by check or draft are subject to collection. Please verify the accuracy of the data presented on this receipt. **If an error is noted on this Filing Receipt, please submit a written request for a Filing Receipt Correction. Please provide a copy of this Filing Receipt with the changes noted thereon. If you received a "Notice to File Missing Parts" for this application, please submit any corrections to this Filing Receipt with your reply to the Notice. When the USPTO processes the reply to the Notice, the USPTO will generate another Filing Receipt incorporating the requested corrections**

Applicant(s)

Aristeidis Karalis, Boston, MA;
Andre B. Kurs, Chestnut Hill, MA;
Andrew J. Campanella, Waltham, MA;
Konrad Kulikowski, Somerville, MA;
Katherine L. Hall, Westford, MA;
Marin Soljacic, Belmont, MA;
Morris P. Kesler, Bedford, MA;

Power of Attorney: None

Domestic Priority data as claimed by applicant

This application is a CIP of 12/567,716 09/25/2009
which claims benefit of 61/100,721 09/27/2008
and claims benefit of 61/108,743 10/27/2008
and claims benefit of 61/147,386 01/26/2009
and claims benefit of 61/152,086 02/12/2009
and claims benefit of 61/178,508 05/15/2009
and claims benefit of 61/182,768 06/01/2009
and claims benefit of 61/121,159 12/09/2008
and claims benefit of 61/142,977 01/07/2009
and claims benefit of 61/142,885 01/06/2009
and claims benefit of 61/142,796 01/06/2009
and claims benefit of 61/142,889 01/06/2009
and claims benefit of 61/142,880 01/06/2009
and claims benefit of 61/142,818 01/06/2009
and claims benefit of 61/142,887 01/06/2009

and claims benefit of 61/156,764 03/02/2009
and claims benefit of 61/143,058 01/07/2009
and claims benefit of 61/152,390 02/13/2009
and claims benefit of 61/163,695 03/26/2009
and claims benefit of 61/172,633 04/24/2009
and claims benefit of 61/169,240 04/14/2009
and claims benefit of 61/173,747 04/29/2009

Foreign Applications

Projected Publication Date: To Be Determined - pending completion of Missing Parts

Non-Publication Request: No

Early Publication Request: No

Title

WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS

Preliminary Class

372

PROTECTING YOUR INVENTION OUTSIDE THE UNITED STATES

Since the rights granted by a U.S. patent extend only throughout the territory of the United States and have no effect in a foreign country, an inventor who wishes patent protection in another country must apply for a patent in a specific country or in regional patent offices. Applicants may wish to consider the filing of an international application under the Patent Cooperation Treaty (PCT). An international (PCT) application generally has the same effect as a regular national patent application in each PCT-member country. The PCT process **simplifies** the filing of patent applications on the same invention in member countries, but **does not result** in a grant of "an international patent" and does not eliminate the need of applicants to file additional documents and fees in countries where patent protection is desired.

Almost every country has its own patent law, and a person desiring a patent in a particular country must make an application for patent in that country in accordance with its particular laws. Since the laws of many countries differ in various respects from the patent law of the United States, applicants are advised to seek guidance from specific foreign countries to ensure that patent rights are not lost prematurely.

Applicants also are advised that in the case of inventions made in the United States, the Director of the USPTO must issue a license before applicants can apply for a patent in a foreign country. The filing of a U.S. patent application serves as a request for a foreign filing license. The application's filing receipt contains further information and guidance as to the status of applicant's license for foreign filing.

Applicants may wish to consult the USPTO booklet, "General Information Concerning Patents" (specifically, the section entitled "Treaties and Foreign Patents") for more information on timeframes and deadlines for filing foreign patent applications. The guide is available either by contacting the USPTO Contact Center at 800-786-9199, or it can be viewed on the USPTO website at http://www.uspto.gov/web/offices/pac/doc/general/index.html.

For information on preventing theft of your intellectual property (patents, trademarks and copyrights), you may wish to consult the U.S. Government website, http://www.stopfakes.gov. Part of a Department of Commerce initiative, this website includes self-help "toolkits" giving innovators guidance on how to protect intellectual property in specific countries such as China, Korea and Mexico. For questions regarding patent enforcement issues, applicants may call the U.S. Government hotline at 1-866-999-HALT (1-866-999-4158).

LICENSE FOR FOREIGN FILING UNDER

Title 35, United States Code, Section 184

Title 37, Code of Federal Regulations, 5.11 & 5.15

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NOT GRANTED

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www.uspto.gov

APPLICATION NUMBER	FILING OR 371(C) DATE	FIRST NAMED APPLICANT	ATTY. DOCKET NO./TITLE
12/647,763	12/28/2009	Aristeidis Karalis	WTCY-0026-P07

CONFIRMATION NO. 2576

43520
STRATEGIC PATENTS P.C..
C/O PORTFOLIOIP
P.O. BOX 52050
MINNEAPOLIS, MN 55402

FORMALITIES LETTER

OC000000039637580

Date Mailed: 01/14/2010

NOTICE TO FILE MISSING PARTS OF NONPROVISIONAL APPLICATION

FILED UNDER 37 CFR 1.53(b)

Filing Date Granted

Items Required To Avoid Abandonment:

An application number and filing date have been accorded to this application. The item(s) indicated below, however, are missing. Applicant is given **TWO MONTHS** from the date of this Notice within which to file all required items and pay any fees required below to avoid abandonment. Extensions of time may be obtained by filing a petition accompanied by the extension fee under the provisions of 37 CFR 1.136(a).

- The statutory basic filing fee is missing.
  *Applicant must submit $**330** to complete the basic filing fee for a non-small entity. If appropriate, applicant may make a written assertion of entitlement to small entity status and pay the small entity filing fee (37 CFR 1.27).*
- The oath or declaration is missing.
  A properly signed oath or declaration in compliance with 37 CFR 1.63, identifying the application by the above Application Number and Filing Date, is required.
  Note: If a petition under 37 CFR 1.47 is being filed, an oath or declaration in compliance with 37 CFR 1.63 signed by all available joint inventors, or if no inventor is available by a party with sufficient proprietary interest, is required.

The application is informal since it does not comply with the regulations for the reason(s) indicated below.

The required item(s) identified below must be timely submitted to avoid abandonment:

- A substitute specification in compliance with 37 CFR 1.52, 1.121(b)(3), and 1.125, is required. The substitute specification must be submitted with markings and be accompanied by a clean version (without markings) as set forth in 37 CFR 1.125(c) and a statement that the substitute specification contains no new matter (see 37 CFR 1.125(b)). The specification, claims, and/or abstract page(s) submitted is not acceptable and cannot be scanned or properly stored because:
  - The application papers (including any electronically submitted papers) are not in compliance with 37 CFR 1.52 because pages 19-25, 30-32, 34, 47, 48, 58, 72, 74 contain text that is written in unacceptable font or font size. The text must be written in nonscript type font (e.g., Arial, Times Roman, Courier, preferably a font size of 12) lettering style having capital letters that should be at least 0.3175 cm. (0.125 inch) high. A font with capital letters smaller than 0.3175 cm. (0.125 inch) high is only acceptable if the writing is clear and legible.

Applicant is cautioned that correction of the above items may cause the specification and drawings page count to exceed 100 pages. If the specification and drawings exceed 100 pages, applicant will need to submit the required application size fee.

The applicant needs to satisfy supplemental fees problems indicated below.

The required item(s) identified below must be timely submitted to avoid abandonment:

- To avoid abandonment, a surcharge (for late submission of filing fee, search fee, examination fee or oath or declaration) as set forth in 37 CFR 1.16(f) of $**130** for a non-small entity, must be submitted with the missing items identified in this notice.

SUMMARY OF FEES DUE:

Total additional fee(s) required for this application is $**1490** for a non-small entity

- $**330** Statutory basic filing fee.
- $**130** Surcharge.
- The application search fee has not been paid. Applicant must submit $**540** to complete the search fee.
- The application examination fee has not been paid. Applicant must submit $**220** to complete the examination fee for a non-small entity.
- The specification and drawings submitted electronically contain the equivalent of more than 100 pages. Applicant owes $**270** for **47** pages in excess of **100** pages for a non-small entity.

Replies should be mailed to:

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P.O. Box 1450
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https://sportal.uspto.gov/authenticate/AuthenticateUserLocalEPF.html

For more information about EFS-Web please call the USPTO Electronic Business Center at **1-866-217-9197** or visit our website at http://www.uspto.gov/ebc.

If you are not using EFS-Web to submit your reply, you must include a copy of this notice.

/schin/

Office of Data Management, Application Assistance Unit (571) 272-4000, or (571) 272-4200, or 1-888-786-0101

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APPLICATION NUMBER	FILING OR 371(C) DATE	FIRST NAMED APPLICANT	ATTY. DOCKET NO./TITLE
12/647,763	12/28/2009	Aristeidis Karalis	WTCY-0026-P07

CONFIRMATION NO. 2576

FORMALITIES LETTER

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STRATEGIC PATENTS P.C..
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MINNEAPOLIS, MN 55402

Date Mailed: 01/14/2010

NOTICE TO FILE MISSING PARTS OF NONPROVISIONAL APPLICATION

FILED UNDER 37 CFR 1.53(b)

Filing Date Granted

Items Required To Avoid Abandonment:

An application number and filing date have been accorded to this application. The item(s) indicated below, however, are missing. Applicant is given **TWO MONTHS** from the date of this Notice within which to file all required items and pay any fees required below to avoid abandonment. Extensions of time may be obtained by filing a petition accompanied by the extension fee under the provisions of 37 CFR 1.136(a).

- The statutory basic filing fee is missing.
  *Applicant must submit $**330** to complete the basic filing fee for a non-small entity. If appropriate, applicant may make a written assertion of entitlement to small entity status and pay the small entity filing fee (37 CFR 1.27).*
- The oath or declaration is missing.
  A properly signed oath or declaration in compliance with 37 CFR 1.63, identifying the application by the above Application Number and Filing Date, is required.
  Note: If a petition under 37 CFR 1.47 is being filed, an oath or declaration in compliance with 37 CFR 1.63 signed by all available joint inventors, or if no inventor is available by a party with sufficient proprietary interest, is required.

The application is informal since it does not comply with the regulations for the reason(s) indicated below.

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  - The application papers (including any electronically submitted papers) are not in compliance with 37 CFR 1.52 because pages 19-25, 30-32, 34, 47, 48, 58, 72, 74 contain text that is written in unacceptable font or font size. The text must be written in nonscript type font (e.g., Arial, Times Roman, Courier, preferably a font size of 12) lettering style having capital letters that should be at least 0.3175 cm. (0.125 inch) high. A font with capital letters smaller than 0.3175 cm. (0.125 inch) high is only acceptable if the writing is clear and legible.

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Applicant is cautioned that correction of the above items may cause the specification and drawings page count to exceed 100 pages. If the specification and drawings exceed 100 pages, applicant will need to submit the required application size fee.

The applicant needs to satisfy supplemental fees problems indicated below.

The required item(s) identified below must be timely submitted to avoid abandonment:

- To avoid abandonment, a surcharge (for late submission of filing fee, search fee, examination fee or oath or declaration) as set forth in 37 CFR 1.16(f) of **$130** for a non-small entity, must be submitted with the missing items identified in this notice.

SUMMARY OF FEES DUE:

Total additional fee(s) required for this application is **$1490** for a non-small entity

- **$330** Statutory basic filing fee.
- **$130** Surcharge.
- The application search fee has not been paid. Applicant must submit **$540** to complete the search fee.
- The application examination fee has not been paid. Applicant must submit **$220** to complete the examination fee for a non-small entity.
- The specification and drawings submitted electronically contain the equivalent of more than 100 pages. Applicant owes **$270** for **47** pages in excess of **100** pages for a non-small entity.

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United States Patent Application

COMBINED DECLARATION AND POWER OF ATTORNEY

As a below named inventor I hereby declare with respect to the U.S. patent application entitled

WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS

the specification of which was filed on December 28, 2009 as application serial no. 12/647,763, that:

(a) my residence, post office address, and citizenship are as stated below next to my name;
(b) I believe I am the original, first and joint inventor of the subject matter which is claimed and for which a patent is sought; and
(c) I have reviewed and understood the contents of the specification, including the claims, as amended by any amendment referred to above.

I acknowledge the duty to disclose information which is material to the patentability of this application as defined in 37 C.F.R. § 1.56 (attached hereto). I also acknowledge my duty to disclose all information known to be material to patentability which became available between a filing date of a prior application and the national or PCT international filing date in the event this is a Continuation-In-Part application in accordance with 37 C.F.R. § 1.63(e).

I hereby appoint the following attorney(s) and/or patent agent(s) to prosecute this application and to transact all business in the Patent and Trademark Office connected herewith. I hereby authorize them to act and rely on instructions from and communicate directly with the person/assignee/attorney/agent/firm/organization who/which first sends/sent this case to them and by whom/which I hereby declare that I have consented after full disclosure to be represented:

All Attorneys and Agents Associated With:
Customer Number: 43520
(Strategic Patents, P.C.)

Please direct all correspondence in this case to:

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I hereby declare that all statements made herein of my own knowledge are true and that all statements made on information and belief are believed to be true; and further that these statements were made with the knowledge that willful false statements and the like so made are punishable by fine or imprisonment, or both, under Section 1001 of Title 18 of the United States Code and that such willful false statements may jeopardize the validity of the application or any patent issued thereon.

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§ 1.56 Duty to disclose information material to patentability.

(a) A patent by its very nature is affected with a public interest. The public interest is best served, and the most effective patent examination occurs when, at the time an application is being examined, the Office is aware of and evaluates the teachings of all information material to patentability. Each individual associated with the filing and prosecution of a patent application has a duty of candor and good faith in dealing with the Office, which includes a duty to disclose to the Office all information known to that individual to be material to patentability as defined in this section. The duty to disclose information exists with respect to each pending claim until the claim is canceled or withdrawn from consideration, or the application becomes abandoned. Information material to the patentability of a claim that is canceled or withdrawn from consideration need not be submitted if the information is not material to the patentability of any claim remaining under consideration in the application. There is no duty to submit information which is not material to the patentability of any existing claim. The duty to disclose all information known to be material to patentability is deemed to be satisfied if all information known to be material to patentability of any claim issued in a patent was cited by the Office or submitted to the Office in the manner prescribed by §§ 1.97(b)-(d) and 1.98. However, no patent will be granted on an application in connection with which fraud on the Office was practiced or attempted or the duty of disclosure was violated through bad faith or intentional misconduct. The Office encourages applicants to carefully examine:

(1) prior art cited in search reports of a foreign patent office in a counterpart application, and

(2) the closest information over which individuals associated with the filing or prosecution of a patent application believe any pending claim patentably defines, to make sure that any material information contained therein is disclosed to the Office.

(b) Under this section, information is material to patentability when it is not cumulative to information already of record or being made of record in the application, and

(1) It establishes, by itself or in combination with other information, a prima facie case of unpatentability of a claim; or

(2) It refutes, or is inconsistent with, a position the applicant takes in:

(i) Opposing an argument of unpatentability relied on by the Office, or

(ii) Asserting an argument of patentability.

A prima facie case of unpatentability is established when the information compels a conclusion that a claim is unpatentable under the preponderance of evidence, burden-of-proof standard, giving each term in the claim its broadest reasonable construction consistent with the specification, and before any consideration is given to evidence which may be submitted in an attempt to establish a contrary conclusion of patentability.

(c) Individuals associated with the filing or prosecution of a patent application within the meaning of this section are:

(1) Each inventor named in the application:

(2) Each attorney or agent who prepares or prosecutes the application; and

(3) Every other person who is substantively involved in the preparation or prosecution of the application and who is associated with the inventor, with the assignee or with anyone to whom there is an obligation to assign the application.

(d) Individuals other than the attorney, agent or inventor may comply with this section by disclosing information to the attorney, agent, or inventor.

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

In re application of Aristeidis Karalis *et al.*	: Group Art Unit: 2828
Serial No. 12/647,763	: Confirmation No. 2576
Filed: December 28, 2009	: Examiner: Not yet known

PRELIMINARY AMENDMENT

Mail Stop Missing Parts
Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

Sir/Madam:

Please enter the following amendments prior to examination on the merits:

Remarks begin on page 2 of this paper.

Remarks

As requested in the Notice to File Missing Parts and in compliance with 37 C.F.R. § 1.52, 1.121(b)(3), and 1.125, a substitute specification is provided as requested by the Notice To File Missing Parts of Non-Provisional Application dated January 14, 2010. In compliance with 37 C.F.R. § 1.125(c), a marked copy of the specification showing all changes relative to the originally filed specification is provided along with a non-marked clean version of the substitute specification. As requested in the Notice To File Missing Parts, the applicant has increased the font size on pages 19-25, 30-32, 34, 47, 48, 58, 72, and 74 for text that includes unacceptable font sizes. More specifically, the applicant has incresed the font size of the equations on these pages. The applicant believes that these equations now include font sizes that are clear and legible. The applicant also notes that since the substitute specification includes only font size corrections, there is no added or deleted text requiring text marking indications in the Marked Specification relative to the filed specification. No new matter is added by this substitute specification.

The Commissioner is hereby authorized to charge any additional required fees to Deposit Account No. 50-5087 in order to have this amendment considered.

Respectfully submitted
STRATEGIC PATENTS, P.C.

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March 15, 2010

WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS

CROSS-REFERENCE TO RELATED APPLICATIONS

[0001] This application is a continuation-in-part of the following U.S. patent application, U.S. 12/567716 filed September 25, 2009 which claims the benefit of the following U.S. provisional applications, U.S. App. No. 61/100,721 filed September 27,2008; U.S. App. No. 61/108,743 filed October 27, 2008; U.S. App. No. 61/147,386 filed January 26, 2009; U.S. App. No. 61/152,086 filed February 12, 2009; U.S. App. No. 61/178,508 filed May 15, 2009; U.S. App. No. 61/182,768 filed June 1, 2009; U.S. App. No. 61/121,159 filed December 9, 2008; U.S. App. No. 61/142,977 filed January 7, 2009; U.S. App. No. 61/142,885 filed January 6, 2009; U.S. App. No. 61/142,796 filed January 6, 2009; U.S. App. No. 61/142,889 filed January 6, 2009; U.S. App. No. 61/142,880 filed January 6, 2009; U.S. App. No. 61/142,818 filed January 6, 2009; U.S. App. No. 61/142,887 filed January 6, 2009; U.S. App. No. 61/156,764 filed March 2, 2009; U.S. App. No. 61/143,058 filed January 7, 2009; U.S. App. No. 61/152,390 filed February 13, 2009; U.S. App. No. 61/163,695 filed March 26, 2009; U.S. App. No. 61/172,633 filed April 24, 2009; U.S. App. No. 61/169,240 filed April 14, 2009, U.S. App. No. 61/173,747 filed April 29, 2009.

[0002] Each of the foregoing applications is incorporated herein by reference in its entirety.

BACKGROUND

[0003] Field:

[0004] This disclosure relates to wireless energy transfer, also referred to as wireless power transmission.

[0005] Description of the Related Art:

[0006] Energy or power may be transferred wirelessly using a variety of known radiative, or far-field, and non-radiative, or near-field, techniques. For example, radiative wireless information transfer using low-directionality antennas, such as those used in radio and cellular communications systems and home computer networks, may be considered wireless energy transfer. However, this type of radiative transfer is very inefficient because only a tiny

portion of the supplied or radiated power, namely, that portion in the direction of, and overlapping with, the receiver is picked up. The vast majority of the power is radiated away in all the other directions and lost in free space. Such inefficient power transfer may be acceptable for data transmission, but is not practical for transferring useful amounts of electrical energy for the purpose of doing work, such as for powering or charging electrical devices. One way to improve the transfer efficiency of some radiative energy transfer schemes is to use directional antennas to confine and preferentially direct the radiated energy towards a receiver. However, these directed radiation schemes may require an uninterruptible line-of-sight and potentially complicated tracking and steering mechanisms in the case of mobile transmitters and/or receivers. In addition, such schemes may pose hazards to objects or people that cross or intersect the beam when modest to high amounts of power are being transmitted. A known non-radiative, or near-field, wireless energy transfer scheme, often referred to as either induction or traditional induction, does not (intentionally) radiate power, but uses an oscillating current passing through a primary coil, to generate an oscillating magnetic near-field that induces currents in a near-by receiving or secondary coil. Traditional induction schemes have demonstrated the transmission of modest to large amounts of power, however only over very short distances, and with very small offset tolerances between the primary power supply unit and the secondary receiver unit. Electric transformers and proximity chargers are examples of devices that utilize this known short range, near-field energy transfer scheme.

[0007] Therefore a need exists for a wireless power transfer scheme that is capable of transferring useful amounts of electrical power over mid-range distances or alignment offsets. Such a wireless power transfer scheme should enable useful energy transfer over greater distances and alignment offsets than those realized with traditional induction schemes, but without the limitations and risks inherent in radiative transmission schemes.

SUMMARY

[0008] There is disclosed herein a non-radiative or near-field wireless energy transfer scheme that is capable of transmitting useful amounts of power over mid-range distances and alignment offsets. This inventive technique uses coupled electromagnetic resonators with long-lived oscillatory resonant modes to transfer power from a power supply to a power drain. The technique is general and may be applied to a wide range of resonators, even where the specific

examples disclosed herein relate to electromagnetic resonators. If the resonators are designed such that the energy stored by the electric field is primarily confined within the structure and that the energy stored by the magnetic field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant magnetic near-field. These types of resonators may be referred to as magnetic resonators. If the resonators are designed such that the energy stored by the magnetic field is primarily confined within the structure and that the energy stored by the electric field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant electric near-field. These types of resonators may be referred to as electric resonators. Either type of resonator may also be referred to as an electromagnetic resonator. Both types of resonators are disclosed herein.

[0009] The omni-directional but stationary (non-lossy) nature of the near-fields of the resonators we disclose enables efficient wireless energy transfer over mid-range distances, over a wide range of directions and resonator orientations, suitable for charging, powering, or simultaneously powering and charging a variety of electronic devices. As a result, a system may have a wide variety of possible applications where a first resonator, connected to a power source, is in one location, and a second resonator, potentially connected to electrical/electronic devices, batteries, powering or charging circuits, and the like, is at a second location, and where the distance from the first resonator to the second resonator is on the order of centimeters to meters. For example, a first resonator connected to the wired electricity grid could be placed on the ceiling of a room, while other resonators connected to devices, such as robots, vehicles, computers, communication devices, medical devices, and the like, move about within the room, and where these devices are constantly or intermittently receiving power wirelessly from the source resonator. From this one example, one can imagine many applications where the systems and methods disclosed herein could provide wireless power across mid-range distances, including consumer electronics, industrial applications, infrastructure power and lighting, transportation vehicles, electronic games, military applications, and the like.

[0010] Energy exchange between two electromagnetic resonators can be optimized when the resonators are tuned to substantially the same frequency and when the losses in the system are minimal. Wireless energy transfer systems may be designed so that the "coupling-time" between resonators is much shorter than the resonators' "loss-times". Therefore, the systems and methods described herein may utilize high quality factor (high-Q) resonators with

low intrinsic-loss rates. In addition, the systems and methods described herein may use sub-wavelength resonators with near-fields that extend significantly longer than the characteristic sizes of the resonators, so that the near-fields of the resonators that exchange energy overlap at mid-range distances. This is a regime of operation that has not been practiced before and that differs significantly from traditional induction designs.

[0011] It is important to appreciate the difference between the high-Q magnetic resonator scheme disclosed here and the known close-range or proximity inductive schemes, namely, that those known schemes do not conventionally utilize high-Q resonators. Using coupled-mode theory (CMT), (see, for example, *Waves and Fields in Optoelectronics*, H.A. Haus, Prentice Hall, 1984), one may show that a high-Q resonator-coupling mechanism can enable orders of magnitude more efficient power delivery between resonators spaced by mid-range distances than is enabled by traditional inductive schemes. Coupled high-Q resonators have demonstrated efficient energy transfer over mid-range distances and improved efficiencies and offset tolerances in short range energy transfer applications.

[0012] The systems and methods described herein may provide for near-field wireless energy transfer via strongly coupled high-Q resonators, a technique with the potential to transfer power levels from picowatts to kilowatts, safely, and over distances much larger than have been achieved using traditional induction techniques. Efficient energy transfer may be realized for a variety of general systems of strongly coupled resonators, such as systems of strongly coupled acoustic resonators, nuclear resonators, mechanical resonators, and the like, as originally described by researchers at M.I.T. in their publications, "Efficient wireless non-radiative mid-range energy transfer", *Annals of Physics*, vol. 323, Issue 1, p. 34 (2008) and "Wireless Power Transfer via Strongly Coupled Magnetic Resonances", *Science*, vol. 317, no. 5834, p. 83, (2007). Disclosed herein are electromagnetic resonators and systems of coupled electromagnetic resonators, also referred to more specifically as coupled magnetic resonators and coupled electric resonators, with operating frequencies below 10 GHz.

[0013] This disclosure describes wireless energy transfer technologies, also referred to as wireless power transmission technologies. Throughout this disclosure, we may use the terms wireless energy transfer, wireless power transfer, wireless power transmission, and the like, interchangeably. We may refer to supplying energy or power from a source, an AC or DC source, a battery, a source resonator, a power supply, a generator, a solar panel, and thermal

collector, and the like, to a device, a remote device, to multiple remote devices, to a device resonator or resonators, and the like. We may describe intermediate resonators that extend the range of the wireless energy transfer system by allowing energy to hop, transfer through, be temporarily stored, be partially dissipated, or for the transfer to be mediated in any way, from a source resonator to any combination of other device and intermediate resonators, so that energy transfer networks, or strings, or extended paths may be realized. Device resonators may receive energy from a source resonator, convert a portion of that energy to electric power for powering or charging a device, and simultaneously pass a portion of the received energy onto other device or mobile device resonators. Energy may be transferred from a source resonator to multiple device resonators, significantly extending the distance over which energy may be wirelessly transferred. The wireless power transmission systems may be implemented using a variety of system architectures and resonator designs. The systems may include a single source or multiple sources transmitting power to a single device or multiple devices. The resonators may be designed to be source or device resonators, or they may be designed to be repeaters. In some cases, a resonator may be a device and source resonator simultaneously, or it may be switched from operating as a source to operating as a device or a repeater. One skilled in the art will understand that a variety of system architectures may be supported by the wide range of resonator designs and functionalities described in this application.

[0014] In the wireless energy transfer systems we describe, remote devices may be powered directly, using the wirelessly supplied power or energy, or the devices may be coupled to an energy storage unit such as a battery, a super-capacitor, an ultra-capacitor, or the like (or other kind of power drain), where the energy storage unit may be charged or re-charged wirelessly, and/or where the wireless power transfer mechanism is simply supplementary to the main power source of the device. The devices may be powered by hybrid battery/energy storage devices such as batteries with integrated storage capacitors and the like. Furthermore, novel battery and energy storage devices may be designed to take advantage of the operational improvements enabled by wireless power transmission systems.

[0015] Other power management scenarios include using wirelessly supplied power to recharge batteries or charge energy storage units while the devices they power are turned off, in an idle state, in a sleep mode, and the like. Batteries or energy storage units may be charged or recharged at high (fast) or low (slow) rates. Batteries or energy storage units may be trickle

charged or float charged. Multiple devices may be charged or powered simultaneously in parallel or power delivery to multiple devices may be serialized such that one or more devices receive power for a period of time after which other power delivery is switched to other devices. Multiple devices may share power from one or more sources with one or more other devices either simultaneously, or in a time multiplexed manner, or in a frequency multiplexed manner, or in a spatially multiplexed manner, or in an orientation multiplexed manner, or in any combination of time and frequency and spatial and orientation multiplexing. Multiple devices may share power with each other, with at least one device being reconfigured continuously, intermittently, periodically, occasionally, or temporarily, to operate as wireless power sources. It would be understood by one of ordinary skill in the art that there are a variety of ways to power and/or charge devices, and the variety of ways could be applied to the technologies and applications described herein.

[0016] Wireless energy transfer has a variety of possible applications including for example, placing a source (e.g. one connected to the wired electricity grid) on the ceiling, under the floor, or in the walls of a room, while devices such as robots, vehicles, computers, PDAs or similar are placed or move freely within the room. Other applications may include powering or recharging electric-engine vehicles, such as buses and/or hybrid cars and medical devices, such as wearable or implantable devices. Additional example applications include the ability to power or recharge autonomous electronics (e.g. laptops, cell-phones, portable music players, house-hold robots, GPS navigation systems, displays, etc), sensors, industrial and manufacturing equipment, medical devices and monitors, home appliances and tools (e.g. lights, fans, drills, saws, heaters, displays, televisions, counter-top appliances, etc.), military devices, heated or illuminated clothing, communications and navigation equipment, including equipment built into vehicles, clothing and protective-wear such as helmets, body armor and vests, and the like, and the ability to transmit power to physically isolated devices such as to implanted medical devices, to hidden, buried, implanted or embedded sensors or tags, to and/or from roof-top solar panels to indoor distribution panels, and the like.

[0017] In one aspect, disclosed herein is a system including a source resonator having a Q-factor Q_1 and a characteristic size x_1, coupled to a power generator with direct electrical connections; and a second resonator having a Q-factor Q_2 and a characteristic size x_2, coupled to a load with direct electrical connections, and located a distance D from the source resonator,

wherein the source resonator and the second resonator are coupled to exchange energy wirelessly among the source resonator and the second resonator in order to transmit power from the power generator to the load, and wherein $\sqrt{Q_1 Q_2}$ is greater than 100.

[0018] Q_1 may be greater than 100 and Q_2 may be less than 100. Q_1 may be greater than 100 and Q_2 may be greater than 100. A useful energy exchange may be maintained over an operating distance from 0 to D, where D is larger than the smaller of x_1 and x_2. At least one of the source resonator and the second resonator may be a coil of at least one turn of a conducting material connected to a first network of capacitors. The first network of capacitors may include at least one tunable capacitor. The direct electrical connections of at least one of the source resonator to the ground terminal of the power generator and the second resonator to the ground terminal of the load may be made at a point on an axis of electrical symmetry of the first network of capacitors. The first network of capacitors may include at least one tunable butterfly-type capacitor, wherein the direct electrical connection to the ground terminal is made on a center terminal of the at least one tunable butterfly-type capacitor. The direct electrical connection of at least one of the source resonator to the power generator and the second resonator to the load may be made via a second network of capacitors, wherein the first network of capacitors and the second network of capacitors form an impedance matching network. The impedance matching network may be designed to match the coil to a characteristic impedance of the power generator or the load at a driving frequency of the power generator.

[0019] At least one of the first network of capacitors and the second network of capacitors may include at least one tunable capacitor. The first network of capacitors and the second network of capacitors may be adjustable to change an impedance of the impedance matching network at a driving frequency of the power generator. The first network of capacitors and the second network of capacitors may be adjustable to match the coil to the characteristic impedance of the power generator or the load at a driving frequency of the power generator. At least one of the first network of capacitors and the second network of capacitors may include at least one fixed capacitor that reduces a voltage across the at least one tunable capacitor. The direct electrical connections of at least one of the source resonator to the power generator and the second resonator to the load may be configured to substantially preserve a resonant mode. At least one of the source resonator and the second resonator may be a tunable resonator. The source resonator may be physically separated from the power generator and the second resonator

may be physically separated from the load. The second resonator may be coupled to a power conversion circuit to deliver DC power to the load. The second resonator may be coupled to a power conversion circuit to deliver AC power to the load. The second resonator may be coupled to a power conversion circuit to deliver both AC and DC power to the load. The second resonator may be coupled to a power conversion circuit to deliver power to a plurality of loads.

[0020] In another aspect, a system disclosed herein includes a source resonator having a *Q*-factor Q_1 and a characteristic size x_1, and a second resonator having a *Q*-factor Q_2 and a characteristic size x_2, and located a distance D from the source resonator; wherein the source resonator and the second resonator are coupled to exchange energy wirelessly among the source resonator and the second resonator; and wherein $\sqrt{Q_1 Q_2}$ is greater than 100, and wherein at least one of the resonators is enclosed in a low loss tangent material.

[0021] In another aspect, a system disclosed herein includes a source resonator having a *Q*-factor Q_1 and a characteristic size x_1, and a second resonator having a *Q*-factor Q_2 and a characteristic size x_2, and located a distance D from the source resonator; wherein the source resonator and the second resonator are coupled to exchange energy wirelessly among the source resonator and the second resonator, and wherein $\sqrt{Q_1 Q_2}$ is greater than 100; and wherein at least one of the resonators includes a coil of a plurality of turns of a conducting material connected to a network of capacitors, wherein the plurality of turns are in a common plane, and wherein a characteristic thickness of the at least one of the resonators is much less than a characteristic size of the at least one of the resonators.

[0022] In embodiments, the present invention may provide for a method and system comprising a source resonator optionally coupled to an energy source and a second resonator located a distance from the source resonator, where the source resonator and the second resonator are coupled to provide near-field wireless energy transfer among the source resonator and the second resonator and where the field of at least one of the source resonator and the second resonator is shaped to avoid a loss-inducing object. In embodiments, at least one of the source resonator and the second resonator may have a quality factor, Q>100. The source resonator Q may be greater than 100 and the second resonator Q may be greater than 100. The square root of the source resonator Q times the second resonator Q may be greater than 100. In embodiments, there may be more than one source resonator, more than one second resonator, more than three resonators, and the like.

[0023] Throughout this disclosure we may refer to the certain circuit components such as capacitors, inductors, resistors, diodes, switches and the like as circuit components or elements. We may also refer to series and parallel combinations of these components as elements, networks, topologies, circuits, and the like. We may describe combinations of capacitors, diodes, varactors, transistors, and/or switches as adjustable impedance networks, tuning networks, matching networks, adjusting elements, and the like. We may also refer to "self-resonant" objects that have both capacitance, and inductance distributed (or partially distributed, as opposed to solely lumped) throughout the entire object. It would be understood by one of ordinary skill in the art that adjusting and controlling variable components within a circuit or network may adjust the performance of that circuit or network and that those adjustments may be described generally as tuning, adjusting, matching, correcting, and the like. Other methods to tune or adjust the operating point of the wireless power transfer system may be used alone, or in addition to adjusting tunable components such as inductors and capacitors, or banks of inductors and capacitors.

[0024] Unless otherwise defined, all technical and scientific terms used herein have the same meaning as commonly understood by one of ordinary skill in the art to which this disclosure belongs. In case of conflict with publications, patent applications, patents, and other references mentioned or incorporated herein by reference, the present specification, including definitions, will control.

[0025] Any of the features described above may be used, alone or in combination, without departing from the scope of this disclosure. Other features, objects, and advantages of the systems and methods disclosed herein will be apparent from the following detailed description and figures.

BRIEF DESCRIPTION OF FIGURES

[0026] Fig. 1 (a) and (b) depict exemplary wireless power systems containing a source resonator 1 and device resonator 2 separated by a distance D.

[0027] Fig. 2 shows an exemplary resonator labeled according to the labeling convention described in this disclosure. Note that there are no extraneous objects or additional resonators shown in the vicinity of resonator 1.

[0028] Fig. 3 shows an exemplary resonator in the presence of a “loading” object, labeled according to the labeling convention described in this disclosure.

[0029] Fig. 4 shows an exemplary resonator in the presence of a “perturbing” object, labeled according to the labeling convention described in this disclosure.

[0030] Fig. 5 shows a plot of efficiency, η, vs. strong coupling factor, $U=\kappa/\sqrt{\Gamma_s\Gamma_d}=k\sqrt{Q_sQ_d}$.

[0031] Fig. 6 (a) shows a circuit diagram of one example of a resonator (b) shows a diagram of one example of a capacitively-loaded inductor loop magnetic resonator, (c) shows a drawing of a self-resonant coil with distributed capacitance and inductance, (d) shows a simplified drawing of the electric and magnetic field lines associated with an exemplary magnetic resonator of the current disclosure, and (e) shows a diagram of one example of an electric resonator.

[0032] Fig. 7 shows a plot of the “quality factor”, Q (solid line), as a function of frequency, of an exemplary resonator that may be used for wireless power transmission at MHz frequencies. The absorptive Q (dashed line) increases with frequency, while the radiative Q (dotted line) decreases with frequency, thus leading the overall Q to peak at a particular frequency.

[0033] Fig. 8 shows a drawing of a resonator structure with its characteristic size, thickness and width indicated.

[0034] Fig. 9 (a) and (b) show drawings of exemplary inductive loop elements.

[0035] Fig. 10 (a) and (b) show two examples of trace structures formed on printed circuit boards and used to realize the inductive element in magnetic resonator structures.

[0036] Fig. 11 (a) shows a perspective view diagram of a planar magnetic resonator, (b) shows a perspective view diagram of a two planar magnetic resonator with various geometries, and c) shows is a perspective view diagram of a two planar magnetic resonators separated by a distance D.

[0037] Fig. 12 is a perspective view of an example of a planar magnetic resonator.

[0038] Fig. 13 is a perspective view of a planar magnetic resonator arrangement with a circular resonator coil.

[0039] Fig. 14 is a perspective view of an active area of a planar magnetic resonator.

[0040] Fig. 15 is a perspective view of an application of the wireless power transfer system with a source at the center of a table powering several devices placed around the source.

[0041] Fig. 16(a) shows a 3D finite element model of a copper and magnetic material structure driven by a square loop of current around the choke point at its center. In this example, a structure may be composed of two boxes made of a conducting material such as copper, covered by a layer of magnetic material, and connected by a block of magnetic material. The inside of the two conducting boxes in this example would be shielded from AC electromagnetic fields generated outside the boxes and may house lossy objects that might lower the Q of the resonator or sensitive components that might be adversely affected by the AC electromagnetic fields. Also shown are the calculated magnetic field streamlines generated by this structure, indicating that the magnetic field lines tend to follow the lower reluctance path in the magnetic material. Fig. 16(b) shows interaction, as indicated by the calculated magnetic field streamlines, between two identical structures as shown in (a). Because of symmetry, and to reduce computational complexity, only one half of the system is modeled (but the computation assumes the symmetrical arrangement of the other half).

[0042] Fig. 17 shows an equivalent circuit representation of a magnetic resonator including a conducting wire wrapped N times around a structure, possibly containing magnetically permeable material. The inductance is realized using conducting loops wrapped around a structure comprising a magnetic material and the resistors represent loss mechanisms in the system (R_{wire} for resistive losses in the loop, R_{μ} denoting the equivalent series resistance of the structure surrounded by the loop). Losses may be minimized to realize high-Q resonators.

[0043] Fig. 18 shows a Finite Element Method (FEM) simulation of two high conductivity surfaces above and below a disk composed of lossy dielectric material, in an external magnetic field of frequency 6.78 MHz. Note that the magnetic field was uniform before the disk and conducting materials were introduced to the simulated environment. This simulation is performed in cylindrical coordinates. The image is azimuthally symmetric around the $r = 0$ axis. The lossy dielectric disk has $\epsilon_r = 1$ and $\sigma = 10$ S/m.

[0044] Fig. 19 shows a drawing of a magnetic resonator with a lossy object in its vicinity completely covered by a high-conductivity surface.

[0045] Fig. 20 shows a drawing of a magnetic resonator with a lossy object in its vicinity partially covered by a high-conductivity surface.

[0046] Fig. 21 shows a drawing of a magnetic resonator with a lossy object in its vicinity placed on top of a high-conductivity surface.

[0047] Fig. 22 shows a diagram of a completely wireless projector.

[0048] Fig. 23 shows the magnitude of the electric and magnetic fields along a line that contains the diameter of the circular loop inductor and along the axis of the loop inductor.

[0049] Fig. 24 shows a drawing of a magnetic resonator and its enclosure along with a necessary but lossy object placed either (a) in the corner of the enclosure, as far away from the resonator structure as possible or (b) in the center of the surface enclosed by the inductive element in the magnetic resonator.

[0050] Fig. 25 shows a drawing of a magnetic resonator with a high-conductivity surface above it and a lossy object, which may be brought into the vicinity of the resonator, but above the high-conductivity sheet.

[0051] Fig. 26(a) shows an axially symmetric FEM simulation of a thin conducting (copper) cylinder or disk (20 cm in diameter, 2 cm in height) exposed to an initially uniform, externally applied magnetic field (gray flux lines) along the z-axis. The axis of symmetry is at r=0. The magnetic streamlines shown originate at $z = -\infty$, where they are spaced from r=3 cm to r=10 cm in intervals of 1 cm. The axes scales are in meters. Fig. 26 (b) shows the same structure and externally applied field as in (a), except that the conducting cylinder has been modified to include a 0.25 mm layer of magnetic material (not visible) with μ_r'=40, on its outside surface. Note that the magnetic streamlines are deflected away from the cylinder significantly less than in (a).

[0052] Fig. 27 shows an axi-symmetric view of a variation based on the system shown in Fig. 26. Only one surface of the lossy material is covered by a layered structure of copper and magnetic materials. The inductor loop is placed on the side of the copper and magnetic material structure opposite to the lossy material as shown.

[0053] Fig. 28 (a) depicts a general topology of a matching circuit including an indirect coupling to a high-Q inductive element.

[0054] Fig. 28 (b) shows a block diagram of a magnetic resonator that includes a conductor loop inductor and a tunable impedance network. Physical electrical connections to this resonator may be made to the terminal connections.

[0055] Fig. 28 (c) depicts a general topology of a matching circuit directly coupled to a high-Q inductive element.

[0056] Fig. 28 (d) depicts a general topology of a symmetric matching circuit directly coupled to a high-Q inductive element and driven anti-symmetrically (balanced drive).

[0057] Fig. 28 (e) depicts a general topology of a matching circuit directly coupled to a high-Q inductive element and connected to ground at a point of symmetry of the main resonator (unbalanced drive).

[0058] Figs. 29(a) and 29(b) depict two topologies of matching circuits transformer-coupled (i.e. indirectly or inductively) to a high-Q inductive element. The highlighted portion of the Smith chart in (c) depicts the complex impedances (arising from L and R of the inductive element) that may be matched to an arbitrary real impedance Z_0 by the topology of Fig. 31(b) in the case $\omega L_2=1/\omega C_2$.

[0059] Figs. 30(a),(b),(c),(d),(e),(f) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in series with Z_0. The topologies shown in Figs. 30(a),(b),(c) are driven with a common-mode signal at the input terminals, while the topologies shown in Figs 30(d),(e),(f) are symmetric and receive a balanced drive. The highlighted portion of the Smith chart in 30(g) depicts the complex impedances that may be matched by these topologies. Figs. 30(h),(i),(j),(k),(l),(m) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in series with Z_0.

[0060] Figs. 31(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in series with Z_0. They are connected to ground at the center point of a capacitor and receive an unbalanced drive. The highlighted portion of the Smith chart in Fig. 31(d) depicts the complex impedances that may be matched by these topologies. Figs. 31(e),(f),(g) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in series with Z_0.

[0061] Figs. 32(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in series with Z_0. They are connected to ground by tapping at the center point of the inductor loop and receive an unbalanced drive. The highlighted portion of the Smith chart in (d) depicts the complex impedances that may be matched by these topologies, (e),(f),(g) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in series with Z_0.

[0062] Figs. 33(a),(b),(c),(d),(e),(f) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in parallel with Z_o. The topologies shown in Figs. 33(a),(b),(c) are driven with a common-mode signal at the input terminals, while the topologies shown in Figs 33(d),(e),(f) are symmetric and receive a balanced drive. The highlighted portion of the Smith chart in Fig. 33(g) depicts the complex impedances that may be matched by these topologies. Figs. 33(h),(i),(j),(k),(l),(m) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in parallel with Z_o.

[0063] Figs. 34(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in parallel with Z_0. They are connected to ground at the center point of a capacitor and receive an unbalanced drive. The highlighted portion of the Smith chart in (d) depicts the complex impedances that may be matched by these topologies. Figs. 34(e),(f),(g) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in parallel with Z_0.

[0064] Figs. 35(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in parallel with Z_0. They are connected to ground by tapping at the center point of the inductor loop and receive an unbalanced drive. The highlighted portion of the Smith chart in Figs. 35(d),(e), and (f) depict the complex impedances that may be matched by these topologies.

[0065] Figs. 36(a),(b),(c),(d) depict four topologies of networks of fixed and variable capacitors designed to produce an overall variable capacitance with finer tuning resolution and some with reduced voltage on the variable capacitor.

[0066] Figs. 37(a) and 37(b) depict two topologies of networks of fixed capacitors and a variable inductor designed to produce an overall variable capacitance.

[0067] Fig. 38 depicts a high level block diagram of a wireless power transmission system.

[0068] Fig. 39 depicts a block diagram of an exemplary wirelessly powered device.

[0069] Fig. 40 depicts a block diagram of the source of an exemplary wireless power transfer system.

[0070] Fig. 41 shows an equivalent circuit diagram of a magnetic resonator. The slash through the capacitor symbol indicates that the represented capacitor may be fixed or variable.

The port parameter measurement circuitry may be configured to measure certain electrical signals and may measure the magnitude and phase of signals.

[0071] Fig. 42 shows a circuit diagram of a magnetic resonator where the tunable impedance network is realized with voltage controlled capacitors. Such an implementation may be adjusted, tuned or controlled by electrical circuits including programmable or controllable voltage sources and/or computer processors. The voltage controlled capacitors may be adjusted in response to data measured by the port parameter measurement circuitry and processed by measurement analysis and control algorithms and hardware. The voltage controlled capacitors may be a switched bank of capacitors.

[0072] Fig. 43 shows an end-to-end wireless power transmission system. In this example, both the source and the device contain port measurement circuitry and a processor. The box labeled "coupler/switch" indicates that the port measurement circuitry may be connected to the resonator by a directional coupler or a switch, enabling the measurement, adjustment and control of the source and device resonators to take place in conjunction with, or separate from, the power transfer functionality.

[0073] Fig. 44 shows an end-to-end wireless power transmission system. In this example, only the source contains port measurement circuitry and a processor. In this case, the device resonator operating characteristics may be fixed or may be adjusted by analog control circuitry and without the need for control signals generated by a processor.

[0074] Fig. 45 shows an end-to-end wireless power transmission system. In this example, both the source and the device contain port measurement circuitry but only the source contains a processor. Data from the device is transmitted through a wireless communication channel, which could be implemented either with a separate antenna, or through some modulation of the source drive signal.

[0075] Fig. 46 shows an end-to-end wireless power transmission system. In this example, only the source contains port measurement circuitry and a processor. Data from the device is transmitted through a wireless communication channel, which could be implemented either with a separate antenna, or through some modulation of the source drive signal.

[0076] Fig. 47 shows coupled magnetic resonators whose frequency and impedance may be automatically adjusted using algorithms implemented using a processor or a computer.

[0077] Fig. 48 shows a varactor array.

[0078] Fig. 49 shows a device (laptop computer) being wirelessly powered or charged by a source, where both the source and device resonator are physically separated from, but electrically connected to, the source and device.

[0079] Fig. 50 (a) is an illustration of a wirelessly powered or charged laptop application where the device resonator is inside the laptop case and is not visible.

[0080] Fig. 50 (b) is an illustration of a wirelessly powered or charged laptop application where the resonator is underneath the laptop base and is electrically connected to the laptop power input by an electrical cable.

[0081] Fig. 50 (c) is an illustration of a wirelessly powered or charged laptop application where the resonator is attached to the laptop base.

[0082] Fig. 50 (d) is an illustration of a wirelessly powered or charged laptop application where the resonator is attached to the laptop display.

[0083] Fig. 51 is a diagram of rooftop PV panels with wireless power transfer.

DETAILED DESCRIPTION

[0084] As described above, this disclosure relates to coupled electromagnetic resonators with long-lived oscillatory resonant modes that may wirelessly transfer power from a power supply to a power drain. However, the technique is not restricted to electromagnetic resonators, but is general and may be applied to a wide variety of resonators and resonant objects. Therefore, we first describe the general technique, and then disclose electromagnetic examples for wireless energy transfer.

[0085] **Resonators**

[0086] A resonator may be defined as a system that can store energy in at least two different forms, and where the stored energy is oscillating between the two forms. The resonance has a specific oscillation mode with a resonant (modal) frequency, *f*, and a resonant (modal) field. The angular resonant frequency, *ω*, may be defined as $\omega = 2\pi f$, the resonant wavelength, λ, may be defined as $\lambda = c/f$, where *c* is the speed of light, and the resonant period, *T*, may be defined as $T = 1/f = 2\pi/\omega$. In the absence of loss mechanisms, coupling mechanisms or external energy supplying or draining mechanisms, the total resonator stored energy, *W*, would stay fixed

and the two forms of energy would oscillate, wherein one would be maximum when the other is minimum and vice versa.

[0087] In the absence of extraneous materials or objects, the energy in the resonator 102 shown in Fig. 1 may decay or be lost by intrinsic losses. The resonator fields then obey the following linear equation:

$$\frac{da(t)}{dt} = -i(\omega - i\Gamma)a(t),$$

where the variable *a(t)* is the resonant field amplitude, defined so that the energy contained within the resonator is given by $|a(t)|^2$. Γ is the intrinsic energy decay or loss rate (e.g. due to absorption and radiation losses).

[0088] The Quality Factor, or *Q*-factor, or *Q*, of the resonator, which characterizes the energy decay, is inversely proportional to these energy losses. It may be defined as $Q=\omega*W/P$, where *P* is the time-averaged power lost at steady state. That is, a resonator 102 with a high-*Q* has relatively low intrinsic losses and can store energy for a relatively long time. Since the resonator loses energy at its intrinsic decay rate, 2Γ, its *Q*, also referred to as its intrinsic *Q*, is given by $Q = \omega/2\Gamma$. The quality factor also represents the number of oscillation periods, *T*, it takes for the energy in the resonator to decay by a factor of *e*.

[0089] As described above, we define the quality factor or *Q* of the resonator as that due only to intrinsic loss mechanisms. A subscript index such as Q_1, indicates the resonator (resonator 1 in this case) to which the *Q* refers. Fig. 2 shows an electromagnetic resonator 102 labeled according to this convention. Note that in this figure, there are no extraneous objects or additional resonators in the vicinity of resonator 1.

[0090] Extraneous objects and/or additional resonators in the vicinity of a first resonator may perturb or load the first resonator, thereby perturbing or loading the *Q* of the first resonator, depending on a variety of factors such as the distance between the resonator and object or other resonator, the material composition of the object or other resonator, the structure of the first resonator, the power in the first resonator, and the like. Unintended external energy losses or coupling mechanisms to extraneous materials and objects in the vicinity of the resonators may be referred to as "perturbing" the *Q* of a resonator, and may be indicated by a subscript within rounded parentheses, (). Intended external energy losses, associated with energy transfer via coupling to other resonators and to generators and loads in the wireless energy transfer system

may be referred to as "loading" the Q of the resonator, and may be indicated by a subscript within square brackets, [].

[0091] The Q of a resonator 102 connected or coupled to a power generator, g, or load 302 , l, may be called the "loaded quality factor" or the "loaded Q" and may be denoted by $Q_{[g]}$ or $Q_{[l]}$, as illustrated in Fig. 3. In general, there may be more than one generator or load 302 connected to a resonator 102. However, we do not list those generators or loads separately but rather use "g" and "l" to refer to the equivalent circuit loading imposed by the combinations of generators and loads. In general descriptions, we may use the subscript "l" to refer to either generators or loads connected to the resonators.

[0092] In some of the discussion herein, we define the "loading quality factor" or the "loading Q" due to a power generator or load connected to the resonator, as $\delta Q_{[l]}$, where, $1/\delta Q_{[l]} \equiv 1/Q_{[l]} - 1/Q$. Note that the larger the loading Q, $\delta Q_{[l]}$, of a generator or load, the less the loaded Q, $Q_{[l]}$, deviates from the unloaded Q of the resonator.

[0093] The Q of a resonator in the presence of an extraneous object 402, p, that is not intended to be part of the energy transfer system may be called the "perturbed quality factor" or the "perturbed Q" and may be denoted by $Q_{(p)}$, as illustrated in Fig. 4. In general, there may be many extraneous objects, denoted as $p1$, $p2$, etc., or a set of extraneous objects $\{p\}$, that perturb the Q of the resonator 102. In this case, the perturbed Q may be denoted $Q_{(p1+p2+\ldots)}$ or $Q_{(\{p\})}$. For example, $Q_{1(brick+wood)}$ may denote the perturbed quality factor of a first resonator in a system for wireless power exchange in the presence of a brick and a piece of wood, and $Q_{2(\{office\})}$ may denote the perturbed quality factor of a second resonator in a system for wireless power exchange in an office environment.

[0094] In some of the discussion herein, we define the "perturbing quality factor" or the "perturbing Q" due to an extraneous object, p, as $\delta Q_{(p)}$, where $1/\delta Q_{(p)} \equiv 1/Q_{(p)} - 1/Q$. As stated before, the perturbing quality factor may be due to multiple extraneous objects, $p1$, $p2$, etc. or a set of extraneous objects, $\{p\}$. The larger the perturbing Q, $\delta Q_{(p)}$, of an object, the less the perturbed Q, $Q_{(p)}$, deviates from the unperturbed Q of the resonator.

[0095] In some of the discussion herein, we also define $\Theta_{(p)} \equiv Q_{(p)} / Q$ and call it the "quality factor insensitivity" or the "Q-insensitivity" of the resonator in the presence of an

extraneous object. A subscript index, such as $\Theta_{1(p)}$, indicates the resonator to which the perturbed and unperturbed quality factors are referring, namely, $\Theta_{1(p)} \equiv Q_{1(p)} / Q_1$.

[0096] Note that the quality factor, Q, may also be characterized as "unperturbed", when necessary to distinguish it from the perturbed quality factor, $Q_{(p)}$, and "unloaded", when necessary to distinguish it from the loaded quality factor, $Q_{[l]}$. Similarly, the perturbed quality factor, $Q_{(p)}$, may also be characterized as "unloaded", when necessary to distinguish them from the loaded perturbed quality factor, $Q_{(p)[l]}$.

[0097] **Coupled Resonators**

[0098] Resonators having substantially the same resonant frequency, coupled through any portion of their near-fields may interact and exchange energy. There are a variety of physical pictures and models that may be employed to understand, design, optimize and characterize this energy exchange. One way to describe and model the energy exchange between two coupled resonators is using coupled mode theory (CMT).

[0099] In coupled mode theory, the resonator fields obey the following set of linear equations:

$$\frac{da_m(t)}{dt} = -i(\omega_m - i\Gamma_m)a_m(t) + i\sum_{n \neq m} \kappa_{mn} a_n(t)$$

where the indices denote different resonators and and κ_{mn} are the coupling coefficients between the resonators. For a reciprocal system, the coupling coefficients may obey the relation $\kappa_{mn} = \kappa_{nm}$. Note that, for the purposes of the present specification, far-field radiation interference effects will be ignored and thus the coupling coefficients will be considered real. Furthermore, since in all subsequent calculations of system performance in this specification the coupling coefficients appear only with their square, κ_{mn}^2, we use κ_{mn} to denote the absolute value of the real coupling coefficients.

[00100] Note that the coupling coefficient, κ_{mn}, from the CMT described above is related to the so-called coupling factor, k_{mn}, between resonators m and n by $k_{mn} = 2\kappa_{mn} / \sqrt{\omega_m \omega_n}$.

We define a "strong-coupling factor", U_{mn}, as the ratio of the coupling and loss rates between resonators *m* and *n*, by $U_{mn}=\kappa_{mn}/\sqrt{\Gamma_m\Gamma_n}=k_{mn}\sqrt{Q_mQ_n}$.

[00101] The quality factor of a resonator *m,* in the presence of a similar frequency resonator *n* or additional resonators, may be loaded by that resonator *n* or additional resonators, in a fashion similar to the resonator being loaded by a connected power generating or consuming device. The fact that resonator *m* may be loaded by resonator *n* and vice versa is simply a different way to see that the resonators are coupled.

[00102] The loaded *Q*'s of the resonators in these cases may be denoted as $Q_{m[n]}$ and $Q_{n[m]}$. For multiple resonators or loading supplies or devices, the total loading of a resonator may be determined by modeling each load as a resistive loss, and adding the multiple loads in the appropriate parallel and/or series combination to determine the equivalent load of the ensemble.

[00103] In some of the discussion herein, we define the "loading quality factor" or the "loading Q_m" of resonator *m* due to resonator *n* as $\delta Q_{m[n]}$, where $1/\delta Q_{m[n]}\equiv 1/Q_{m[n]}-1/Q_m$. Note that resonator *n* is also loaded by resonator *m* and its "loading Q_n" is given by $1/\delta Q_{n[m]}\equiv 1/Q_{n[m]}-1/Q_n$.

[00104] When one or more of the resonators are connected to power generators or loads, the set of linear equations is modified to:

$$\frac{da_m(t)}{dt}=-i\left(\omega_m-i\Gamma_m\right)a_m(t)+i\sum_{n\neq m}\kappa_{mn}a_n(t)-\kappa_m a_m(t)+\sqrt{2\kappa_m}s_{+m}(t)$$
$$s_{-m}(t)=\sqrt{2\kappa_m}a_m(t)-s_{+m}(t)\quad,$$

where $s_{+m}(t)$ and $s_{-m}(t)$ are respectively the amplitudes of the fields coming from a generator into the resonator *m* and going out of the resonator *m* either back towards the generator or into a load, defined so that the power they carry is given by $\left|s_{+m}(t)\right|^2$ and $\left|s_{-m}(t)\right|^2$. The loading coefficients κ_m relate to the rate at which energy is exchanged between the resonator *m* and the generator or load connected to it.

[00105] Note that the loading coefficient, κ_m, from the CMT described above is related to the loading quality factor, $\delta Q_{m[l]}$, defined earlier, by $\delta Q_{m[l]}=\omega_m/2\kappa_m$.

[00106] We define a "strong-loading factor", $U_{m[l]}$, as the ratio of the loading and loss rates of resonator m, $U_{m[l]} = \kappa_m / \Gamma_m = Q_m / \delta Q_{m[l]}$.

[00107] Fig. 1(a) shows an example of two coupled resonators 1000, a first resonator 102S, configured as a source resonator and a second resonator 102D, configured as a device resonator. Energy may be transferred over a distance *D* between the resonators. The source resonator 102S may be driven by a power supply or generator (not shown). Work may be extracted from the device resonator 102D by a power consuming drain or load (e.g. a load resistor, not shown). Let us use the subscripts "s" for the source, "d" for the device, "g" for the generator, and "l" for the load, and, since in this example there are only two resonators and $\kappa_{sd}=\kappa_{ds}$, let us drop the indices on κ_{sd}, k_{sd}, and U_{sd}, and denote them as κ, k, and U, respectively.

[00108] The power generator may be constantly driving the source resonator at a constant driving frequency, *f*, corresponding to an angular driving frequency, ω, where $\omega = 2\pi f$.

[00109] In this case, the efficiency, $\eta = |s_{-d}|^2 / |s_{+s}|^2$, of the power transmission from the generator to the load (via the source and device resonators) is maximized under the following conditions: The source resonant frequency, the device resonant frequency and the generator driving frequency have to be matched, namely

$$\omega_s = \omega_d = \omega.$$

Furthermore, the loading Q of the source resonator due to the generator, $\delta Q_{s[g]}$, has to be matched (equal) to the loaded Q of the source resonator due to the device resonator and the load, $Q_{s[dl]}$, and inversely the loading Q of the device resonator due to the load, $\delta Q_{d[l]}$, has to be matched (equal) to the loaded Q of the device resonator due to the source resonator and the generator, $Q_{d[sg]}$, namely

$$\delta Q_{s[g]} = Q_{s[dl]} \text{ and } \delta Q_{d[l]} = Q_{d[sg]}.$$

These equations determine the optimal loading rates of the source resonator by the generator and of the device resonator by the load as

$$U_{d[l]} = \kappa_d / \Gamma_d = Q_d / \delta Q_{d[l]} = \sqrt{1+U^2} = \sqrt{1+\left(\kappa / \sqrt{\Gamma_s \Gamma_d}\right)^2} = Q_s / \delta Q_{s[g]} = \kappa_s / \Gamma_s = U_{s[g]}.$$

Note that the above frequency matching and Q matching conditions are together known as "impedance matching" in electrical engineering.

[00110] Under the above conditions, the maximized efficiency is a monotonically increasing function of only the strong-coupling factor, $U = \kappa / \sqrt{\Gamma_s \Gamma_d} = k\sqrt{Q_s Q_d}$, between the source and device resonators and is given by, $\eta = U^2 / \left(1+\sqrt{1+U^2}\right)^2$, as shown in Fig. 5. Note that the coupling efficiency, η, is greater than 1% when U is greater than 0.2, is greater than 10% when U is greater than 0.7, is greater than 17% when U is greater than 1, is greater than 52% when U is greater than 3, is greater than 80% when U is greater than 9, is greater than 90% when U is greater than 19, and is greater than 95% when U is greater than 45. In some applications, the regime of operation where $U>1$ may be referred to as the "strong-coupling" regime.

[00111] Since a large $U = \kappa / \sqrt{\Gamma_s \Gamma_d} = \left(2\kappa / \sqrt{\omega_s \omega_d}\right)\sqrt{Q_s Q_d}$ is desired in certain circumstances, resonators may be used that are high-Q. The Q of each resonator may be high. The geometric mean of the resonator Q's, $\sqrt{Q_s Q_d}$ may also or instead be high.

[00112] The coupling factor, k, is a number between $0 \le k \le 1$, and it may be independent (or nearly independent) of the resonant frequencies of the source and device resonators, rather it may determined mostly by their relative geometry and the physical decay-law of the field mediating their coupling. In contrast, the coupling coefficient, $\kappa = k\sqrt{\omega_s \omega_d}/2$, may be a strong function of the resonant frequencies. The resonant frequencies of the resonators may be chosen preferably to achieve a high Q rather than to achieve a low Γ, as these two goals may be achievable at two separate resonant frequency regimes.

[00113] A high-Q resonator may be defined as one with Q>100. Two coupled resonators may be referred to as a system of high-Q resonators when each resonator has a Q greater than 100, Q_s>100 and Q_d>100. In other implementationss, two coupled resonators may be referred to as a system of high-Q resonators when the geometric mean of the resonator Q's is greater than 100, $\sqrt{Q_s Q_d}$ >100.

[00114] The resonators may be named or numbered. They may be referred to as source resonators, device resonators, first resonators, second resonators, repeater resonators, and the like. It is to be understood that while two resonators are shown in Fig. 1, and in many of the examples below, other implementations may include three (3) or more resonators. For example, a single source resonator 102S may transfer energy to multiple device resonators 102D or multiple devices. Energy may be transferred from a first device to a second, and then from the second device to the third, and so forth. Multiple sources may transfer energy to a single device or to multiple devices connected to a single device resonator or to multiple devices connected to multiple device resonators. Resonators 102 may serve alternately or simultaneously as sources, devices, or they may be used to relay power from a source in one location to a device in another location. Intermediate electromagnetic resonators 102 may be used to extend the distance range of wireless energy transfer systems. Multiple resonators 102 may be daisy chained together, exchanging energy over extended distances and with a wide range of sources and devices. High power levels may be split between multiple sources 102S, transferred to multiple devices and recombined at a distant location.

[00115] The analysis of a single source and a single device resonator may be extended to multiple source resonators and/or multiple device resonators and/or multiple intermediate resonators. In such an analysis, the conclusion may be that large strong-coupling factors, U_{mn}, between at least some or all of the multiple resonators is preferred for a high system efficiency in the wireless energy transfer. Again, implementations may use source, device and intermediate resonators that have a high Q. The Q of each resonator may be high. The geometric mean $\sqrt{Q_m Q_n}$ of the Q's for pairs of resonators m and n, for which a large U_{mn} is desired, may also or instead be high.

[00116] Note that since the strong-coupling factor of two resonators may be determined by the relative magnitudes of the loss mechanisms of each resonator and the coupling mechanism between the two resonators, the strength of any or all of these mechanisms may be perturbed in the presence of extraneous objects in the vicinity of the resonators as described above.

[00117] Continuing the conventions for labeling from the previous sections, we describe k as the coupling factor in the absence of extraneous objects or materials. We denote the coupling factor in the presence of an extraneous object, p, as $k_{(p)}$, and call it the "perturbed coupling factor" or the "perturbed k". Note that the coupling factor, k, may also be characterized as "unperturbed", when necessary to distinguish from the perturbed coupling factor $k_{(p)}$.

[00118] We define $\delta k_{(p)} \equiv k_{(p)} - k$ and we call it the "perturbation on the coupling factor" or the "perturbation on k" due to an extraneous object, p.

[00119] We also define $\beta_{(p)} \equiv k_{(p)}/k$ and we call it the "coupling factor insensitivity" or the "k-insensitivity". Lower indices, such as $\beta_{12(p)}$, indicate the resonators to which the perturbed and unperturbed coupling factor is referred to, namely $\beta_{12(p)} \equiv k_{12(p)}/k_{12}$.

[00120] Similarly, we describe U as the strong-coupling factor in the absence of extraneous objects. We denote the strong-coupling factor in the presence of an extraneous object, p, as $U_{(p)}$, $U_{(p)} = k_{(p)}\sqrt{Q_{1(p)}Q_{2(p)}}$, and call it the "perturbed strong-coupling factor" or the "perturbed U". Note that the strong-coupling factor U may also be characterized as "unperturbed", when necessary to distinguish from the perturbed strong-coupling factor $U_{(p)}$. Note that the strong-coupling factor U may also be characterized as "unperturbed", when necessary to distinguish from the perturbed strong-coupling factor $U_{(p)}$.

[00121] We define $\delta U_{(p)} \equiv U_{(p)} - U$ and call it the "perturbation on the strong-coupling factor" or the "perturbation on U" due to an extraneous object, p.

[00122] We also define $\Xi_{(p)} \equiv U_{(p)}/U$ and call it the "strong-coupling factor insensitivity" or the "U-insensitivity". Lower indices, such as $\Xi_{12(p)}$, indicate the resonators to which the perturbed and unperturbed coupling factor refers, namely $\Xi_{12(p)} \equiv U_{12(p)}/U_{12}$.

[00123] The efficiency of the energy exchange in a perturbed system may be given by the same formula giving the efficiency of the unperturbed system, where all parameters such as strong-coupling factors, coupling factors, and quality factors are replaced by their perturbed equivalents. For example, in a system of wireless energy transfer including one source and one device resonator, the optimal efficiency may calculated as $\eta_{(p)} = \left[U_{(p)} \Big/ \left(1 + \sqrt{1 + U_{(p)}{}^2} \right) \right]^2$. Therefore, in a system of wireless energy exchange which is perturbed by extraneous objects, large perturbed strong-coupling factors, $U_{mn(p)}$, between at least some or all of the multiple resonators may be desired for a high system efficiency in the wireless energy transfer. Source, device and/or intermediate resonators may have a high $Q_{(p)}$.

[00124] Some extraneous perturbations may sometimes be detrimental for the perturbed strong-coupling factors (via large perturbations on the coupling factors or the quality factors). Therefore, techniques may be used to reduce the effect of extraneous perturbations on the system and preserve large strong-coupling factor insensitivites.

[00125] **<u>Efficiency of Energy Exchange</u>**

[00126] The so-called "useful" energy in a useful energy exchange is the energy or power that must be delivered to a device (or devices) in order to power or charge the device. The transfer efficiency that corresponds to a useful energy exchange may be system or application dependent. For example, high power vehicle charging applications that transfer kilowatts of power may need to be at least 80% efficient in order to supply useful amounts of power resulting in a useful energy exchange sufficient to recharge a vehicle battery, without significantly heating up various components of the transfer system. In some consumer electronics applications, a useful energy exchange may include any energy transfer efficiencies greater than 10%, or any other amount acceptable to keep rechargeable batteries "topped off" and running for long periods of time. For some wireless sensor applications, transfer efficiencies that are much less than 1% may be adequate for powering multiple low power sensors from a single source located a significant distance from the sensors. For still other applications, where wired power transfer is either impossible or impractical, a wide range of transfer efficiencies may be acceptable for a useful energy exchange and may be said to supply useful power to devices in those applications. In general, an operating distance is any distance over which a useful energy exchange is or can be maintained according to the principles disclosed herein.

[00127] A useful energy exchange for a wireless energy transfer in a powering or recharging application may be efficient, highly efficient, or efficient enough, as long as the wasted energy levels, heat dissipation, and associated field strengths are within tolerable limits. The tolerable limits may depend on the application, the environment and the system location. Wireless energy transfer for powering or recharging applications may be efficient, highly efficient, or efficient enough, as long as the desired system performance may be attained for the reasonable cost restrictions, weight restrictions, size restrictions, and the like. Efficient energy transfer may be determined relative to that which could be achieved using traditional inductive techniques that are not high-Q systems. Then, the energy transfer may be defined as being efficient, highly efficient, or efficient enough, if more energy is delivered than could be delivered by similarly sized coil structures in traditional inductive schemes over similar distances or alignment offsets.

[00128] Note that, even though certain frequency and Q matching conditions may optimize the system efficiency of energy transfer, these conditions may not need to be exactly met in order to have efficient enough energy transfer for a useful energy exchange. Efficient energy exchange may be realized so long as the relative offset of the resonant frequencies ($|\omega_m - \omega_n| / \sqrt{\omega_m \omega_n}$) is less than approximately the maximum among $1/Q_{m(p)}$, $1/Q_{n(p)}$ and $k_{mn(p)}$. The Q matching condition may be less critical than the frequency matching condition for efficient energy exchange. The degree by which the strong-loading factors, $U_{m[l]}$, of the resonators due to generators and/or loads may be away from their optimal values and still have efficient enough energy exchange depends on the particular system, whether all or some of the generators and/or loads are Q-mismatched and so on.

[00129] Therefore, the resonant frequencies of the resonators may not be exactly matched, but may be matched within the above tolerances. The strong-loading factors of at least some of the resonators due to generators and/or loads may not be exactly matched to their optimal value. The voltage levels, current levels, impedance values, material parameters, and the like may not be at the exact values described in the disclosure but will be within some acceptable tolerance of those values. The system optimization may include cost, size, weight, complexity, and the like, considerations, in addition to efficiency, Q, frequency, strong coupling factor, and the like, considerations. Some system performance parameters, specifications, and designs may be far from optimal in order to optimize other system performance parameters, specifications and designs.

[00130] In some applications, at least some of the system parameters may be varying in time, for example because components, such as sources or devices, may be mobile or aging or because the loads may be variable or because the perturbations or the environmental conditions are changing etc. In these cases, in order to achieve acceptable matching conditions, at least some of the system parameters may need to be dynamically adjustable or tunable. All the system parameters may be dynamically adjustable or tunable to achieve approximately the optimal operating conditions. However, based on the discussion above, efficient enough energy exchange may be realized even if some system parameters are not variable. In some examples, at least some of the devices may not be dynamically adjusted. In some examples, at least some of the sources may not be dynamically adjusted. In some examples, at least some of the intermediate resonators may not be dynamically adjusted. In some examples, none of the system parameters may be dynamically adjusted.

[00131] **Electromagnetic Resonators**

[00132] The resonators used to exchange energy may be electromagnetic resonators. In such resonators, the intrinsic energy decay rates, Γ_m, are given by the absorption (or resistive) losses and the radiation losses of the resonator.

[00133] The resonator may be constructed such that the energy stored by the electric field is primarily confined within the structure and that the energy stored by the magnetic field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant magnetic near-field. These types of resonators may be referred to as magnetic resonators.

[00134] The resonator may be constructed such that the energy stored by the magnetic field is primarily confined within the structure and that the energy stored by the electric field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant electric near-field. These types of resonators may be referred to as electric resonators.

[00135] Note that the total electric and magnetic energies stored by the resonator have to be equal, but their localizations may be quite different. In some cases, the ratio of the average electric field energy to the average magnetic field energy specified at a distance from a resonator may be used to characterize or describe the resonator.

[00136] Electromagnetic resonators may include an inductive element, a distributed inductance, or a combination of inductances with inductance, L, and a capacitive element, a distributed capacitance, or a combination of capacitances, with capacitance, C. A minimal circuit model of an electromagnetic resonator 102 is shown in Fig. 6a. The resonator may include an inductive element 108 and a capacitive element 104. Provided with initial energy, such as electric field energy stored in the capacitor 104, the system will oscillate as the capacitor discharges transferring energy into magnetic field energy stored in the inductor 108 which in turn transfers energy back into electric field energy stored in the capacitor 104.

[00137] The resonators 102 shown in Figs. 6(b)(c)(d) may be referred to as magnetic resonators. Magnetic resonators may be preferred for wireless energy transfer applications in populated environments because most everyday materials including animals, plants, and humans are non-magnetic (i.e., $\mu_r \approx 1$), so their interaction with magnetic fields is minimal and due primarily to eddy currents induced by the time-variation of the magnetic fields, which is a second-order effect. This characteristic is important both for safety reasons and because it reduces the potential for interactions with extraneous environmental objects and materials that could alter system performance.

[00138] Fig. 6d shows a simplified drawing of some of the electric and magnetic field lines associated with an exemplary magnetic resonator 102B. The magnetic resonator 102B may include a loop of conductor acting as an inductive element 108 and a capacitive element 104 at the ends of the conductor loop. Note that this drawing depicts most of the energy in the region surrounding the resonator being stored in the magnetic field, and most of the energy in the resonator (between the capacitor plates) stored in the electric field. Some electric field, owing to fringing fields, free charges, and the time varying magnetic field, may be stored in the region around the resonator, but the magnetic resonator may be designed to confine the electric fields to be close to or within the resonator itself, as much as possible.

[00139] The inductor 108 and capacitor 104 of an electromagnetic resonator 102 may be bulk circuit elements, or the inductance and capacitance may be distributed and may result from the way the conductors are formed, shaped, or positioned, in the structure. For example, the inductor 108 may be realized by shaping a conductor to enclose a surface area, as shown in Figs. 6(b)(c)(d). This type of resonator 102 may be referred to as a capacitively-loaded loop inductor. Note that we may use the terms “loop” or “coil” to indicate generally a conducting structure (wire, tube, strip, etc.), enclosing a surface of any shape and dimension, with any number of turns. In Fig. 6b, the enclosed surface area is circular, but the surface may be any of a wide variety of other shapes and sizes and may be designed to achieve certain system performance specifications. As an example to indicate how inductance scales with physical dimensions, the inductance for a length of circular conductor arranged to form a circular single-turn loop is approximately,

$$L = \mu_0 x(\ln \frac{8x}{a} - 2),$$

where μ_0 is the magnetic permeability of free space, x, is the radius of the enclosed circular surface area and, a, is the radius of the conductor used to form the inductor loop. A more precise value of the inductance of the loop may be calculated analytically or numerically.

[00140] The inductance for other cross-section conductors, arranged to form other enclosed surface shapes, areas, sizes, and the like, and of any number of wire turns, may be calculated analytically, numerically or it may be determined by measurement. The inductance may be realized using inductor elements, distributed inductance, networks, arrays, series and parallel combinations of inductors and inductances, and the like. The inductance may be fixed or variable and may be used to vary impedance matching as well as resonant frequency operating conditions.

[00141] There are a variety of ways to realize the capacitance required to achieve the desired resonant frequency for a resonator structure. Capacitor plates 110 may be formed and utilized as shown in Fig. 6b, or the capacitance may be distributed and be realized between adjacent windings of a multi-loop conductor 114, as shown in Fig. 6c. The capacitance may be realized using capacitor elements, distributed capacitance, networks, arrays, series and parallel combinations of capacitances, and the like. The capacitance may be fixed or variable and may be used to vary impedance matching as well as resonant frequency operating conditions.

[00142] It is to be understood that the inductance and capacitance in an electromagnetic resonator 102 may be lumped, distributed, or a combination of lumped and distributed inductance and capacitance and that electromagnetic resonators may be realized by combinations of the various elements, techniques and effects described herein.

[00143] Electromagnetic resonators 102 may be include inductors, inductances, capacitors, capacitances, as well as additional circuit elements such as resistors, diodes, switches, amplifiers, diodes, transistors, transformers, conductors, connectors and the like.

[00144] **<u>Resonant Frequency of an Electromagnetic Resonator</u>**

[00145] An electromagnetic resonator 102 may have a characteristic, natural, or resonant frequency determined by its physical properties. This resonant frequency is the frequency at which the energy stored by the resonator oscillates between that stored by the electric field, W_E, ($W_E = q^2/2C$, where q is the charge on the capacitor, C) and that stored by the magnetic field, W_B, ($W_B=Li^2/2$, where *i* is the current through the inductor, *L*) of the resonator. In the absence of any losses in the system, energy would continually be exchanged between the electric field in the capacitor 104 and the magnetic field in the inductor 108. The frequency at which this energy is exchanged may be called the characteristic frequency, the natural frequency, or the resonant frequency of the resonator, and is given by ω ,

$$\omega = 2\pi f = \sqrt{\frac{1}{LC}}.$$

[00146] The resonant frequency of the resonator may be changed by tuning the inductance, L, and/or the capacitance, C, of the resonator. The resonator frequency may be design to operate at the so-called ISM (Industrial, Scientific and Medical) frequencies as specified by the FCC. The resonator frequency may be chosen to meet certain field limit specifications, specific absorption rate (SAR) limit specifications, electromagnetic compatibility (EMC) specifications, electromagnetic interference (EMI) specifications, component size, cost or performance specifications, and the like.

[00147] **Quality Factor of an Electromagnetic Resonator**

[00148] The energy in the resonators 102 shown in Fig. 6 may decay or be lost by intrinsic losses including absorptive losses (also called ohmic or resistive losses) and/or radiative losses. The Quality Factor, or *Q*, of the resonator, which characterizes the energy decay, is inversely proportional to these losses. Absorptive losses may be caused by the finite conductivity of the conductor used to form the inductor as well as by losses in other elements, components, connectors, and the like, in the resonator. An inductor formed from low loss materials may be referred to as a "high-*Q* inductive element" and elements, components, connectors and the like with low losses may be referred to as having "high resistive *Q*'s". In general, the total absorptive loss for a resonator may be calculated as the appropriate series and/or parallel combination of resistive losses for the various elements and components that make up the resonator. That is, in the absence of any significant radiative or component/connection losses, the *Q* of the resonator may be given by, Q_{abs},

$$Q_{abs} = \frac{\omega L}{R_{abs}},$$

where ω, is the resonant frequency, *L,* is the total inductance of the resonator and the resistance for the conductor used to form the inductor, for example, may be given by $R_{abs} = l\rho/A$, (*l* is the length of the wire, ρ is the resistivity of the conductor material, and A is the cross-sectional area over which current flows in the wire). For alternating currents, the cross-sectional area over which current flows may be less than the physical cross-sectional area of the conductor owing to the skin effect. Therefore, high-Q magnetic resonators may be composed of conductors with high conductivity, relatively large surface areas and/or with specially designed profiles (e.g. Litz wire) to minimize proximity effects and reduce the AC resistance.

[00149] The magnetic resonator structures may include high-Q inductive elements composed of high conductivity wire, coated wire, Litz wire, ribbon, strapping or plates, tubing, paint, gels, traces, and the like. The magnetic resonators may be self-resonant, or they may include external coupled elements such as capacitors, inductors, switches, diodes, transistors, transformers, and the like. The magnetic resonators may include distributed and lumped capacitance and inductance. In general, the Q of the resonators will be determined by the Q's of all the individual components of the resonator.

[00150] Because Q is proportional to inductance, L, resonators may be designed to increase L, within certain other constraints. One way to increase L, for example, is to use more than one turn of the conductor to form the inductor in the resonator. Design techniques and trade-offs may depend on the application, and a wide variety of structures, conductors, components, and resonant frequencies may be chosen in the design of high-Q magnetic resonators.

[00151] In the absence of significant absorption losses, the Q of the resonator may be determined primarily by the radiation losses, and given by, $Q_{rad} = \omega L / R_{rad}$, where R_{rad} is the radiative loss of the resonator and may depend on the size of the resonator relative to the frequency, ω, or wavelength, λ, of operation. For the magnetic resonators discussed above, radiative losses may scale as $R_{rad} \sim (x/\lambda)^4$ (characteristic of magnetic dipole radiation), where x is a characteristic dimension of the resonator, such as the radius of the inductive element shown in Fig. 6b, and where $\lambda = c / f$, where c is the speed of light and f is as defined above. The size of the magnetic resonator may be much less than the wavelength of operation so radiation losses may be very small. Such structures may be referred to as sub-wavelength resonators. Radiation may be a loss mechanism for non-radiative wireless energy transfer systems and designs may be chosen to reduce or minimize R_{rad}. Note that a high-Q_{rad} may be desirable for non-radiative wireless energy transfer schemes.

[00152] Note too that the design of resonators for non-radiative wireless energy transfer differs from antennas designed for communication or far-field energy transmission purposes. Specifically, capacitively-loaded conductive loops may be used as resonant antennas (for example in cell phones), but those operate in the far-field regime where the radiation *Q*'s are intentionally designed to be small to make the antenna efficient at radiating energy. Such designs are not appropriate for the efficient near-field wireless energy transfer technique disclosed in this application.

[00153] The quality factor of a resonator including both radiative and absorption losses is $Q=\omega L/(R_{abs}+R_{rad})$. Note that there may be a maximum *Q* value for a particular resonator and that resonators may be designed with special consideration given to the size of the resonator, the materials and elements used to construct the resonator, the operating frequency, the connection mechanisms, and the like, in order to achieve a high-Q resonator. Fig. 7 shows a plot of *Q* of an exemplary magnetic resonator (in this case a coil with a diameter of 60 cm made of copper pipe with an outside diameter (OD) of 4 cm) that may be used for wireless power transmission at MHz frequencies. The absorptive *Q* (dashed line) 702 increases with frequency, while the radiative *Q* (dotted line) 704 decreases with frequency, thus leading the overall *Q* to peak 708 at a particular frequency. Note that the *Q* of this exemplary resonator is greater than 100 over a wide frequency range. Magnetic resonators may be designed to have high-Q over a range of frequencies and system operating frequency may set to any frequency in that range.

[00154] When the resonator is being described in terms of loss rates, the *Q* may be defined using the intrinsic decay rate, 2Γ, as described previously. The intrinsic decay rate is the rate at which an uncoupled and undriven resonator loses energy. For the magnetic resonators described above, the intrinsic loss rate may be given by $\Gamma=(R_{abs}+R_{rad})/2L$, and the quality factor, *Q*, of the resonator is given by $Q=\omega/2\Gamma$.

[00155] Note that a quality factor related only to a specific loss mechanism may be denoted as $Q_{mechanism}$, if the resonator is not specified, or as $Q_{1,mechanism}$, if the resonator is specified (e.g. resonator 1). For example, $Q_{1,rad}$ is the quality factor for resonator 1 related to its radiation losses.

[00156] **<u>Electromagnetic Resonator Near-Fields</u>**

[00157] The high-Q electromagnetic resonators used in the near-field wireless energy transfer system disclosed here may be sub-wavelength objects. That is, the physical dimensions of the resonator may be much smaller than the wavelength corresponding to the resonant frequency. Sub-wavelength magnetic resonators may have most of the energy in the region surrounding the resonator stored in their magnetic near-fields, and these fields may also be described as stationary or non-propagating because they do not radiate away from the resonator. The extent of the near-field in the area surrounding the resonator is typically set by the wavelength, so it may extend well beyond the resonator itself for a sub-wavelength resonator. The limiting surface, where the field behavior changes from near-field behavior to far-field behavior may be called the "radiation caustic".

[00158] The strength of the near-field is reduced the farther one gets away from the resonator. While the field strength of the resonator near-fields decays away from the resonator, the fields may still interact with objects brought into the general vicinity of the resonator. The degree to which the fields interact depends on a variety of factors, some of which may be controlled and designed, and some of which may not. The wireless energy transfer schemes described herein may be realized when the distance between coupled resonators is such that one resonator lies within the radiation caustic of the other.

[00159] The near-field profiles of the electromagnetic resonators may be similar to those commonly associated with dipole resonators or oscillators. Such field profiles may be described as omni-directional, meaning the magnitudes of the fields are non-zero in all directions away from the object.

[00160] <u>Characteristic Size of An Electromagnetic Resonator</u>

[00161] Spatially separated and/or offset magnetic resonators of sufficient Q may achieve efficient wireless energy transfer over distances that are much larger than have been seen in the prior art, even if the sizes and shapes of the resonator structures are different. Such resonators may also be operated to achieve more efficient energy transfer than was achievable with previous techniques over shorter range distances. We describe such resonators as being capable of mid-range energy transfer.

[00162] Mid-range distances may be defined as distances that are larger than the characteristic dimension of the smallest of the resonators involved in the transfer, where the distance is measured from the center of one resonator structure to the center of a spatially separated second resonator structure. In this definition, two-dimensional resonators are spatially separated when the areas circumscribed by their inductive elements do not intersect and three-dimensional resonators are spatially separated when their volumes do not intersect. A two-dimensional resonator is spatially separated from a three-dimensional resonator when the area circumscribed by the former is outside the volume of the latter.

[00163] Fig. 8 shows some example resonators with their characteristic dimensions labeled. It is to be understood that the characteristic sizes 802 of resonators 102 may be defined in terms of the size of the conductor and the area circumscribed or enclosed by the inductive element in a magnetic resonator and the length of the conductor forming the capacitive element of an electric resonator. Then, the characteristic size 802 of a resonator 102, x_{char}, may be equal to the radius of the smallest sphere that can fit around the inductive or capacitive element of the magnetic or electric resonator respectively, and the center of the resonator structure is the center of the sphere. The characteristic thickness 804, t_{char}, of a resonator 102 may be the smallest possible height of the highest point of the inductive or capacitive element in the magnetic or capacitive resonator respectively, measured from a flat surface on which it is placed. The characteristic width 808 of a resonator 102, w_{char}, may be the radius of the smallest possible circle through which the inductive or capacitive element of the magnetic or electric resonator respectively, may pass while traveling in a straight line. For example, the characteristic width 808 of a cylindrical resonator may be the radius of the cylinder.

[00164] In this inventive wireless energy transfer technique, energy may be exchanged efficiently over a wide range of distances, but the technique is distinguished by the ability to exchange useful energy for powering or recharging devices over mid-range distances and between resonators with different physical dimensions, components and orientations. Note that while k may be small in these circumstances, strong coupling and efficient energy transfer may be realized by using high-Q resonators to achieve a high U, $U = k\sqrt{Q_s Q_d}$. That is, increases in Q may be used to at least partially overcome decreases in k, to maintain useful energy transfer efficiencies.

[00165] Note too that while the near-field of a single resonator may be described as omni-directional, the efficiency of the energy exchange between two resonators may depend on the relative position and orientation of the resonators. That is, the efficiency of the energy exchange may be maximized for particular relative orientations of the resonators. The sensitivity of the transfer efficiency to the relative position and orientation of two uncompensated resonators may be captured in the calculation of either k or κ. While coupling may be achieved between resonators that are offset and/or rotated relative to each other, the efficiency of the exchange may depend on the details of the positioning and on any feedback, tuning, and compensation techniques implemented during operation.

[00166] <u>High-*Q* Magnetic Resonators</u>

[00167] In the near-field regime of a sub-wavelength capacitively-loaded loop magnetic resonator ($x \ll \lambda$), the resistances associated with a circular conducting loop inductor composed of N turns of wire whose radius is larger than the skin depth, are approximately $R_{abs} = \sqrt{\mu_o \rho \omega / 2} \cdot Nx / a$ and $R_{rad} = \pi / 6 \cdot \eta_o N^2 (\omega x / c)^4$, where ρ is the resistivity of the conductor material and $\eta_o \approx 120\pi\ \Omega$ is the impedance of free space. . The inductance, L, for such a N-turn loop is approximately N^2 times the inductance of a single-turn loop given previously. The quality factor of such a resonator, $Q = \omega L / (R_{abs} + R_{rad})$, is highest for a particular frequency determined by the system parameters (Fig. 4). As described previously, at lower frequencies the Q is determined primarily by absorption losses and at higher frequencies the Q is determined primarily by radiation losses.

[00168] Note that the formulas given above are approximate and intended to illustrate the functional dependence of R_{abs}, R_{rad} and L on the physical parameters of the structure. More accurate numerical calculations of these parameters that take into account deviations from the strict quasi-static limit, for example a non-uniform current/charge distribution along the conductor, may be useful for the precise design of a resonator structure.

[00169] Note that the absorptive losses may be minimized by using low loss conductors to form the inductive elements. The loss of the conductors may be minimized by using large surface area conductors such as conductive tubing, strapping, strips, machined objects, plates, and the like, by using specially designed conductors such as Litz wire, braided wires, wires of any cross-section, and other conductors with low proximity losses, in which case the frequency scaled behavior described above may be different, and by using low resistivity materials such as high-purity copper and silver, for example. One advantage of using conductive tubing as the conductor at higher operating frequencies is that it may be cheaper and lighter than a similar diameter solid conductor, and may have similar resistance because most of the current is traveling along the outer surface of the conductor owing to the skin effect.

[00170] To get a rough estimate of achievable resonator designs made from copper wire or copper tubing and appropriate for operation in the microwave regime, one may calculate the optimum Q and resonant frequency for a resonator composed of one circular inductive element ($N=1$) of copper wire ($\rho=1.69\cdot 10^{-8}\ \Omega m$) with various cross sections. Then for an inductive element with characteristic size $x=1\ cm$ and conductor diameter a=1 mm, appropriate for a cell phone for example, the quality factor peaks at Q=1225 when $f=380$ MHz. For $x=30\ cm$ and $a=2\ mm$, an inductive element size that might be appropriate for a laptop or a household robot, $Q=1103$ at $f=17$ MHz. For a larger source inductive element that might be located in the ceiling for example, x=1 m and a=4 mm, Q may be as high as Q=1315 at $f=5$ MHz. Note that a number of practical examples yield expected quality factors of $Q\approx 1000\text{-}1500$ at $\lambda/x\approx 50\text{-}80$. Measurements of a wider variety of coil shapes, sizes, materials and operating frequencies than described above show that Q's >100 may be realized for a variety of magnetic resonator structures using commonly available materials.

[00171] As described above, the rate for energy transfer between two resonators of characteristic size x_1 and x_2, and separated by a distance D between their centers, may be given by κ . To give an example of how the defined parameters scale, consider the cell phone, laptop, and ceiling resonator examples from above, at three (3) distances; D/x=10, 8, 6. In the examples considered here, the source and device resonators are the same size, $x_1=x_2$, and shape, and are oriented as shown in Fig. 1(b). In the cell phone example, $\omega/2\kappa$ =3033, 1553, 655 respectively. In the laptop example, $\omega/2\kappa$ =7131, 3651, 1540 respectively and for the ceiling resonator example, $\omega/2\kappa$ =6481, 3318, 1400. The corresponding coupling-to-loss ratios peak at the frequency where the inductive element Q peaks and are κ/Γ =0.4, 0.79, 1.97 and 0.15, 0.3, 0.72 and 0.2, 0.4, 0.94 for the three inductive element sizes and distances described above. An example using different sized inductive elements is that of an x_1=1 m inductor (e.g. source in the ceiling) and an x_2=30 cm inductor (e.g. household robot on the floor) at a distance D=3 m apart (e.g. room height). In this example, the strong-coupling figure of merit, $U=\kappa/\sqrt{\Gamma_1\Gamma_2}$ =0.88, for an efficiency of approximately 14%, at the optimal operating frequency of f=6.4 MHz. Here, the optimal system operating frequency lies between the peaks of the individual resonator Q's.

[00172] Inductive elements may be formed for use in high-Q magnetic resonators. We have demonstrated a variety of high-*Q* magnetic resonators based on copper conductors that are formed into inductive elements that enclose a surface. Inductive elements may be formed using a variety of conductors arranged in a variety of shapes, enclosing any size or shaped area, and they may be single turn or multiple turn elements. Drawings of exemplary inductive elements 900A-B are shown in Fig. 9. The inductive elements may be formed to enclose a circle, a rectangle, a square, a triangle, a shape with rounded corners, a shape that follows the contour of a particular structure or device, a shape that follows, fills, or utilizes, a dedicated space within a structure or device, and the like. The designs may be optimized for size, cost, weight, appearance, performance, and the like.

[00173] These conductors may be bent or formed into the desired size, shape, and number of turns. However, it may be difficult to accurately reproduce conductor shapes and sizes using manual techniques. In addition, it may be difficult to maintain uniform or desired center-to-center spacings between the conductor segments in adjacent turns of the inductive elements. Accurate or uniform spacing may be important in determining the self capacitance of the structure as well as any proximity effect induced increases in AC resistance, for example.

[00174] Molds may be used to replicate inductor elements for high-*Q* resonator designs. In addition, molds may be used to accurately shape conductors into any kind of shape without creating kinks, buckles or other potentially deleterious effects in the conductor. Molds may be used to form the inductor elements and then the inductor elements may be removed from the forms. Once removed, these inductive elements may be built into enclosures or devices that may house the high-Q magnetic resonator. The formed elements may also or instead remain in the mold used to form them.

[00175] The molds may be formed using standard CNC (computer numerical control) routing or milling tools or any other known techniques for cutting or forming grooves in blocks. The molds may also or instead be formed using machining techniques, injection molding techniques, casting techniques, pouring techniques, vacuum techniques, thermoforming techniques, cut-in-place techniques, compression forming techniques and the like.

[00176] The formed element may be removed from the mold or it may remain in the mold. The mold may be altered with the inductive element inside. The mold may be covered, machined, attached, painted and the like. The mold and conductor combination may be integrated into another housing, structure or device. The grooves cut into the molds may be any dimension and may be designed to form conducting tubing, wire, strapping, strips, blocks, and the like into the desired inductor shapes and sizes.

[00177] The inductive elements used in magnetic resonators may contain more than one loop and may spiral inward or outward or up or down or in some combination of directions. In general, the magnetic resonators may have a variety of shapes, sizes and number of turns and they may be composed of a variety of conducing materials.

[00178] The magnetic resonators may be free standing or they may be enclosed in an enclosure, container, sleeve or housing. The magnetic resonators may include the form used to make the inductive element. These various forms and enclosures may be composed of almost any kind of material. Low loss materials such as Teflon, REXOLITE, styrene, and the like may be preferable for some applications. These enclosures may contain fixtures that hold the inductive elements.

[00179] Magnetic resonators may be composed of self-resonant coils of copper wire or copper tubing. Magnetic resonators composed of self resonant conductive wire coils may include a wire of length *l,* and cross section radius *a,* wound into a helical coil of radius *x*, height *h,* and number of turns *N*, which may for example be characterized as $N = \sqrt{l^2 - h^2} / 2\pi x$.

[00180] A magnetic resonator structure may be configured so that *x* is about 30 cm, *h* is about 20 cm, *a* is about 3 mm and *N* is about 5.25, and, during operation, a power source coupled to the magnetic resonator may drive the resonator at a resonant frequency, *f*, where *f* is about 10.6 MHz. Where *x* is about 30 cm, *h* is about 20 cm, *a* is about 1 cm and *N* is about 4, the resonator may be driven at a frequency, *f*, where *f* is about 13.4 MHz. Where *x* is about 10 cm, *h* is about 3 cm, *a* is about 2 mm and *N* is about 6, the resonator may be driven at a frequency, *f*, where *f* is about 21.4 MHz.

[00181] High-Q inductive elements may be designed using printed circuit board traces. Printed circuit board traces may have a variety of advantages compared to mechanically formed inductive elements including that they may be accurately reproduced and easily integrated using established printed circuit board fabrication techniques, that their AC resistance may be lowered using custom designed conductor traces, and that the cost of mass-producing them may be significantly reduced.

[00182] High-Q inductive elements may be fabricated using standard PCB techniques on any PCB material such as FR-4 (epoxy E-glass), multi-functional epoxy, high performance epoxy, bismalaimide triazine/epoxy, polyimide, Cyanate Ester, polytetraflouroethylene (Teflon), FR-2, FR-3, CEM-1, CEM-2, Rogers, Resolute, and the like. The conductor traces may be formed on printed circuit board materials with lower loss tangents.

[00183] The conducting traces may be composed of copper, silver, gold, aluminum, nickel and the like, and they may be composed of paints, inks, or other cured materials. The circuit board may be flexible and it may be a flex-circuit. The conducting traces may be formed by chemical deposition, etching, lithography, spray deposition, cutting, and the like. The conducting traces may be applied to form the desired patterns and they may be formed using crystal and structure growth techniques.

[00184] The dimensions of the conducting traces, as well as the number of layers containing conducting traces, the position, size and shape of those traces and the architecture for interconnecting them may be designed to achieve or optimize certain system specifications such as resonator Q, $Q_{(p)}$, resonator size, resonator material and fabrication costs, U, $U_{(p)}$, and the like.

[00185] As an example, a three-turn high-Q inductive element 1001A was fabricated on a four-layer printed circuit board using the rectangular copper trace pattern as shown in Fig. 10(a). The copper trace is shown in black and the PCB in white. The width and thickness of the copper traces in this example was approximately 1 cm (400 mils) and 43 μ m (1.7 mils) respectively. The edge-to-edge spacing between turns of the conducting trace on a single layer was approximately 0.75 cm (300 mils) and each board layer thickness was approximately 100 μ m (4 mils). The pattern shown in Fig. 10(a) was repeated on each layer of the board and the conductors were connected in parallel. The outer dimensions of the 3-loop structure were approximately 30 cm by 20 cm. The measured inductance of this PCB loop was 5.3 μ H. A magnetic resonator using this inductor element and tunable capacitors had a quality factor, Q, of 550 at its designed resonance frequency of 6.78 MHz. The resonant frequency could be tuned by changing the inductance and capacitance values in the magnetic resonator.

[00186] As another example, a two-turn inductor 1001B was fabricated on a four-layer printed circuit board using the rectangular copper trace pattern shown in Fig. 10(b). The copper trace is shown in black and the PCB in white. The width and height of the copper traces in this example were approximately 0.75 cm (300 mils) and 43 μ m (1.7 mils) respectively. The edge-to-edge spacing between turns of the conducting trace on a single layer was approximately 0.635 cm (250 mils) and each board layer thickness was approximately 100 μ m (4 mils). The pattern shown in Fig. 10(b) was repeated on each layer of the board and the conductors were connected in parallel. The outer dimensions of the two-loop structure were approximately 7.62 cm by 26.7 cm. The measured inductance of this PCB loop was 1.3 μ H. Stacking two boards together with a vertical separation of approximately 0.635 cm (250 mils) and connecting the two boards in series produced a PCB inductor with an inductance of approximately 3.4 μ H. A magnetic resonator using this stacked inductor loop and tunable capacitors had a quality factor, Q, of 390 at its designed resonance frequency of 6.78 MHz. The resonant frequency could be tuned by changing the inductance and capacitance values in the magnetic resonator.

[00187] The inductive elements may be formed using magnetic materials of any size, shape thickness, and the like, and of materials with a wide range of permeability and loss values. These magnetic materials may be solid blocks, they may enclose hollow volumes, they may be formed from many smaller pieces of magnetic material tiled and or stacked together, and they may be integrated with conducting sheets or enclosures made from highly conducting materials. Wires may be wrapped around the magnetic materials to generate the magnetic near-field. These wires may be wrapped around one or more than one axis of the structure. Multiple wires may be wrapped around the magnetic materials and combined in parallel, or in series, or via a switch to form customized near-field patterns.

[00188] The magnetic resonator may include 15 turns of Litz wire wound around a 19.2 cm x 10 cm x 5 mm tiled block of 3F3 ferrite material. The Litz wire may be wound around the ferrite material in any direction or combination of directions to achieve the desire resonator performance. The number of turns of wire, the spacing between the turns, the type of wire, the size and shape of the magnetic materials and the type of magnetic material are all design parameters that may be varied or optimized for different application scenarios.

[00189] **<u>High-*Q* Magnetic resonators using magnetic material structures</u>**

[00190] It may be possible to use magnetic materials assembled to form an open magnetic circuit, albeit one with an air gap on the order of the size of the whole structure, to realize a magnetic resonator structure. In these structures, high conductivity materials are wound around a structure made from magnetic material to form the inductive element of the magnetic resonator. Capacitive elements may be connected to the high conductivity materials, with the resonant frequency then determined as described above. These magnetic resonators have their dipole moment in the plane of the two dimensional resonator structures, rather than perpendicular to it, as is the case for the capacitively-loaded inductor loop resonators.

[00191] A diagram of a single planar resonator structure is shown in Fig. 11(a). The planar resonator structure is constructed of a core of magnetic material 1121, such as ferrite with a loop or loops of conducting material 1122 wrapped around the core 1121. The structure may be used as the source resonator that transfers power and the device resonator that captures energy. When used as a source, the ends of the conductor may be coupled to a power source. Alternating electrical current flowing through the conductor loops excites alternating magnetic fields. When the structure is being used to receive power, the ends of the conductor may be coupled to a power drain or load. Changing magnetic fields induce an electromotive force in the loop or loops of the conductor wound around the core magnetic material. The dipole moment of these types of structures is in the plane of the structures and is, for example, directed along the Y axis for the structure in Figure 11(a). Two such structures have strong coupling when placed substantially in the same plane (i.e. the X,Y plane of Figure 11). The structures of Figure 11(a) have the most favorable orientation when the resonators are aligned in the same plane along their Y axis.

[00192] The geometry and the coupling orientations of the described planar resonators may be preferable for some applications. The planar or flat resonator shape may be easier to integrate into many electronic devices that are relatively flat and planar. The planar resonators may be integrated into the whole back or side of a device without requiring a change in geometry of the device. Due to the flat shape of many devices, the natural position of the devices when placed on a surface is to lay with their largest dimension being parallel to the surface they are placed on. A planar resonator integrated into a flat device is naturally parallel to the plane of the surface and is in a favorable coupling orientation relative to the resonators of other devices or planar resonator sources placed on a flat surface.

[00193] As mentioned, the geometry of the planar resonators may allow easier integration into devices. Their low profile may allow a resonator to be integrated into or as part of a complete side of a device. When a whole side of a device is covered by the resonator, magnetic flux can flow through the resonator core without being obstructed by lossy material that may be part of the device or device circuitry.

[00194] The core of the planar resonator structure may be of a variety of shapes and thicknesses and may be flat or planar such that the minimum dimension does not exceed 30% of the largest dimension of the structure. The core may have complex geometries and may have indentations, notches, ridges, and the like. Geometric enhancements may be used to reduce the coupling dependence on orientation and they may be used to facilitate integration into devices, packaging, packages, enclosures, covers, skins, and the like. Two exemplary variations of core geometries are shown in Figure 11(b). For example, the planar core 1131 may be shaped such that the ends are substantially wider than the middle of the structure to create an indentation for the conductor winding. The core material may be of varying thickness with ends that are thicker and wider than the middle. The core material 1132 may have any number of notches or cutouts 1133 of various depths, width, and shapes to accommodate conductor loops, housing, packaging, and the like.

[00195] The shape and dimensions of the core may be further dictated by the dimensions and characteristics of the device that they are integrated into. The core material may curve to follow the contours of the device, or may require non-symmetric notches or cutouts to allow clearance for parts of the device. The core structure may be a single monolithic piece of magnetic material or may be composed of a plurality of tiles, blocks, or pieces that are arranged together to form the larger structure. The different layers, tiles, blocks, or pieces of the structure may be of similar or may be of different materials. It may be desirable to use materials with different magnetic permeability in different locations of the structure. Core structures with different magnetic permeability may be useful for guiding the magnetic flux, improving coupling, and affecting the shape or extent of the active area of a system.

[00196] The conductor of the planar resonator structure may be wound at least once around the core. In certain circumstances, it may be preferred to wind at least three loops. The conductor can be any good conductor including conducting wire, Litz wire, conducting tubing, sheets, strips, gels, inks, traces and the like.

[00197] The size, shape, or dimensions of the active area of source may be further enhanced, altered, or modified with the use of materials that block, shield, or guide magnetic fields. To create non-symmetric active area around a source once side of the source may be covered with a magnetic shield to reduce the strength of the magnetic fields in a specific direction. The shield may be a conductor or a layered combination of conductor and magnetic material which can be used to guide magnetic fields away from a specific direction. Structures composed of layers of conductors and magnetic materials may be used to reduce energy losses that may occur due to shielding of the source.

[00198] The plurality of planar resonators may be integrated or combined into one planar resonator structure. A conductor or conductors may be wound around a core structure such that the loops formed by the two conductors are not coaxial. An example of such a structure is shown in Figure 12 where two conductors 1201,1202 are wrapped around a planar rectangular core 1203 at orthogonal angles. The core may be rectangular or it may have various geometries with several extensions or protrusions. The protrusions may be useful for wrapping of a conductor, reducing the weight, size, or mass of the core, or may be used to enhance the directionality or omni-directionality of the resonator. A multi wrapped planar resonator with four protrusions is shown by the inner structure 1310 in Figure 13, where four conductors 1301, 1302, 1303, 1304 are wrapped around the core. The core may have extensions 1305,1306,1307,1308 with one or more conductor loops. A single conductor may be wrapped around a core to form loops that are not coaxial. The four conductor loops of Figure 13, for example, may be formed with one continuous piece of conductor, or using two conductors where a single conductor is used to make all coaxial loops.

[00199] Non-uniform or asymmetric field profiles around the resonator comprising a plurality of conductor loops may be generated by driving some conductor loops with non-identical parameters. Some conductor loops of a source resonator with a plurality of conductor loops may be driven by a power source with a different frequency, voltage, power level, duty cycle, and the like all of which may be used to affect the strength of the magnetic field generated by each conductor.

[00200] The planar resonator structures may be combined with a capacitively-loaded inductor resonator coil to provide an omni-directional active area all around, including above and below the source while maintaining a flat resonator structure. As shown in Figure 13, an additional resonator loop coil 1309 comprising of a loop or loops of a conductor, may be placed in a common plane as the planar resonator structure 1310. The outer resonator coil provides an active area that is substantially above and below the source. The resonator coil can be arranged with any number of planar resonator structures and arrangements described herein.

[00201] The planar resonator structures may be enclosed in magnetically permeable packaging or integrated into other devices. The planar profile of the resonators within a single, common plane allows packaging and integration into flat devices. A diagram illustrating the application of the resonators is shown in Figure 14. A flat source 1411 comprising one or more planar resonators 1414 each with one or more conductor loops may transfer power to devices 1412,1413 that are integrated with other planar resonators 1415,1416 and placed within an active area 1417 of the source. The devices may comprise a plurality of planar resonators such that regardless of the orientation of the device with respect to the source the active area of the source does not change. In addition to invariance to rotational misalignment, a flat device comprising of planar resonators may be turned upside down without substantially affecting the active area since the planar resonator is still in the plane of the source.

[00202] Another diagram illustrating a possible use of a power transfer system using the planar resonator structures is shown in Figure 15. A planar source 1521 placed on top of a surface 1525 may create an active area that covers a substantial surface area creating an "energized surface" area. Devices such as computers 1524, mobile handsets 1522, games, and other electronics 1523 that are coupled to their respective planar device resonators may receive energy from the source when placed within the active area of the source, which may be anywhere on top of the surface. Several devices with different dimensions may be placed in the active area and used normally while charging or being powered from the source without having strict placement or alignment constraints. The source may be placed under the surface of a table, countertop, desk, cabinet, and the like, allowing it to be completely hidden while energizing the top surface of the table, countertop, desk, cabinet and the like, creating an active area on the surface that is much larger than the source.

[00203] The source may include a display or other visual, auditory, or vibration indicators to show the direction of charging devices or what devices are being charged, error or problems with charging, power levels, charging time, and the like.

[00204] The source resonators and circuitry may be integrated into any number of other devices. The source may be integrated into devices such as clocks, keyboards, monitors, picture frames, and the like. For example, a keyboard integrated with the planar resonators and appropriate power and control circuitry may be used as a source for devices placed around the keyboard such as computer mice, webcams, mobile handsets, and the like without occupying any additional desk space.

[00205] While the planar resonator structures have been described in the context of mobile devices it should be clear to those skilled in the art that a flat planar source for wireless power transfer with an active area that extends beyond its physical dimensions has many other consumer and industrial applications. The structures and configuration may be useful for a large number of applications where electronic or electric devices and a power source are typically located, positioned, or manipulated in substantially the same plane and alignment. Some of the possible application scenarios include devices on walls, floor, ceilings or any other substantially planar surfaces.

[00206] Flat source resonators may be integrated into a picture frame or hung on a wall thereby providing an active area within the plane of the wall where other electronic devices such as digital picture frames, televisions, lights, and the like can be mounted and powered without wires. Planar resonators may be integrated into a floor resulting in an energized floor or active area on the floor on which devices can be placed to receive power. Audio speakers, lamps, heaters, and the like can be placed within the active are and receive power wirelessly.

[00207] The planar resonator may have additional components coupled to the conductor. Components such as capacitors, inductors, resistors, diodes, and the like may be coupled to the conductor and may be used to adjust or tune the resonant frequency and the impedance matching for the resonators.

[00208] A planar resonator structure of the type described above and shown in Fig. 11(a), may be created, for example, with a quality factor, Q, of 100 or higher and even Q of 1,000 or higher. Energy may be wirelessly transferred from one planar resonator structure to another over a distance larger than the characteristic size of the resonators, as shown in Fig. 11(c).

[00209] In addition to utilizing magnetic materials to realize a structure with properties similar to the inductive element in the magnetic resonators, it may be possible to use a combination of good conductor materials and magnetic material to realize such inductive structures. Fig. 16(a) shows a magnetic resonator structure 1602 that may include one or more enclosures made of high-conductivity materials (the inside of which would be shielded from AC electromagnetic fields generated outside) surrounded by at least one layer of magnetic material and linked by blocks of magnetic material 1604.

A structure may include a high-conductivity sheet of material covered on one side by a layer of magnetic material. The layered structure may instead be applied conformally to an electronic device, so that parts of the device may be covered by the high-conductivity and magnetic material layers, while other parts that need to be easily accessed (such as buttons or screens) may be left uncovered. The structure may also or instead include only layers or bulk pieces of magnetic material. Thus, a magnetic resonator may be incorporated into an existing device without significantly interfering with its existing functions and with little or no need for extensive redesign. Moreover, the layers of good conductor and/or magnetic material may be made thin enough (of the order of a millimeter or less) that they would add little extra weight and volume to the completed device. An oscillating current applied to a length of conductor wound around the structure, as shown by the square loop in the center of the structure in Figure 16 may be used to excite the electromagnetic fields associated with this structure.

[00210] Quality factor of the structure

[00211] A structure of the type described above may be created with a quality factor, *Q*, of the order of 1,000 or higher. This high-Q is possible even if the losses in the magnetic material are high, if the fraction of magnetic energy within the magnetic material is small compared to the total magnetic energy associated with the object. For structures composed of layers conducting materials and magnetic materials, the losses in the conducting materials may be reduced by the presence of the magnetic materials as described previously. In structures where the magnetic material layer's thickness is of the order of 1/100 of the largest dimension of the system (e.g., the magnetic material may be of the order of 1 mm thick, while the area of the structure is of the order of 10 cm x 10 cm), and the relative permeability is of the order of 1,000, it is possible to make the fraction of magnetic energy contained within the magnetic material only a few hundredths of the total magnetic energy associated with the object or resonator. To see how that comes about, note that the expression for the magnetic energy contained in a volume is $U_m = \int_V d\mathbf{r}\mathbf{B}(\mathbf{r})^2 / (2\mu_r \mu_0)$, so as long as ***B*** (rather than ***H***) is the main field conserved across the magnetic material-air interface (which is typically the case in open magnetic circuits), the fraction of magnetic energy contained in the high-μ_r region may be significantly reduced compared to what it is in air.

[00212] If the fraction of magnetic energy in the magnetic material is denoted by *frac*, and the loss tangent of the material is *tanδ*, then the *Q* of the resonator, assuming the magnetic material is the only source of losses, is $Q=1/(frac \times \tan\delta)$. Thus, even for loss tangents as high as 0.1, it is possible to achieve *Q's* of the order of 1,000 for these types of resonator structures.

[00213] If the structure is driven with *N* turns of wire wound around it, the losses in the excitation inductor loop can be ignored if *N* is sufficiently high. Fig. 17 shows an equivalent circuit 1700 schematic for these structures and the scaling of the loss mechanisms and inductance with the number of turns, *N,* wound around a structure made of conducting and magnetic material. If proximity effects can be neglected (by using an appropriate winding, or a wire designed to minimize proximity effects, such as Litz wire and the like), the resistance 1702 due to the wire in the looped conductor scales linearly with the length of the loop, which is in turn proportional to the number of turns. On the other hand, both the equivalent resistance 1708 and equivalent inductance 1704 of these special structures are proportional to the square of the magnetic field inside the structure. Since this magnetic field is proportional to *N*, the equivalent resistance 1708 and equivalent inductance 1704 are both proportional to N^2. Thus, for large enough N, the resistance 1702 of the wire is much smaller than the equivalent resistance 1708 of the magnetic structure, and the *Q* of the resonator asymptotes to $Q_{max} = \omega L_\mu / R_\mu$.

[00214] Fig. 16 (a) shows a drawing of a copper and magnetic material structure 1602 driven by a square loop of current around the narrowed segment at the center of the structure 1604 and the magnetic field streamlines generated by this structure 1608. This exemplary structure includes two 20 cm x 8 cm x 2 cm hollow regions enclosed with copper and then completely covered with a 2 mm layer of magnetic material having the properties $\mu_r' = 1{,}400$, $\mu_r'' = 5$, and $\sigma = 0.5$ S/m. These two parallelepipeds are spaced 4 cm apart and are connected by a 2 cm x 4 cm x 2 cm block of the same magnetic material. The excitation loop is wound around the center of this block. At a frequency of 300 kHz, this structure has a calculated *Q* of 890. The conductor and magnetic material structure may be shaped to optimize certain system parameters. For example, the size of the structure enclosed by the excitation loop may be small to reduce the resistance of the excitation loop, or it may be large to mitigate losses in the magnetic material associated with large magnetic fields. Note that the magnetic streamlines and *Q's* associated with the same structure composed of magnetic material only would be similar to the layer conductor and magnetic material design shown here.

[00215] **Electromagnetic Resonators Interacting with Other Objects**

[00216] For electromagnetic resonators, extrinsic loss mechanisms that perturb the intrinsic Q may include absorption losses inside the materials of nearby extraneous objects and radiation losses related to scattering of the resonant fields from nearby extraneous objects. Absorption losses may be associated with materials that, over the frequency range of interest, have non-zero, but finite, conductivity, σ , (or equivalently a non-zero and finite imaginary part of the dielectric permittivity), such that electromagnetic fields can penetrate it and induce currents in it, which then dissipate energy through resistive losses. An object may be described as lossy if it at least partly includes lossy materials.

[00217] Consider an object including a homogeneous isotropic material of conductivity, σ and magnetic permeability, μ . The penetration depth of electromagnetic fields inside this object is given by the skin depth, $\delta = \sqrt{2/\omega\mu\sigma}$. The power dissipated inside the object, P_d , can be determined from $P_d = \int_V d\mathbf{r}\sigma |\mathbf{E}|^2 = \int_V d\mathbf{r} |\mathbf{J}|^2 /\sigma$ where we made use of Ohm's law, $\mathbf{J} = \sigma\mathbf{E}$, and where $\mathbf{E}$ is the electric field and $\mathbf{J}$ is the current density.

[00218] If over the frequency range of interest, the conductivity, σ , of the material that composes the object is low enough that the material's skin depth, δ, may be considered long, (i.e. δ is longer than the objects' characteristic size, or δ is longer than the characteristic size of the portion of the object that is lossy) then the electromagnetic fields, $\mathbf{E}$ and $\mathbf{H}$, where $\mathbf{H}$ is the magnetic field, may penetrate significantly into the object. Then, these finite-valued fields may give rise to a dissipated power that scales as $P_d \sim \sigma V_{ol} \left\langle |\mathbf{E}|^2 \right\rangle$, where V_{ol} is the volume of the object that is lossy and $\left\langle |\mathbf{E}|^2 \right\rangle$ is the spatial average of the electric-field squared, in the volume under consideration. Therefore, in the low-conductivity limit, the dissipated power scales proportionally to the conductivity and goes to zero in the limit of a non-conducting (purely dielectric) material.

[00219] If over the frequency range of interest, the conductivity, σ , of the material that composes the object is high enough that the material's skin depth may be considered short, then the electromagnetic fields, $\mathbf{E}$ and $\mathbf{H}$, may penetrate only a short distance into the object (namely they stay close to the 'skin' of the material, where δ is smaller than the characteristic thickness of the portion of the object that is lossy). In this case, the currents induced inside the material may be concentrated very close to the material surface, approximately within a skin depth, and their magnitude may be approximated by the product of a surface current density (mostly determined by the shape of the incident electromagnetic fields and, as long as the thickness of the conductor is much larger than the skin-depth, independent of frequency and conductivity to first order) $K(x,y)$ (where x and y are coordinates parameterizing the surface) and a function decaying exponentially into the surface: $\exp(-z/\delta)/\delta$ (where z denotes the coordinate locally normal to the surface): $J(x,y,z)=K(x,y)\exp(-z/\delta)/\delta$. Then, the dissipated power, P_d, may be estimated by,

$$P_d=\int_V d\mathrm{r}\,|\mathbf{J}(\mathrm{r})|^2/\sigma \simeq \left(\int_S \mathbf{dxdy}\,|\mathbf{K}(\mathbf{x},\mathbf{y})|^2\right)\left(\int_0^\infty \mathbf{dz}\exp(\mathbf{2z}/\delta)/(\sigma\delta^2)\right)=\sqrt{\mu\omega/8\sigma}\left(\int_S dxdy\,|K(x,y)|^2\right)$$

[00220] Therefore, in the high-conductivity limit, the dissipated power scales inverse proportionally to the square-root of the conductivity and goes to zero in the limit of a perfectly-conducting material.

[00221] If over the frequency range of interest, the conductivity, σ, of the material that composes the object is finite, then the material's skin depth, δ, may penetrate some distance into the object and some amount of power may be dissipated inside the object, depending also on the size of the object and the strength of the electromagnetic fields. This description can be generalized to also describe the general case of an object including multiple different materials with different properties and conductivities, such as an object with an arbitrary inhomogeneous and anisotropic distribution of the conductivity inside the object.

[00222] Note that the magnitude of the loss mechanisms described above may depend on the location and orientation of the extraneous objects relative to the resonator fields as well as the material composition of the extraneous objects. For example, high-conductivity materials may shift the resonant frequency of a resonator and detune it from other resonant objects. This frequency shift may be fixed by applying a feedback mechanism to a resonator that corrects its frequency, such as through changes in the inductance and/or capacitance of the resonator. These changes may be realized using variable capacitors and inductors, in some cases achieved by changes in the geometry of components in the resonators. Other novel tuning mechanisms, described below, may also be used to change the resonator frequency.

[00223] Where external losses are high, the perturbed Q may be low and steps may be taken to limit the absorption of resonator energy inside such extraneous objects and materials. Because of the functional dependence of the dissipated power on the strength of the electric and magnetic fields, one might optimize system performance by designing a system so that the desired coupling is achieved with shorter evanescent resonant field tails at the source resonator and longer at the device resonator, so that the perturbed Q of the source in the presence of other objects is optimized (or vice versa if the perturbed Q of the device needs to be optimized).

[00224] Note that many common extraneous materials and objects such as people, animals, plants, building materials, and the like, may have low conductivities and therefore may have little impact on the wireless energy transfer scheme disclosed here. An important fact related to the magnetic resonator designs we describe is that their electric fields may be confined primarily within the resonator structure itself, so it should be possible to operate within the commonly accepted guidelines for human safety while providing wireless power exchange over mid range distances.

[00225] **<u>Electromagnetic Resonators with Reduced Interactions</u>**

[00226] One frequency range of interest for near-field wireless power transmission is between 10 kHz and 100 MHz. In this frequency range, a large variety of ordinary non-metallic materials, such as for example several types of wood and plastic may have relatively low conductivity, such that only small amounts of power may be dissipated inside them. In addition, materials with low loss tangents, $\tan\Delta$, where $\tan\Delta = \varepsilon'' / \varepsilon'$, and ε'' and ε' are the imaginary and real parts of the permittivity respectively, may also have only small amounts of power dissipated inside them. Metallic materials, such as copper, silver, gold, and the like, with relatively high conductivity, may also have little power dissipated in them, because electromagnetic fields are not able to significantly penetrate these materials, as discussed earlier. These very high and very low conductivity materials, and low loss tangent materials and objects may have a negligible impact on the losses of a magnetic resonator.

[00227] However, in the frequency range of interest, there are materials and objects such as some electronic circuits and some lower-conductivity metals, which may have moderate (in general inhomogeneous and anisotropic) conductivity, and/or moderate to high loss tangents, and which may have relatively high dissipative losses. Relatively larger amounts of power may be dissipated inside them. These materials and objects may dissipate enough energy to reduce $Q_{(p)}$ by non-trivial amounts, and may be referred to as "lossy objects".

[00228] One way to reduce the impact of lossy materials on the $Q_{(p)}$ of a resonator is to use high-conductivity materials to shape the resonator fields such that they avoid the lossy objects. The process of using high-conductivity materials to tailor electromagnetic fields so that they avoid lossy objects in their vicinity may be understood by visualizing high-conductivity materials as materials that deflect or reshape the fields. This picture is qualitatively correct as long as the thickness of the conductor is larger than the skin-depth because the boundary conditions for electromagnetic fields at the surface of a good conductor force the electric field to be nearly completely perpendicular to, and the magnetic field to be nearly completely tangential to, the conductor surface. Therefore, a perpendicular magnetic field or a tangential electric field will be "deflected away" from the conducting surface. Furthermore, even a tangential magnetic field or a perpendicular electric field may be forced to decrease in magnitude on one side and/or in particular locations of the conducting surface, depending on the relative position of the sources of the fields and the conductive surface.

[00229] As an example, Fig. 18 shows a finite element method (FEM) simulation of two high conductivity surfaces 1802 above and below a lossy dielectric material 1804 in an external, initially uniform, magnetic field of frequency f= 6.78 MHz. The system is azimuthally symmetric around the r=0 axis. In this simulation, the lossy dielectric material 1804 is sandwiched between two conductors 1802, shown as the white lines at approximately z = ±0.01m. In the absence of the conducting surfaces above and below the dielectric disk, the magnetic field (represented by the drawn magnetic field lines) would have remained essentially uniform (field lines straight and parallel with the z-axis), indicating that the magnetic field would have passed straight through the lossy dielectric material. In this case, power would have been dissipated in the lossy dielectric disk. In the presence of conducting surfaces, however, this simulation shows the magnetic field is reshaped. The magnetic field is forced to be tangential to surface of the conductor and so is deflected around those conducting surfaces 1802, minimizing the amount of power that may be dissipated in the lossy dielectric material 1804 behind or between the conducting surfaces. As used herein, an axis of electrical symmetry refers to any axis about which a fixed or time-varying electrical or magnetic field is substantially symmetric during an exchange of energy as disclosed herein.

[00230] A similar effect is observed even if only one conducting surface, above or below, the dielectric disk, is used. If the dielectric disk is thin, the fact that the electric field is essentially zero at the surface, and continuous and smooth close to it, means that the electric field is very low everywhere close to the surface (i.e. within the dielectric disk). A single surface implementation for deflecting resonator fields away from lossy objects may be preferred for applications where one is not allowed to cover both sides of the lossy material or object (e.g. an LCD screen). Note that even a very thin surface of conducting material, on the order of a few skin-depths, may be sufficient (the skin depth in pure copper at 6.78 MHz is ~20 μm, and at 250 kHz is ~100 μm) to significantly improve the $Q_{(p)}$ of a resonator in the presence of lossy materials.

[00231] Lossy extraneous materials and objects may be parts of an apparatus, in which a high-Q resonator is to be integrated. The dissipation of energy in these lossy materials and objects may be reduced by a number of techniques including:

- by positioning the lossy materials and objects away from the resonator, or, in special positions and orientations relative to the resonator.
- by using a high conductivity material or structure to partly or entirely cover lossy materials and objects in the vicinity of a resonator
- by placing a closed surface (such as a sheet or a mesh) of high-conductivity material around a lossy object to completely cover the lossy object and shape the resonator fields such that they avoid the lossy object.
- by placing a surface (such as a sheet or a mesh) of a high-conductivity material around only a portion of a lossy object, such as along the top, the bottom, along the side, and the like, of an object or material.
- by placing even a single surface (such as a sheet or a mesh) of high-conductivity material above or below or on one side of a lossy object to reduce the strength of the fields at the location of the lossy object.

[00232] Fig. 19 shows a capacitively-loaded loop inductor forming a magnetic resonator 102 and a disk-shaped surface of high-conductivity material 1802 that completely surrounds a lossy object 1804 placed inside the loop inductor. Note that some lossy objects may be components, such as electronic circuits, that may need to interact with, communicate with, or be connected to the outside environment and thus cannot be completely electromagnetically isolated. Partially covering a lossy material with high conductivity materials may still reduce extraneous losses while enabling the lossy material or object to function properly.

[00233] Fig. 20 shows a capacitively-loaded loop inductor that is used as the resonator 102 and a surface of high-conductivity material 1802, surrounding only a portion of a lossy object 1804, that is placed inside the inductor loop.

[00234] Extraneous losses may be reduced, but may not be completely eliminated, by placing a single surface of high-conductivity material above, below, on the side, and the like, of a lossy object or material. An example is shown in Fig. 21, where a capacitively-loaded loop inductor is used as the resonator 102 and a surface of high-conductivity material 1802 is placed inside the inductor loop under a lossy object 1804 to reduce the strength of the fields at the location of the lossy object. It may be preferable to cover only one side of a material or object because of considerations of cost, weight, assembly complications, air flow, visual access, physical access, and the like.

[00235] A single surface of high-conductivity material may be used to avoid objects that cannot or should not be covered from both sides (e.g. LCD or plasma screens). Such lossy objects may be avoided using optically transparent conductors. High-conductivity optically opaque materials may instead be placed on only a portion of the lossy object, instead of, or in addition to, optically transparent conductors. The adequacy of single-sided vs. multi-sided covering implementations, and the design trade-offs inherent therein may depend on the details of the wireless energy transfer scenario and the properties of the lossy materials and objects.

[00236] Below we describe an example using high-conductivity surfaces to improve the Q-insensitivity, $\Theta_{(p)}$, of an integrated magnetic resonator used in a wireless energy-transfer system. Fig. 22 shows a wireless projector 2200. The wireless projector may include a device resonator 102C, a projector 2202, a wireless network/video adapter 2204, and power conversion circuits 2208, arranged as shown. The device resonator 102C may include a three-turn conductor loop, arranged to enclose a surface, and a capacitor network 2210. The conductor loop may be designed so that the device resonator 102C has a high Q (e.g., >100) at its operating resonant frequency. Prior to integration in the completely wireless projector 2200, this device resonator

102C has a Q of approximately 477 at the designed operating resonant frequency of 6.78 MHz. Upon integration, and placing the wireless network/video adapter card 2204 in the center of the resonator loop inductor, the resonator $Q_{(integrated)}$ was decreased to approximately 347. At least some of the reduction from Q to $Q_{(integrated)}$ was attributed to losses in the perturbing wireless network/video adapter card. As described above, electromagnetic fields associated with the magnetic resonator 102C may induce currents in and on the wireless network/video adapter card 2204, which may be dissipated in resistive losses in the lossy materials that compose the card. We observed that $Q_{(integrated)}$ of the resonator may be impacted differently depending on the composition, position, and orientation, of objects and materials placed in its vicinity.

[00237] In a completely wireless projector example, covering the network/video adapter card with a thin copper pocket (a folded sheet of copper that covered the top and the bottom of the wireless network/video adapter card, but not the communication antenna) improved the $Q_{(integrated)}$ of the magnetic resonator to a $Q_{(integrated\ +\ copper\ pocket)}$ of approximately 444. In other words, most of the reduction in $Q_{(integrated)}$ due to the perturbation caused by the extraneous network/video adapter card could be eliminated using a copper pocket to deflect the resonator fields away from the lossy materials.

[00238] In another completely wireless projector example, covering the network/video adapter card with a single copper sheet placed beneath the card provided a $Q_{(integrated\ +\ copper\ sheet)}$ approximately equal to $Q_{(integrated\ +\ copper\ pocket)}$. In that example, the high perturbed Q of the system could be maintained with a single high-conductivity sheet used to deflect the resonator fields away from the lossy adapter card.

[00239] It may be advantageous to position or orient lossy materials or objects, which are part of an apparatus including a high-Q electromagnetic resonator, in places where the fields produced by the resonator are relatively weak, so that little or no power may be dissipated in these objects and so that the Q-insensitivity, $\Theta_{(p)}$, may be large. As was shown earlier, materials of different conductivity may respond differently to electric versus magnetic fields. Therefore, according to the conductivity of the extraneous object, the positioning technique may be specialized to one or the other field.

[00240] Fig. 23 shows the magnitude of the electric 2312 and magnetic fields 2314 along a line that contains the diameter of the circular loop inductor and the electric 2318 and magnetic fields 2320 along the axis of the loop inductor for a capacitively-loaded circular loop inductor of wire of radius 30 cm, resonant at 10 MHz. It can be seen that the amplitude of the resonant near-fields reach their maxima close to the wire and decay away from the loop, 2312 , 2314 . In the plane of the loop inductor 2318, 2320, the fields reach a local minimum at the center of the loop. Therefore, given the finite size of the apparatus, it may be that the fields are weakest at the extrema of the apparatus or it may be that the field magnitudes have local minima somewhere within the apparatus. This argument holds for any other type of electromagnetic resonator 102 and any type of apparatus. Examples are shown in Figs. 24a and 24b, where a capacitively-loaded inductor loop forms a magnetic resonator 102 and an extraneous lossy object 1804 is positioned where the electromagnetic fields have minimum magnitude.

[00241] In a demonstration example, a magnetic resonator was formed using a three-turn conductor loop, arranged to enclose a square surface (with rounded corners), and a capacitor network. The Q of the resonator was approximately 619 at the designed operating resonant frequency of 6.78 MHz. The perturbed Q of this resonator depended on the placement of the perturbing object, in this case a pocket projector, relative to the resonator. When the perturbing projector was located inside the inductor loop and at its center or on top of the inductor wire turns, $Q_{(projector)}$ was approximately 96, lower than when the perturbing projector was placed outside of the resonator, in which case $Q_{(projector)}$ was approximately 513. These measurements support the analysis that shows the fields inside the inductor loop may be larger than those outside it, so lossy objects placed inside such a loop inductor may yield lower perturbed *Q's* for the system than when the lossy object is placed outside the loop inductor. Depending on the resonator designs and the material composition and orientation of the lossy object, the arrangement shown in Fig. 24b may yield a higher Q-insensitivity, $\Theta_{(projector)}$, than the arrangement shown in Fig. 24a.

[00242] High-Q resonators may be integrated inside an apparatus. Extraneous materials and objects of high dielectric permittivity, magnetic permeability, or electric conductivity may be part of the apparatus into which a high-Q resonator is to be integrated. For these extraneous materials and objects in the vicinity of a high-Q electromagnetic resonator, depending on their size, position and orientation relative to the resonator, the resonator field-profile may be distorted and deviate significantly from the original unperturbed field-profile of the resonator. Such a distortion of the unperturbed fields of the resonator may significantly decrease the Q to a lower $Q_{(p)}$, even if the extraneous objects and materials are lossless.

[00243] It may be advantageous to position high-conductivity objects, which are part of an apparatus including a high-Q electromagnetic resonator, at orientations such that the surfaces of these objects are, as much as possible, perpendicular to the electric field lines produced by the unperturbed resonator and parallel to the magnetic field lines produced by the unperturbed resonator, thus distorting the resonant field profiles by the smallest amount possible. Other common objects that may be positioned perpendicular to the plane of a magnetic resonator loop include screens (LCD, plasma, etc), batteries, cases, connectors, radiative antennas, and the like. The Q-insensitivity, $\Theta_{(p)}$, of the resonator may be much larger than if the objects were positioned at a different orientation with respect to the resonator fields.

[00244] Lossy extraneous materials and objects, which are not part of the integrated apparatus including a high-Q resonator, may be located or brought in the vicinity of the resonator, for example, during the use of the apparatus. It may be advantageous in certain circumstances to use high conductivity materials to tailor the resonator fields so that they avoid the regions where lossy extraneous objects may be located or introduced to reduce power dissipation in these materials and objects and to increase Q-insensitivity, $\Theta_{(p)}$. An example is shown in Fig. 25, where a capacitively-loaded loop inductor and capacitor are used as the resonator 102 and a surface of high-conductivity material 1802 is placed above the inductor loop to reduce the magnitude of the fields in the region above the resonator, where lossy extraneous objects 1804 may be located or introduced.

[00245] Note that a high-conductivity surface brought in the vicinity of a resonator to reshape the fields may also lead to $Q_{(cond.\ surface)}<Q$. The reduction in the perturbed Q may be due to the dissipation of energy inside the lossy conductor or to the distortion of the unperturbed resonator field profiles associated with matching the field boundary conditions at the surface of the conductor. Therefore, while a high-conductivity surface may be used to reduce the extraneous losses due to dissipation inside an extraneous lossy object, in some cases, especially in some of those where this is achieved by significantly reshaping the electromagnetic fields, using such a high-conductivity surface so that the fields avoid the lossy object may result effectively in $Q_{(p\ +\ cond.\ surface)}<Q_{(p)}$ rather than the desired result $Q_{(p\ +\ cond.\ surface)}>Q_{(p)}$.

[00246] As described above, in the presence of loss inducing objects, the perturbed quality factor of a magnetic resonator may be improved if the electromagnetic fields associated with the magnetic resonator are reshaped to avoid the loss inducing objects. Another way to reshape the unperturbed resonator fields is to use high permeability materials to completely or partially enclose or cover the loss inducing objects, thereby reducing the interaction of the magnetic field with the loss inducing objects.

[00247] Magnetic field shielding has been described previously, for example in *Electrodynamics 3rd Ed.*, Jackson, pp. 201-203. There, a spherical shell of magnetically permeable material was shown to shield its interior from external magnetic fields. For example, if a shell of inner radius a, outer radius b, and relative permeability μ_r, is placed in an initially uniform magnetic field H_0, then the field inside the shell will have a constant magnitude, $9\mu_r H_0 \big/ \left[(2\mu_r+1)(\mu_r+2)-2(a/b)^3(\mu_r-1)^2\right]$, which tends to $9H_0 \big/ 2\mu_r\left(1-(a/b)^3\right)$ if $\mu_r >> 1$. This result shows that an incident magnetic field (but not necessarily an incident electric field) may be greatly attenuated inside the shell, even if the shell is quite thin, provided the magnetic permeability is high enough. It may be advantageous in certain circumstances to use high permeability materials to partly or entirely cover lossy materials and objects so that they are avoided by the resonator magnetic fields and so that little or no power is dissipated in these materials and objects. In such an approach, the Q-insensitivity, $\Theta_{(p)}$, may be larger than if the materials and objects were not covered, possibly larger than 1.

[00248] It may be desirable to keep both the electric and magnetic fields away from loss inducing objects. As described above, one way to shape the fields in such a manner is to use high-conductivity surfaces to either completely or partially enclose or cover the loss inducing objects. A layer of magnetically permeable material, also referred to as magnetic material, (any material or meta-material having a non-trivial magnetic permeability), may be placed on or around the high-conductivity surfaces. The additional layer of magnetic material may present a lower reluctance path (compared to free space) for the deflected magnetic field to follow and may partially shield the electric conductor underneath it from the incident magnetic flux. This arrangement may reduce the losses due to induced currents in the high-conductivity surface. Under some circumstances the lower reluctance path presented by the magnetic material may improve the perturbed Q of the structure.

[00249] Fig. 26a shows an axially symmetric FEM simulation of a thin conducting 2604 (copper) disk (20 cm in diameter, 2 cm in height) exposed to an initially uniform, externally applied magnetic field (gray flux lines) along the z-axis. The axis of symmetry is at

r=0. The magnetic streamlines shown originate at $z = -\infty$, where they are spaced from r=3 cm to r=10 cm in intervals of 1 cm. The axes scales are in meters. Imagine, for example, that this conducing cylinder encloses loss-inducing objects within an area circumscribed by a magnetic resonator in a wireless energy transfer system such as shown in Fig. 19.

[00250] This high-conductivity enclosure may increase the perturbing Q of the lossy objects and therefore the overall perturbed Q of the system, but the perturbed Q may still be less than the unperturbed Q because of induced losses in the conducting surface and changes to the profile of the electromagnetic fields. Decreases in the perturbed Q associated with the high-conductivity enclosure may be at least partially recovered by including a layer of magnetic material along the outer surface or surfaces of the high-conductivity enclosure. Fig. 26b shows an axially symmetric FEM simulation of the thin conducting 2604A (copper) disk (20 cm in diameter, 2 cm in height) from Fig. 26a, but with an additional layer of magnetic material placed directly on the outer surface of the high-conductivity enclosure. Note that the presence of the magnetic material may provide a lower reluctance path for the magnetic field, thereby at least partially shielding the underlying conductor and reducing losses due to induced eddy currents in the conductor.

[00251] Fig. 27 depicts a variation (in axi-symmetric view) to the system shown in Fig. 26 where not all of the lossy material 2708 may be covered by a high-conductivity surface 2706. In certain circumstances it may be useful to cover only one side of a material or object, such as due to considerations of cost, weight, assembly complications, air flow, visual access, physical access, and the like. In the exemplary arrangement shown in Fig. 27, only one surface of the lossy material 2708 is covered and the resonator inductor loop is placed on the opposite side of the high-conductivity surface.

[00252] Mathematical models were used to simulate a high-conductivity enclosure made of copper and shaped like a 20 cm diameter by 2 cm high cylindrical disk placed within an area circumscribed by a magnetic resonator whose inductive element was a single-turn wire loop with loop radius r=11 cm and wire radius a= 1 mm. Simulations for an applied 6.78 MHz electromagnetic field suggest that the perturbing quality factor of this high-conductivity enclosure, $\delta Q_{(enclosure)}$, is 1,870. When the high-conductivity enclosure was modified to include a 0.25 cm-thick layer of magnetic material with real relative permeability, $\mu_r' = 40$, and imaginary relative permeability, $\mu_r'' = 10^{-2}$, simulations suggest the perturbing quality factor is increased to $\delta Q_{(enclosure+magnetic\ material)}$=5,060.

[00253] The improvement in performance due to the addition of thin layers of magnetic material 2702 may be even more dramatic if the high-conductivity enclosure fills a larger portion of the area circumscribed by the resonator's loop inductor 2704. In the example above, if the radius of the inductor loop 2704 is reduced so that it is only 3 mm away from the surface of the high-conductivity enclosure, the perturbing quality factor may be improved from 670 (conducting enclosure only) to 2,730 (conducting enclosure with a thin layer of magnetic material) by the addition of a thin layer of magnetic material 2702around the outside of the enclosure.

[00254] The resonator structure may be designed to have highly confined electric fields, using shielding, or distributed capacitors, for example, which may yield high, even when the resonator is very close to materials that would typically induce loss.

[00255] **<u>Coupled Electromagnetic Resonators</u>**

[00256] The efficiency of energy transfer between two resonators may be determined by the strong-coupling figure-of-merit, $U = \kappa/\sqrt{\Gamma_s \Gamma_d} = \left(2\kappa/\sqrt{\omega_s \omega_d}\right)\sqrt{Q_s Q_d}$. In magnetic resonator implementations the coupling factor between the two resonators may be related to the inductance of the inductive elements in each of the resonators, L_1 and L_2, and the mutual inductance, *M,* between them by $\kappa_{12} = \omega M / 2\sqrt{L_1 L_2}$. Note that this expression assumes there is negligible coupling through electric-dipole coupling. For capacitively-loaded inductor loop resonators where the inductor loops are formed by circular conducting loops with N turns, separated by a distance D, and oriented as shown in Fig. 1(b), the mutual inductance is $M = \pi / 4 \cdot \mu_o N_1 N_2 (x_1 x_2)^2 / D^3$ where x_1 , N_1 and x_2 , N_2 are the characteristic size and number of turns of the conductor loop of the first and second resonators respectively. Note that this is a quasi-static result, and so assumes that the resonator's size is much smaller than the wavelength and the resonators' distance is much smaller than the wavelength, but also that their distance is at least a few times their size. For these circular resonators operated in the quasi-static limit and at mid-range distances, as described above, $k = 2\kappa / \sqrt{\omega_1 \omega_2} \sim \left(\sqrt{x_1 x_2}/D\right)^3$. Strong coupling (a large

U) between resonators at mid-range distances may be established when the quality factors of the resonators are large enough to compensate for the small k at mid-range distances

[00257] For electromagnetic resonators, if the two resonators include conducting parts, the coupling mechanism may be that currents are induced on one resonator due to electric and magnetic fields generated from the other. The coupling factor may be proportional to the flux of the magnetic field produced from the high-Q inductive element in one resonator crossing a closed area of the high-Q inductive element of the second resonator.

[00258] **Coupled Electromagnetic Resonators with Reduced Interactions**

[00259] As described earlier, a high-conductivity material surface may be used to shape resonator fields such that they avoid lossy objects, p, in the vicinity of a resonator, thereby reducing the overall extraneous losses and maintaining a high Q-insensitivity $\Theta_{(p\ +\ cond.\ surface)}$ of the resonator. However, such a surface may also lead to a perturbed coupling factor, $\mathrm{k}_{(\mathrm{p\ +\ cond.\ surface})}$, between resonators that is smaller than the perturbed coupling factor, $k_{(p)}$ and depends on the size, position, and orientation of the high-conductivity material relative to the resonators. For example, if high-conductivity materials are placed in the plane and within the area circumscribed by the inductive element of at least one of the magnetic resonators in a wireless energy transfer system, some of the magnetic flux through the area of the resonator, mediating the coupling, may be blocked and k may be reduced.

[00260] Consider again the example of Fig. 19. In the absence of the high-conductivity disk enclosure, a certain amount of the external magnetic flux may cross the circumscribed area of the loop. In the presence of the high-conductivity disk enclosure, some of this magnetic flux may be deflected or blocked and may no longer cross the area of the loop, thus leading to a smaller perturbed coupling factor $k_{12(p\ +\ cond.\ surfaces)}$. However, because the deflected magnetic-field lines may follow the edges of the high-conductivity surfaces closely, the reduction in the flux through the loop circumscribing the disk may be less than the ratio of the areas of the face of the disk to the area of the loop.

[00261] One may use high-conductivity material structures, either alone, or combined with magnetic materials to optimize perturbed quality factors, perturbed coupling factors, or perturbed efficiencies.

[00262] Consider the example of Fig. 21. Let the lossy object have a size equal to the size of the capacitively-loaded inductor loop resonator, thus filling its area A 2102. A high-conductivity surface 1802 may be placed under the lossy object 1804. Let this be resonator 1 in a system of two coupled resonators 1 and 2, and let us consider how $U_{12(object + cond. surface)}$ scales compared to U_{12} as the area A_s 2104 of the conducting surface increases. Without the conducting surface 1802 below the lossy object 1804, the k-insensitivity, $\beta_{12(object)}$, may be approximately one, but the Q-insensitivity, $\Theta_{1(object)}$, may be small, so the U-insensitivity $\Xi_{12(object)}$ may be small.

[00263] Where the high-conductivity surface below the lossy object covers the entire area of the inductor loop resonator (A_s=A), $k_{12(object + cond. surface)}$ may approach zero, because little flux is allowed to cross the inductor loop, so $U_{12(object + cond. surface)}$ may approach zero. For intermediate sizes of the high-conductivity surface, the suppression of extrinsic losses and the associated Q-insensitivity, $\Theta_{1(object + cond. surface)}$, may be large enough compared to $\Theta_{1(object)}$, while the reduction in coupling may not be significant and the associated k-insensitivity, $\beta_{12(object + cond. surface)}$, may be not much smaller than $\beta_{12(object)}$, so that the overall $U_{12(object + cond. surface)}$ may be increased compared to $U_{12(object)}$. The optimal degree of avoiding of extraneous lossy objects via high-conductivity surfaces in a system of wireless energy transfer may depend on the details of the system configuration and the application.

[00264] We describe using high-conductivity materials to either completely or partially enclose or cover loss inducing objects in the vicinity of high-Q resonators as one potential method to achieve high perturbed Q's for a system. However, using a good conductor alone to cover the objects may reduce the coupling of the resonators as described above, thereby reducing the efficiency of wireless power transfer. As the area of the conducting surface approaches the area of the magnetic resonator, for example, the perturbed coupling factor, $k_{(p)}$, may approach zero, making the use of the conducting surface incompatible with efficient wireless power transfer.

[00265] One approach to addressing the aforementioned problem is to place a layer of magnetic material around the high-conductivity materials because the additional layer of permeable material may present a lower reluctance path (compared to free space) for the deflected magnetic field to follow and may partially shield the electric conductor underneath it from incident magnetic flux. Under some circumstances the lower reluctance path presented by the magnetic material may improve the electromagnetic coupling of the resonator to other resonators. Decreases in the perturbed coupling factor associated with using conducting materials to tailor resonator fields so that they avoid lossy objects in and around high-Q magnetic resonators may be at least partially recovered by including a layer of magnetic material along the outer surface or surfaces of the conducting materials. The magnetic materials may increase the perturbed coupling factor relative to its initial unperturbed value.

[00266] Note that the simulation results in Fig. 26 show that an incident magnetic field may be deflected less by a layered magnetic material and conducting structure than by a conducting structure alone. If a magnetic resonator loop with a radius only slightly larger than that of the disks shown in Figs. 26(a) and 26(b) circumscribed the disks, it is clear that more flux lines would be captured in the case illustrated in Fig. 26(b) than in Fig. 26(a), and therefore $k_{(disk)}$ would be larger for the case illustrated in Fig. 26(b). Therefore, including a layer of magnetic material on the conducting material may improve the overall system performance. System analyses may be performed to determine whether these materials should be partially, totally, or minimally integrated into the resonator.

[00267] As described above, Fig. 27 depicts a layered conductor 2706 and magnetic material 2702 structure that may be appropriate for use when not all of a lossy material 2708 may be covered by a conductor and/or magnetic material structure. It was shown earlier that for a copper conductor disk with a 20 cm diameter and a 2 cm height, circumscribed by a resonator with an inductor loop radius of 11 cm and a wire radius a=1 mm, the calculated perturbing Q for the copper cylinder was 1,870. If the resonator and the conducting disk shell are placed in a uniform magnetic field (aligned along the axis of symmetry of the inductor loop), we calculate that the copper conductor has an associated coupling factor insensitivity of 0.34. For comparison, we model the same arrangement but include a 0.25 cm-thick layer of magnetic material with a real relative permeability, $\mu_r' = 40$, and an imaginary relative permeability, $\mu_r'' = 10^{-2}$. Using the same model and parameters described above, we find that the coupling factor insensitivity is improved to 0.64 by the addition of the magnetic material to the surface of the conductor.

[00268] Magnetic materials may be placed within the area circumscribed by the magnetic resonator to increase the coupling in wireless energy transfer systems. Consider a solid sphere of a magnetic material with relative permeability, μ_r, placed in an initially uniform magnetic field. In this example, the lower reluctance path offered by the magnetic material may cause the magnetic field to concentrate in the volume of the sphere. We find that the magnetic flux through the area circumscribed by the equator of the sphere is enhanced by a factor of

$3\mu_r/(\mu_r+2)$, by the addition of the magnetic material. If μ_r>>*1*, this enhancement factor may be close to 3.

[00269] One can also show that the dipole moment of a system comprising the magnetic sphere circumscribed by the inductive element in a magnetic resonator would have its magnetic dipole enhanced by the same factor. Thus, the magnetic sphere with high permeability practically triples the dipole magnetic coupling of the resonator. It is possible to keep most of this increase in coupling if we use a spherical shell of magnetic material with inner radius *a,* and outer radius *b,* even if this shell is on top of block or enclosure made from highly conducting materials. In this case, the enhancement in the flux through the equator is

$$\frac{3\mu_r\left(1-\left(\frac{a}{b}\right)^3\right)}{\mu_r\left(1-\left(\frac{a}{b}\right)^3\right)+2\left(1+\frac{1}{2}\left(\frac{a}{b}\right)^3\right)}.$$

For μ_r=1,000 and *(a/b)*=0.99, this enhancement factor is still 2.73, so it possible to significantly improve the coupling even with thin layers of magnetic material.

[00270] As described above, structures containing magnetic materials may be used to realize magnetic resonators. Fig. 16(a) shows a 3 dimensional model of a copper and magnetic material structure 1600 driven by a square loop of current around the choke point at its center. Fig. 16(b) shows the interaction, indicated by magnetic field streamlines , between two identical structures 1600A-B with the same properties as the one shown in Fig. 16(a). Because of symmetry, and to reduce computational complexity, only one half of the system is modeled. If we fix the relative orientation between the two objects and vary their center-to-center distance (the image shown is at a relative separation of 50 cm), we find that, at 300 kHz, the coupling efficiency varies from 87% to 55% as the separation between the structures varies from 30 cm to 60 cm. Each of the example structures shown 1600 A-B includes two 20 cm x 8 cm x 2cm parallelepipeds made of copper joined by a 4 cm x 4 cm x 2 cm block of magnetic material and entirely covered with a 2 mm layer of the same magnetic material (assumed to have μ_r=1,400+j5). Resistive losses in the driving loop are ignored. Each structure has a calculated *Q* of 815.

[00271] ELECTROMAGNETIC RESONATORS AND IMPEDANCE MATCHING

[00272] Impedance Matching Architectures for Low-Loss Inductive Elements

[00273] For purposes of the present discussion, an inductive element may be any coil or loop structure (the 'loop') of any conducting material, with or without a (gapped or ungapped) core made of magnetic material, which may also be coupled inductively or in any other contactless way to other systems. The element is inductive because its impedance, including both the impedance of the loop and the so-called 'reflected' impedances of any potentially coupled systems, has positive reactance, X, and resistance, *R*.

[00274] Consider an external circuit, such as a driving circuit or a driven load or a transmission line, to which an inductive element may be connected. The external circuit (e.g. a driving circuit) may be delivering power to the inductive element and the inductive element may be delivering power to the external circuit (e.g. a driven load). The efficiency and amount of power delivered between the inductive element and the external circuit at a desired frequency may depend on the impedance of the inductive element relative to the properties of the external circuit. Impedance-matching networks and external circuit control techniques may be used to regulate the power delivery between the external circuit and the inductive element, at a desired frequency, *f*.

[00275] The external circuit may be a driving circuit configured to form a amplifier of class A, B, C, D, DE, E, F and the like, and may deliver power at maximum efficiency (namely with minimum losses within the driving circuit) when it is driving a resonant network with specific impedance Z_o^*, where Z_o may be complex and * denotes complex conjugation. The external circuit may be a driven load configured to form a rectifier of class A, B, C, D, DE, E, F and the like, and may receive power at maximum efficiency (namely with minimum losses within the driven load) when it is driven by a resonant network with specific impedance Z_o^*, where Z_o may be complex. The external circuit may be a transmission line with characteristic impedance, Z_o, and may exchange power at maximum efficiency (namely with zero reflections) when connected to an impedance Z_o^*. We will call the characteristic impedance Z_o of an external circuit the complex conjugate of the impedance that may be connected to it for power exchange at maximum efficiency.

[00276] Typically the impedance of an inductive element, $R+jX$, may be much different from Z_o^*. For example, if the inductive element has low loss (a high X/R), its resistance, R, may be much lower than the real part of the characteristic impedance, Z_0, of the external circuit. Furthermore, an inductive element by itself may not be a resonant network. An impedance-matching network connected to an inductive element may typically create a resonant network, whose impedance may be regulated.

[00277] Therefore, an impedance-matching network may be designed to maximize the efficiency of the power delivered between the external circuit and the inductive element (including the reflected impedances of any coupled systems). The efficiency of delivered power may be maximized by matching the impedance of the combination of an impedance-matching network and an inductive element to the characteristic impedance of an external circuit (or transmission line) at the desired frequency.

[00278] An impedance-matching network may be designed to deliver a specified amount of power between the external circuit and the inductive element (including the reflected impedances of any coupled systems). The delivered power may be determined by adjusting the complex ratio of the impedance of the combination of the impedance-matching network and the inductive element to the impedance of the external circuit (or transmission line) at the desired frequency.

[00279] Impedance-matching networks connected to inductive elements may create magnetic resonators. For some applications, such as wireless power transmission using strongly-coupled magnetic resonators, a high Q may be desired for the resonators. Therefore, the inductive element may be chosen to have low losses (high X/R).

[00280] Since the matching circuit may typically include additional sources of loss inside the resonator, the components of the matching circuit may also be chosen to have low losses. Furthermore, in high-power applications and/or due to the high resonator Q, large currents may run in parts of the resonator circuit and large voltages may be present across some circuit elements within the resonator. Such currents and voltages may exceed the specified tolerances for particular circuit elements and may be too high for particular components to withstand. In some cases, it may be difficult to find or implement components, such as tunable capacitors for example, with size, cost and performance (loss and current/voltage-rating) specifications sufficient to realize high-Q and high-power resonator designs for certain applications. We disclose matching circuit designs, methods, implementations and techniques that may preserve the high Q for magnetic resonators, while reducing the component requirements for low loss and/or high current/voltage-rating.

[00281] Matching-circuit topologies may be designed that minimize the loss and current-rating requirements on some of the elements of the matching circuit. The topology of a circuit matching a low-loss inductive element to an impedance, Z_0, may be chosen so that some of its components lie outside the associated high-Q resonator by being in series with the external circuit. The requirements for low series loss or high current-ratings for these components may be reduced. Relieving the low series loss and/or high-current-rating requirement on a circuit element may be particularly useful when the element needs to be variable and/or to have a large voltage-rating and/or low parallel loss.

[00282] Matching-circuit topologies may be designed that minimize the voltage rating requirements on some of the elements of the matching circuit. The topology of a circuit matching a low-loss inductive element to an impedance, Z_0, may be chosen so that some of its components lie outside the associated high-Q resonator by being in parallel with Z_0. The requirements for low parallel loss or high voltage-rating for these components may be reduced. Relieving the low parallel loss and/or high-voltage requirement on a circuit element may be particularly useful when the element needs to be variable and/or to have a large current-rating and/or low series loss.

[00283] The topology of the circuit matching a low-loss inductive element to an external characteristic impedance, Z_0, may be chosen so that the field pattern of the associated resonant mode and thus its high Q are preserved upon coupling of the resonator to the external impedance. Otherwise inefficient coupling to the desired resonant mode may occur (potentially due to coupling to other undesired resonant modes), resulting in an effective lowering of the resonator Q.

[00284] For applications where the low-loss inductive element or the external circuit, may exhibit variations, the matching circuit may need to be adjusted dynamically to match the inductive element to the external circuit impedance, Z_0, at the desired frequency, f. Since there may typically be two tuning objectives, matching or controlling both the real and imaginary part of the impedance level, Z_0, at the desired frequency, f, there may be two variable elements in the matching circuit. For inductive elements, the matching circuit may need to include at least one variable capacitive element.

[00285] A low-loss inductive element may be matched by topologies using two variable capacitors, or two networks of variable capacitors. A variable capacitor may, for example, be a tunable butterfly-type capacitor having, e.g., a center terminal for connection to a ground or other lead of a power source or load, and at least one other terminal across which a capacitance of the tunable butterfly-type capacitor can be varied or tuned, or any other capacitor having a user-configurable, variable capacitance.

[00286] A low-loss inductive element may be matched by topologies using one, or a network of, variable capacitor(s) and one, or a network of, variable inductor(s).

[00287] A low-loss inductive element may be matched by topologies using one, or a network of, variable capacitor(s) and one, or a network of, variable mutual inductance(s), which transformer-couple the inductive element either to an external circuit or to other systems.

[00288] In some cases, it may be difficult to find or implement tunable lumped elements with size, cost and performance specifications sufficient to realize high-Q, high-power, and potentially high-speed, tunable resonator designs. The topology of the circuit matching a variable inductive element to an external circuit may be designed so that some of the variability is assigned to the external circuit by varying the frequency, amplitude, phase, waveform, duty cycle, and the like, of the drive signals applied to transistors, diodes, switches and the like, in the external circuit.

[00289] The variations in resistance, R, and inductance, L, of an inductive element at the resonant frequency may be only partially compensated or not compensated at all. Adequate system performance may thus be preserved by tolerances designed into other system components or specifications. Partial adjustments, realized using fewer tunable components or less capable tunable components, may be sufficient.

[00290] Matching-circuit architectures may be designed that achieve the desired variability of the impedance matching circuit under high-power conditions, while minimizing the voltage/current rating requirements on its tunable elements and achieving a finer (i.e. more precise, with higher resolution) overall tunability. The topology of the circuit matching a variable inductive element to an impedance, Z_0, may include appropriate combinations and placements of fixed and variable elements, so that the voltage/current requirements for the variable components may be reduced and the desired tuning range may be covered with finer tuning resolution. The voltage/current requirements may be reduced on components that are not variable.

[00291] The disclosed impedance matching architectures and techniques may be used to achieve the following:

- To maximize the power delivered to, or to minimize impedance mismatches between, the source low-loss inductive elements (and any other systems wirelessly coupled to them) from the power driving generators.
- To maximize the power delivered from, or to minimize impedance mismatches between, the device low-loss inductive elements (and any other systems wirelessly coupled to them) to the power driven loads.
- To deliver a controlled amount of power to, or to achieve a certain impedance relationship between, the source low-loss inductive elements (and any other systems wirelessly coupled to them) from the power driving generators.
- To deliver a controlled amount of power from, or to achieve a certain impedance relationship between, the device low-loss inductive elements (and any other systems wirelessly coupled to them) to the power driven loads.

[00292] TOPOLOGIES FOR PRESERVATION OF MODE PROFILE (HIGH-*Q*)

[00293] The resonator structure may be designed to be connected to the generator or the load wirelessly (indirectly) or with a hard-wired connection (directly).

[00294] Consider a general indirectly coupled matching topology such as that shown by the block diagram in Fig. 28(a). There, an inductive element 2802 , labeled as (R,L) and represented by the circuit symbol for an inductor, may be any of the inductive elements discussed in this disclosure or in the references provided herein, and where an impedance-matching circuit 2402 includes or consists of parts A and B. B may be the part of the matching circuit that connects the impedance 2804 , Z_0, to the rest of the circuit (the combination of A and the inductive element (A+(R,L)) via a wireless connection (an inductive or capacitive coupling mechanism).

[00295] The combination of A and the inductive element 2802 may form a resonator 102, which in isolation may support a high-Q resonator electromagnetic mode, with an associated current and charge distribution. The lack of a wired connection between the external circuit, Z_0 and B, and the resonator, A + (R,L), may ensure that the high-Q resonator electromagnetic mode and its current/charge distributions may take the form of its intrinsic (in-isolation) profile, so long as the degree of wireless coupling is not too large. That is, the electromagnetic mode, current/charge distributions, and thus the high-Q of the resonator may be automatically maintained using an indirectly coupled matching topology.

[00296] This matching topology may be referred to as indirectly coupled, or transformer-coupled, or inductively-coupled, in the case where inductive coupling is used between the external circuit and the inductor loop. This type of coupling scenario was used to couple the power supply to the source resonator and the device resonator to the light bulb in the demonstration of wireless energy transfer over mid-range distances described in the referenced *Science* article.

[00297] Next consider examples in which the inductive element may include the inductive element and any indirectly coupled systems. In this case, as disclosed above, and again because of the lack of a wired connection between the external circuit or the coupled systems and the resonator, the coupled systems may not, with good approximation for not-too-large degree of indirect coupling, affect the resonator electromagnetic mode profile and the current/charge distributions of the resonator. Therefore, an indirectly-coupled matching circuit may work equally well for any general inductive element as part of a resonator as well as for inductive elements wirelessly-coupled to other systems, as defined herein. Throughout this disclosure, the matching topologies we disclose refer to matching topologies for a general inductive element of this type, that is, where any additional systems may be indirectly coupled to the low-loss inductive element, and it is to be understood that those additional systems do not greatly affect the resonator electromagnetic mode profile and the current/charge distributions of the resonator.

[00298] Based on the argument above, in a wireless power transmission system of any number of coupled source resonators, device resonators and intermediate resonators the wireless magnetic (inductive) coupling between resonators does not affect the electromagnetic mode profile and the current/charge distributions of each one of the resonators. Therefore, when these resonators have a high (unloaded and unperturbed) Q, their (unloaded and unperturbed) Q may be preserved in the presence of the wireless coupling. (Note that the loaded Q of a resonator may be reduced in the presence of wireless coupling to another resonator, but we may be interested in preserving the unloaded Q, which relates only to loss mechanisms and not to coupling/loading mechanisms.)

[00299] Consider a matching topology such as is shown in Fig. 28(b). The capacitors shown in Fig. 28(b) may represent capacitor circuits or networks. The capacitors shown may be used to form the resonator 102 and to adjust the frequency and/or impedance of the source and device resonators. This resonator 102 may be directly coupled to an impedance, Z_0, using the ports labeled "terminal connections" 2808. Fig. 28(c) shows a generalized directly coupled matching topology, where the impedance-matching circuit 2602 includes or consists of parts A, B and C. Here, circuit elements in A, B and C may be considered part of the resonator 102 as well as part of the impedance matching 2402 (and frequency tuning) topology. B and C may be the parts of the matching circuit 2402 that connect the impedance Z_0 2804 (or the network terminals) to the rest of the circuit (A and the inductive element) via a single wire connection each. Note that B and C could be empty (short-circuits). If we disconnect or open circuit parts B and C (namely those single wire connections), then, the combination of A and the inductive element (R,L) may form the resonator.

[00300] The high-Q resonator electromagnetic mode may be such that the profile of the voltage distribution along the inductive element has nodes, namely positions where the voltage is zero. One node may be approximately at the center of the length of the inductive element, such as the center of the conductor used to form the inductive element, (with or without magnetic materials) and at least one other node may be within A. The voltage distribution may be approximately anti-symmetric along the inductive element with respect to its voltage node. A high Q may be maintained by designing the matching topology (A, B, C) and/or the terminal voltages (V1, V2) so that this high-Q resonator electromagnetic mode distribution may be approximately preserved on the inductive element. This high-Q resonator electromagnetic mode distribution may be approximately preserved on the inductive element by preserving the voltage node (approximately at the center) of the inductive element. Examples that achieve these design goals are provided herein.

[00301] A, B, and C may be arbitrary (namely not having any special symmetry), and V1 and V2 may be chosen so that the voltage across the inductive element is symmetric (voltage node at the center inductive). These results may be achieved using simple matching circuits but potentially complicated terminal voltages, because a topology-dependent common-mode signal (V1+V2)/2 may be required on both terminals.

[00302] Consider an 'axis' that connects all the voltage nodes of the resonator, where again one node is approximately at the center of the length of the inductive element and the others within A. (Note that the 'axis' is really a set of points (the voltage nodes) within the electric-circuit topology and may not necessarily correspond to a linear axis of the actual physical structure. The 'axis' may align with a physical axis in cases where the physical structure has symmetry.) Two points of the resonator are electrically symmetric with respect to the 'axis', if the impedances seen between each of the two points and a point on the 'axis', namely a voltage-node point of the resonator, are the same.

[00303] B and C may be the same (C=B), and the two terminals may be connected to any two points of the resonator (A + (R,L)) that are electrically symmetric with respect to the 'axis' defined above and driven with opposite voltages (V2=-V1) as shown in Fig. 28(d). The two electrically symmetric points of the resonator 102 may be two electrically symmetric points on the inductor loop. The two electrically symmetric points of the resonator may be two electrically symmetric points inside A. If the two electrically symmetric points, (to which each of the equal parts B and C is connected), are inside A, A may need to be designed so that these electrically-symmetric points are accessible as connection points within the circuit. This topology may be referred to as a 'balanced drive' topology. These balanced-drive examples may have the advantage that any common-mode signal that may be present on the ground line, due to perturbations at the external circuitry or the power network, for example, may be automatically rejected (and may not reach the resonator). In some balanced-drive examples, this topology may require more components than other topologies.

[00304] In other examples, C may be chosen to be a short-circuit and the corresponding terminal to be connected to ground (V=0) and to any point on the electric-symmetry (zero-voltage) 'axis' of the resonator, and B to be connected to any other point of the resonator not on the electric-symmetry 'axis', as shown in Fig. 28(e). The ground-connected point on the electric-symmetry 'axis' may be the voltage node on the inductive element, approximately at the center of its conductor length. The ground-connected point on the electric-symmetry 'axis' may be inside the circuit A. Where the ground-connected point on the electric-symmetry 'axis' is inside A, A may need to be designed to include one such point on the electrical-symmetric 'axis' that is electrically accessible, namely where connection is possible.

[00305] This topology may be referred to as an 'unbalanced drive' topology. The approximately anti-symmetric voltage distribution of the electromagnetic mode along the inductive element may be approximately preserved, even though the resonator may not be driven exactly symmetrically. The reason is that the high *Q* and the large associated R-vs.-Z_0 mismatch necessitate that a small current may run through B and ground, compared to the much larger current that may flow inside the resonator, (A+(R,L)). In this scenario, the perturbation on the resonator mode may be weak and the location of the voltage node may stay at approximately the center location of the inductive element. These unbalanced-drive examples may have the advantage that they may be achieved using simple matching circuits and that there is no restriction on the driving voltage at the V1 terminal. In some unbalanced-drive examples, additional designs may be required to reduce common-mode signals that may appear at the ground terminal.

[00306] The directly-coupled impedance-matching circuit, generally including or consisting of parts A, B and C, as shown in Fig. 28(c), may be designed so that the wires and components of the circuit do not perturb the electric and magnetic field profiles of the electromagnetic mode of the inductive element and/or the resonator and thus preserve the high resonator *Q*. The wires and metallic components of the circuit may be oriented to be perpendicular to the electric field lines of the electromagnetic mode. The wires and components of the circuit may be placed in regions where the electric and magnetic field of the electromagnetic mode are weak.

[00307] TOPOLOGIES FOR ALLEVIATING LOW-SERIES-LOSS AND HIGH-CURRENT-RATING REQUIREMENTS ON ELEMENTS

[00308] If the matching circuit used to match a small resistance, *R*, of a low-loss inductive element to a larger characteristic impedance, Z_0, of an external circuit may be considered lossless, then $I_{Z_o}^2 Z_o = I_R^2 R \leftrightarrow I_{Z_o} / I_R = \sqrt{R / Z_o}$ and the current flowing through the terminals is much smaller than the current flowing through the inductive element. Therefore, elements connected immediately in series with the terminals (such as in directly-coupled B, C (Fig. 28(c))) may not carry high currents. Then, even if the matching circuit has lossy elements, the resistive loss present in the elements in series with the terminals may not result in a significant reduction in the high-*Q* of the resonator. That is, resistive loss in those series elements may not significantly reduce the efficiency of power transmission from Z_0 to the inductive element or vice versa. Therefore, strict requirements for low-series-loss and/or high current-ratings may not be necessary for these components. In general, such reduced requirements may lead to a wider selection of components that may be designed into the high-Q and/or high-power impedance matching and resonator topologies. These reduced requirements may be especially helpful in expanding the variety of variable and/or high voltage and/or low-parallel-loss components that may be used in these high-Q and/or high-power impedance-matching circuits.

[00309] TOPOLOGIES FOR ALLEVIATING LOW-PARALLEL-LOSS AND HIGH-VOLTAGE-RATING REQUIREMENTS ON ELEMENTS

[00310] If, as above, the matching circuit used to match a small resistance, R, of a low-loss inductive element to a larger characteristic impedance, Z_0, of an external circuit is lossless, then using the previous analysis,

$$|V_{Z_o} / V_{load}| = |I_{Z_o} Z_o / I_R (R + jX)| \approx \sqrt{R / Z_o} \cdot Z_o / X = \sqrt{Z_o / R} / \left(X / R\right),$$

and, for a low-loss (high-X/R) inductive element, the voltage across the terminals may be typically much smaller than the voltage across the inductive element. Therefore, elements connected immediately in parallel to the terminals may not need to withstand high voltages. Then, even if the matching circuit has lossy elements, the resistive loss present in the elements in parallel with the terminals may not result in a significant reduction in the high-Q of the resonator. That is, resistive loss in those parallel elements may not significantly reduce the efficiency of power transmission from Z_0 to the inductive element or vice versa. Therefore, strict requirements for low-parallel-loss and/or high voltage-ratings may not be necessary for these components. In general, such reduced requirements may lead to a wider selection of components that may be designed into the high-Q and/or high-power impedance matching and resonator topologies. These reduced requirements may be especially helpful in expanding the variety of variable and/or high current and/or low-series-loss components that may be used in these high-Q and/or high-power impedance-matching and resonator circuits.

[00311] Note that the design principles above may reduce currents and voltages on various elements differently, as they variously suggest the use of networks in series with Z_0 (such as directly-coupled B, C) or the use of networks in parallel with Z_0. The preferred topology for a given application may depend on the availability of low-series-loss/high-current-rating or low-parallel-loss/high-voltage-rating elements.

[00312] COMBINATIONS OF FIXED AND VARIABLE ELEMENTS FOR ACHIEVING FINE TUNABILITY AND ALLEVIATING HIGH-RATING REQUIREMENTS ON VARIABLE ELEMENTS

[00313] **Circuit topologies**

[00314] Variable circuit elements with satisfactory low-loss and high-voltage or current ratings may be difficult or expensive to obtain. In this disclosure, we describe impedance-matching topologies that may incorporate combinations of fixed and variable elements, such that large voltages or currents may be assigned to fixed elements in the circuit, which may be more likely to have adequate voltage and current ratings, and alleviating the voltage and current rating requirements on the variable elements in the circuit.

[00315] Variable circuit elements may have tuning ranges larger than those required by a given impedance-matching application and, in those cases, fine tuning resolution may be difficult to obtain using only such large-range elements. In this disclosure, we describe impedance-matching topologies that incorporate combinations of both fixed and variable elements, such that finer tuning resolution may be accomplished with the same variable elements.

[00316] Therefore, topologies using combinations of both fixed and variable elements may bring two kinds of advantages simultaneously: reduced voltage across, or current through, sensitive tuning components in the circuit and finer tuning resolution. Note that the maximum achievable tuning range may be related to the maximum reduction in voltage across, or current through, the tunable components in the circuit designs.

[00317] **Element topologies**

[00318] A single variable circuit-element (as opposed to the network of elements discussed above) may be implemented by a topology using a combination of fixed and variable components, connected in series or in parallel, to achieve a reduction in the rating requirements of the variable components and a finer tuning resolution. This can be demonstrated mathematically by the fact that:

$$\text{If } x_{|total|} = x_{|fixed|} + x_{|\text{var}iable|},$$
$$\text{then } \Delta x_{|total|} / x_{|total|} = \Delta x_{|\text{variable}|} / (x_{|fixed|} + x_{|\text{variable}|}),$$
$$\text{and } X_{\text{variable}} / X_{total} = X_{\text{variable}} / (X_{fixed} + X_{\text{variable}}),$$

where $x_{|subscript|}$ is any element value (e.g. capacitance, inductance), X is voltage or current, and the “+ sign” denotes the appropriate (series-addition or parallel-addition) combination of elements. Note that the subscript format for $x_{|subscript|}$, is chosen to easily distinguish it from the radius of the area enclosed by a circular inductive element (e.g. x, x_1, etc.).

[00319] Furthermore, this principle may be used to implement a variable electric element of a certain type (e.g. a capacitance or inductance) by using a variable element of a different type, if the latter is combined appropriately with other fixed elements.

[00320] In conclusion, one may apply a topology optimization algorithm that decides on the required number, placement, type and values of fixed and variable elements with the required tunable range as an optimization constraint and the minimization of the currents and/or voltages on the variable elements as the optimization objective.

[00321] EXAMPLES

[00322] In the following schematics, we show different specific topology implementations for impedance matching to and resonator designs for a low-loss inductive element. In addition, we indicate for each topology: which of the principles described above are used, the equations giving the values of the variable elements that may be used to achieve the matching, and the range of the complex impedances that may be matched (using both inequalities and a Smith-chart description). For these examples, we assume that Z_0 is real, but an extension to a characteristic impedance with a non-zero imaginary part is straightforward, as it implies only a small adjustment in the required values of the components of the matching network. We will use the convention that the subscript, n, on a quantity implies normalization to (division by) Z_0.

[00323] Fig. 29 shows two examples of a transformer-coupled impedance-matching circuit, where the two tunable elements are a capacitor and the mutual inductance between two inductive elements. If we define respectively $X_2=\omega L_2$ for Fig. 29(a) and $X_2=\omega L_2-1/\omega C_2$ for Fig. 29(b), and $X \equiv \omega L$, then the required values of the tunable elements are:

$$\omega C_1 = \frac{1}{X + RX_{2n}}$$

$$\omega M = \sqrt{Z_o R(1+X_{2n}^2)}\,.$$

For the topology of Fig. 29(b), an especially straightforward design may be to choose X_2=0. In that case, these topologies may match the impedances satisfying the inequalities:

$$R_n > 0,\ X_n > 0,$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 29(c).

[00324] Given a well pre-chosen fixed M, one can also use the above matching topologies with a tunable C_2 instead.

[00325] Fig. 30 shows six examples (a)-(f) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and six examples (h)-(m) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 30(a),(b),(c),(h),(i),(j), a common-mode signal may be required at the two terminals to preserve the voltage node of the resonator at the center of the inductive element and thus the high *Q*. Note that these examples may be described as implementations of the general topology shown in Fig. 28(c). For the symmetric topologies of Figs. 30(d),(e),(f),(k),(l),(m), the two terminals may need to be driven anti-symmetrically (balanced drive) to preserve the voltage node of the resonator at the center of the inductive element and thus the high *Q*. Note that these examples may be described as implementations of the general topology shown in Fig. 28(d). It will be appreciated that a network of capacitors, as used herein, may in general refer to any circuit topology including one or more capacitors, including without limitation any of the circuits specifically disclosed herein using capacitors, or any other equivalent or different circuit structure(s), unless another meaning is explicitly provided or otherwise clear from the context.

[00326] Let us define respectively $Z=R+j\omega L$ for Figs. 30(a),(d),(h),(k), $Z=R+j\omega L+1/j\omega C_3$ for Figs. 30(b),(e),(i),(l), and $Z=(R+j\omega L)||(1/j\omega C_3)$ for Figs. 30(c),(f),(j),(m), where the symbol "||" means "the parallel combination of", and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.30(a)-(f) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X - \sqrt{X^2 R_n - R^2(1-R_n)}}{X^2 + R^2},$$

$$\omega C_2 = \frac{R_n \omega C_1}{1 - X\omega C_1 - R_n},$$

and these topologies can match the impedances satisfying the inequalities:

$$R_n \leq 1,\ X_n \geq \sqrt{R_n(1-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 30(g).
For Figs.30(h)-(m) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X + \sqrt{X^2 R_n - R^2(1-R_n)}}{X^2 + R^2},$$
$$\omega L_2 = -\frac{1 - X\omega C_1 - R_n}{R_n \omega C_1}.$$

[00327] Fig. 31 shows three examples (a)-(c) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and three examples (e)-(g) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 31(a),(b),(c),(e),(f),(g), the ground terminal is connected between two equal-value capacitors, $2C_1$, (namely on the axis of symmetry of the main resonator) to preserve the voltage node of the resonator at the center of the inductive element and thus the high *Q*. Note that these examples may be described as implementations of the general topology shown in Fig. 28(e).

[00328] Let us define respectively Z=R+jωL for Figs. 31(a),(e), $Z=R+j\omega L+1/j\omega C_3$ for Figs. 31(b),(f), and $Z=(R+j\omega L)||(1/j\omega C_3)$ for Fig. 31(c),(g), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.31(a)-(c) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X - \frac{1}{2}\sqrt{X^2 R_n - R^2(4-R_n)}}{X^2 + R^2},$$

$$\omega C_2 = \frac{R_n \omega C_1}{1 - X\omega C_1 - \frac{R_n}{2}},$$

and these topologies can match the impedances satisfying the inequalities:

$$R_n \le 1, \quad X_n \ge \sqrt{\frac{R_n}{1-R_n}(2-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 31(d). For Figs.31(e)-(g) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X + \frac{1}{2}\sqrt{X^2 R_n - R^2(4-R_n)}}{X^2 + R^2},$$

$$\omega L_2 = -\frac{1 - X\omega C_1 - \frac{R_n}{2}}{R_n \omega C_1}.$$

[00329] Fig. 32 shows three examples (a)-(c) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and three examples (e)-(g) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 32(a),(b),(c),(e),(f),(g), the ground terminal may be connected at the center of the inductive element to preserve the voltage node of the resonator at that point and thus the high *Q*. Note that these example may be described as implementations of the general topology shown in Fig. 28(e).

[00330] Let us define respectively $Z=R+j\omega L$ for Fig. 32(a), $Z=R+j\omega L+1/j\omega C_3$ for Fig. 32(b), and $Z=(R+j\omega L)||(1/j\omega C_3)$ for Fig. 32(c), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.32(a)-(c) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X - \sqrt{\frac{X^2 R_n - 2R^2(2-R_n)}{4-R_n}}}{X^2 + R^2},$$

$$\omega C_2 = \frac{R_n \omega C_1}{1 - X\omega C_1 - \frac{R_n}{2} + \frac{R_n X \omega C_1}{2(1+k)}},$$

where k is defined by $M' = -kL'$, where L' is the inductance of each half of the inductor loop and M' is the mutual inductance between the two halves, and these topologies can match the impedances satisfying the inequalities:

$$R_n \le 2, \quad X_n \ge \sqrt{2R_n(2-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 32(d). For Figs.32(e)-(g) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X + \sqrt{\dfrac{X^2 R_n - 2R^2(2-R_n)}{4-R_n}}}{X^2+R^2},$$

[00331] In the circuits of Figs. 30, 31, 32, the capacitor, C_2, or the inductor, L_2, is (or the two capacitors, $2C_2$, or the two inductors, $L_2/2$, are) in series with the terminals and may not need to have very low series-loss or withstand a large current.

[00332] Fig. 33 shows six examples (a)-(f) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and six examples (h)-(m) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 33(a),(b),(c),(h),(i),(j), a common-mode signal may be required at the two terminals to preserve the voltage node of the resonator at the center of the inductive element and thus the high *Q*. Note that these examples may be described as implementations of the general topology shown in Fig. 28(c), where B and C are short-circuits and A is not balanced. For the symmetric topologies of Figs. 33(d),(e),(f),(k),(l),(m), the two terminals may need to be driven anti-symmetrically (balanced drive) to preserve the voltage node of the resonator at the center of the inductive element and thus the high *Q*. Note that these examples may be described as implementations of the general topology shown in Fig. 28(d), where B and C are short-circuits and A is balanced.

[00333] Let us define respectively $Z=R+j\omega L$ for Figs. 33(a),(d),(h),(k), $Z=R+j\omega L+1/j\omega C_3$ for Figs. 33(b),(e),(i),(l), and $Z=(R+j\omega L)\|(1/j\omega C_3)$ for Figs. 33(c),(f),(j),(m), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.33(a)-(f) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{1}{X - Z_o\sqrt{R_n(1-R_n)}},$$

$$\omega C_2 = \frac{1}{Z_o}\sqrt{\frac{1}{R_n} - 1},$$

and these topologies can match the impedances satisfying the inequalities:

$$R_n \le 1, \quad X_n \ge \sqrt{R_n(1-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 33(g). For Figs.35(h)-(m) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{1}{X + Z_o\sqrt{R_n(1-R_n)}},$$

$$\omega L_2 = \frac{Z_o}{\sqrt{\frac{1}{R_n} - 1}}.$$

[00334] Fig. 34 shows three examples (a)-(c) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and three examples (e)-(g) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 34(a),(b),(c),(e),(f),(g), the ground terminal is connected between two equal-value capacitors, $2C_2$, (namely on the axis of symmetry of the main resonator) to preserve the voltage node of the resonator at the center of the inductive element and thus the high Q. Note that these examples may be described as implementations of the general topology shown in Fig. 28(e).

[00335] Let us define respectively $Z=R+j\omega L$ for Fig. 34(a),(e), $Z=R+j\omega L+1/j\omega C_3$ for Fig. 34(b),(f), and $Z=(R+j\omega L)\|(1/j\omega C_3)$ for Fig. 34(c),(g), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.34(a)-(c) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{1}{X - Z_o\sqrt{\frac{1-R_n}{R_n}(2-R_n)}},$$

$$\omega C_2 = \frac{1}{2Z_o}\sqrt{\frac{1}{R_n}-1},$$

and these topologies can match the impedances satisfying the inequalities:

$$R_n \le 1,\;\; X_n \ge \sqrt{\frac{R_n}{1-R_n}(2-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 34(d).
For Figs.34(e)-(g) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{1}{X + Z_o\sqrt{\frac{1-R_n}{R_n}(2-R_n)}},$$

$$\omega L_2 = \frac{2Z_o}{\sqrt{\frac{1}{R_n}-1}}.$$

[00336] Fig. 35 shows three examples of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors. For the topologies of Figs. 35, the ground terminal may be connected at the center of the inductive element to preserve the voltage node of the resonator at that point and thus the high Q. Note that these examples may be described as implementations of the general topology shown in Fig. 28(e).

[00337] Let us define respectively Z=R+jωL for Fig. 35(a), Z=R+jωL+1/jωC_3 for Fig. 35(b), and Z=(R+jωL)||(1/jωC_3) for Fig. 35(c), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{2}{X(1+a) - \sqrt{Z_o R(4-R_n)(1+a^2)}},$$
$$\omega C_2 = \frac{2}{X(1+a) + \sqrt{Z_o R(4-R_n)(1+a^2)}},$$

where $a = \frac{R}{2Z_o - R} \cdot \frac{k}{1+k}$ and k is defined by $M' = -kL'$, where L' is the inductance of each half of the inductive element and M' is the mutual inductance between the two halves. These topologies can match the impedances satisfying the inequalities:

$$R_n \leq 2 \,\&\, \frac{2}{\gamma} \leq R_n \leq 4,$$
$$X_n \geq \sqrt{\frac{R_n(4-R_n)(2-R_n)}{2-\gamma R_n}},$$

where

$$\gamma = \frac{1-6k+k^2}{1+2k+k^2} \leq 1$$

which are shown by the area enclosed by the bold lines on the three Smith charts shown in Fig. 35(d) for k=0, Fig. 35(e) for k=0.05, and Fig. 35(f) for k=1. Note that for 0<k<1 there are two disconnected regions of the Smith chart that this topology can match.

[00338] In the circuits of Figs. 33, 34, 35, the capacitor, C_2, or the inductor, L_2, is (or one of the two capacitors, $2C_2$, or one of the two inductors, $2L_2$, are) in parallel with the terminals and thus may not need to have a high voltage-rating. In the case of two capacitors, $2C_2$, or two inductors, $2L_2$, both may not need to have a high voltage-rating, since approximately the same current flows through them and thus they experience approximately the same voltage across them.

[00339] For the topologies of Figs. 30-35, where a capacitor, C_3, is used, the use of the capacitor, C_3, may lead to finer tuning of the frequency and the impedance. For the topologies of Figs. 30-35, the use of the fixed capacitor, C_3, in series with the inductive element may ensure that a large percentage of the high inductive-element voltage will be across this fixed capacitor, C_3, thus potentially alleviating the voltage rating requirements for the other elements of the impedance matching circuit, some of which may be variable. Whether or not such topologies are preferred depends on the availability, cost and specifications of appropriate fixed and tunable components.

[00340] In all the above examples, a pair of equal-value variable capacitors without a common terminal may be implemented using ganged-type capacitors or groups or arrays of varactors or diodes biased and controlled to tune their values as an ensemble. A pair of equal-value variable capacitors with one common terminal can be implemented using a tunable butterfly-type capacitor or any other tunable or variable capacitor or group or array of varactors or diodes biased and controlled to tune their capacitance values as an ensemble.

[00341] Another criterion which may be considered upon the choice of the impedance matching network is the response of the network to different frequencies than the desired operating frequency. The signals generated in the external circuit, to which the inductive element is coupled, may not be monochromatic at the desired frequency but periodic with the desired frequency, as for example the driving signal of a switching amplifier or the reflected signal of a switching rectifier. In some such cases, it may be desirable to suppress the amount of higher-order harmonics that enter the inductive element (for example, to reduce radiation of these harmonics from this element). Then the choice of impedance matching network may be one that sufficiently suppresses the amount of such harmonics that enters the inductive element.

[00342] The impedance matching network may be such that the impedance seen by the external circuit at frequencies higher than the fundamental harmonic is high, when the external periodic signal is a signal that can be considered to behave as a voltage-source signal (such as the driving signal of a class-D amplifier with a series resonant load), so that little current flows through the inductive element at higher frequencies. Among the topologies of Figs. 30-35, those which use an inductor, L_2, may then be preferable, as this inductor presents a high impedance at high frequencies.

[00343] The impedance matching network may be such that the impedance seen by the external circuit at frequencies higher than the fundamental harmonic is low, when the external periodic signal is a signal that can be considered to behave as a current-source signal, so that little voltage is induced across the inductive element at higher frequencies. Among the topologies of Figs. 30-35, those which use a capacitor, C_2, are then preferable, as this capacitor presents a low impedance at high frequencies.

[00344] Fig. 36 shows four examples of a variable capacitance, using networks of one variable capacitor and the rest fixed capacitors. Using these network topologies, fine tunability of the total capacitance value may be achieved. Furthermore, the topologies of Figs. 36(a),(c),(d), may be used to reduce the voltage across the variable capacitor, since most of the voltage may be assigned across the fixed capacitors.

[00345] Fig. 37 shows two examples of a variable capacitance, using networks of one variable inductor and fixed capacitors. In particular, these networks may provide implementations for a variable reactance, and, at the frequency of interest, values for the variable inductor may be used such that each network corresponds to a net negative variable reactance, which may be effectively a variable capacitance.

[00346] Tunable elements such as tunable capacitors and tunable inductors may be mechanically-tunable, electrically-tunable, thermally-tunable and the like. The tunable elements may be variable capacitors or inductors, varactors, diodes, Schottky diodes, reverse-biased PN diodes, varactor arrays, diode arrays, Schottky diode arrays and the like. The diodes may be Si diodes, GaN diodes, SiC diodes, and the like. GaN and SiC diodes may be particularly attractive for high power applications. The tunable elements may be electrically switched capacitor banks, electrically-switched mechanically-tunable capacitor banks, electrically-switched varactor-array banks, electrically-switched transformer-coupled inductor banks, and the like. The tunable elements may be combinations of the elements listed above.

[00347] As described above, the efficiency of the power transmission between coupled high-Q magnetic resonators may be impacted by how closely matched the resonators are in resonant frequency and how well their impedances are matched to the power supplies and power consumers in the system. Because a variety of external factors including the relative position of extraneous objects or other resonators in the system, or the changing of those relative positions, may alter the resonant frequency and/or input impedance of a high-Q magnetic resonator, tunable impedance networks may be required to maintain sufficient levels of power transmission in various environments or operating scenarios.

[00348] The capacitance values of the capacitors shown may be adjusted to adjust the resonant frequency and/or the impedance of the magnetic resonator. The capacitors may be adjusted electrically, mechanically, thermally, or by any other known methods. They may be adjusted manually or automatically, such as in response to a feedback signal. They may be adjusted to achieve certain power transmission efficiencies or other operating characteristics between the power supply and the power consumer.

[00349] The inductance values of the inductors and inductive elements in the resonator may be adjusted to adjust the frequency and/or impedance of the magnetic resonator. The inductance may be adjusted using coupled circuits that include adjustable components such as tunable capacitors, inductors and switches. The inductance may be adjusted using transformer coupled tuning circuits. The inductance may be adjusted by switching in and out different sections of conductor in the inductive elements and/or using ferro-magnetic tuning and/or mu-tuning, and the like.

[00350] The resonant frequency of the resonators may be adjusted to or may be allowed to change to lower or higher frequencies. The input impedance of the resonator may be adjusted to or may be allowed to change to lower or higher impedance values. The amount of power delivered by the source and/or received by the devices may be adjusted to or may be allowed to change to lower or higher levels of power. The amount of power delivered to the source and/or received by the devices from the device resonator may be adjusted to or may be allowed to change to lower or higher levels of power. The resonator input impedances, resonant frequencies, and power levels may be adjusted depending on the power consumer or consumers in the system and depending on the objects or materials in the vicinity of the resonators. The resonator input impedances, frequencies, and power levels may be adjusted manually or automatically, and may be adjusted in response to feedback or control signals or algorithms.

[00351] Circuit elements may be connected directly to the resonator, that is, by physical electrical contact, for example to the ends of the conductor that forms the inductive element and/or the terminal connectors. The circuit elements may be soldered to, welded to, crimped to, glued to, pinched to, or closely position to the conductor or attached using a variety of electrical components, connectors or connection techniques. The power supplies and the power consumers may be connected to magnetic resonators directly or indirectly or inductively. Electrical signals may be supplied to, or taken from, the resonators through the terminal connections.

[00352] It is to be understood by one of ordinary skill in the art that in real implementations of the principles described herein, there may be an associated tolerance, or acceptable variation, to the values of real components (capacitors, inductors, resistors and the like) from the values calculated via the herein stated equations, to the values of real signals (voltages, currents and the like) from the values suggested by symmetry or anti-symmetry or otherwise, and to the values of real geometric locations of points (such as the point of connection of the ground terminal close to the center of the inductive element or the 'axis' points and the like) from the locations suggested by symmetry or otherwise.

[00353] **<u>Examples</u>**

[00354] <u>SYSTEM BLOCK DIAGRAMS</u>

[00355] We disclose examples of high-Q resonators for wireless power transmission systems that may wirelessly power or charge devices at mid-range distances. High-*Q* resonator wireless power transmission systems also may wirelessly power or charge devices with magnetic resonators that are different in size, shape, composition, arrangement, and the like, from any source resonators in the system.

[00356] Fig. 1(a)(b) shows high level diagrams of two exemplary two-resonator systems. These exemplary systems each have a single source resonator 102S or 104S and a single device resonator 102D or 104D. Fig. 38 shows a high level block diagram of a system with a few more features highlighted. The wirelessly powered or charged device 2310 may include or consist of a device resonator 102D, device power and control circuitry 2304, and the like, along with the device 2308 or devices, to which either DC or AC or both AC and DC power is transferred. The energy or power source for a system may include the source power and control circuitry 2302, a source resonator 102S, and the like. The device 2308 or devices that receive power from the device resonator 102D and power and control circuitry 2304 may be any kind of device 2308 or devices as described previously. The device resonator 102D and circuitry 2304 delivers power to the device/devices 2308 that may be used to recharge the battery of the device/devices, power the device/devices directly, or both when in the vicinity of the source resonator 102S.

[00357] The source and device resonators may be separated by many meters or they may be very close to each other or they may be separated by any distance in between. The source and device resonators may be offset from each other laterally or axially. The source and device resonators may be directly aligned (no lateral offset), or they may be offset by meters, or anything in between. The source and device resonators may be oriented so that the surface areas enclosed by their inductive elements are approximately parallel to each other. The source and device resonators may be oriented so that the surface areas enclosed by their inductive elements are approximately perpendicular to each other, or they may be oriented for any relative angle (0 to 360 degrees) between them.

[00358] The source and device resonators may be free standing or they may be enclosed in an enclosure, container, sleeve or housing. These various enclosures may be composed of almost any kind of material. Low loss tangent materials such as Teflon, REXOLITE, styrene, and the like may be preferable for some applications. The source and device resonators may be integrated in the power supplies and power consumers. For example, the source and device resonators may be integrated into keyboards, computer mice, displays, cell phones, etc. so that they are not visible outside these devices. The source and device resonators may be separate from the power supplies and power consumers in the system and may be connected by a standard or custom wires, cables, connectors or plugs.

[00359] The source 102S may be powered from a number of DC or AC voltage, current or power sources including a USB port of a computer. The source 102S may be powered from the electric grid, from a wall plug, from a battery, from a power supply, from an engine, from a solar cell, from a generator, from another source resonator, and the like. The source power and control circuitry 2302 may include circuits and components to isolate the source electronics from the power source, so that any reflected power or signals are not coupled out through the source input terminals. The source power and control circuits 2302 may include power factor correction circuits and may be configured to monitor power usage for monitoring accounting, billing, control, and like functionalities.

[00360] The system may be operated bi-directionally. That is, energy or power that is generated or stored in a device resonator may be fed back to a power source including the electric grid, a battery, any kind of energy storage unit, and the like. The source power and control circuits may include power factor correction circuits and may be configured to monitor power usage for monitoring accounting, billing, control, and like functionalities for bi-directional energy flow. Wireless energy transfer systems may enable or promote vehicle-to-grid (V2G) applications.

[00361] The source and the device may have tuning capabilities that allow adjustment of operating points to compensate for changing environmental conditions, perturbations, and loading conditions that can affect the operation of the source and device resonators and the efficiency of the energy exchange. The tuning capability may also be used to multiplex power delivery to multiple devices, from multiple sources, to multiple systems, to multiple repeaters or relays, and the like. The tuning capability may be manually controlled, or automatically controlled and may be performed continuously, periodically, intermittently or at scheduled times or intervals.

[00362] The device resonator and the device power and control circuitry may be integrated into any portion of the device, such as a battery compartment, or a device cover or sleeve, or on a mother board, for example, and may be integrated alongside standard rechargeable batteries or other energy storage units. The device resonator may include a device field reshaper which may shield any combination of the device resonator elements and the device power and control electronics from the electromagnetic fields used for the power transfer and which may deflect the resonator fields away from the lossy device resonator elements as well as the device power and control electronics. A magnetic material and/or high-conductivity field reshaper may be used to increase the perturbed quality factor Q of the resonator and increase the perturbed coupling factor of the source and device resonators.

[00363] The source resonator and the source power and control circuitry may be integrated into any type of furniture, structure, mat, rug, picture frame (including digital picture frames, electronic frames), plug-in modules, electronic devices, vehicles, and the like. The source resonator may include a source field reshaper which may shield any combination of the source resonator elements and the source power and control electronics from the electromagnetic fields used for the power transfer and which may deflect the resonator fields away from the lossy source resonator elements as well as the source power and control electronics. A magnetic material and/or high-conductivity field reshaper may be used to increase the perturbed quality factor Q of the resonator and increase the perturbed coupling factor of the source and device resonators.

[00364] A block diagram of the subsystems in an example of a wirelessly powered device is shown in Fig. 39. The power and control circuitry may be designed to transform the alternating current power from the device resonator 102D and convert it to stable direct current power suitable for powering or charging a device. The power and control circuitry may be designed to transform an alternating current power at one frequency from the device resonator to alternating current power at a different frequency suitable for powering or charging a device. The power and control circuitry may include or consist of impedance matching circuitry 2402D, rectification circuitry 2404, voltage limiting circuitry (not shown), current limiting circuitry (not shown), AC-to-DC converter 2408 circuitry, DC-to-DC converter 2408 circuitry, DC-to-AC converter 2408 circuitry, AC-to-AC converter 2408 circuitry, battery charge control circuitry (not shown), and the like.

[00365] The impedance-matching 2402D network may be designed to maximize the power delivered between the device resonator 102D and the device power and control circuitry 2304 at the desired frequency. The impedance matching elements may be chosen and connected such that the high-Q of the resonators is preserved. Depending on the operating conditions, the impedance matching circuitry 2402D may be varied or tuned to control the power delivered from the source to the device, from the source to the device resonator, between the device resonator and the device power and control circuitry, and the like. The power, current and voltage signals may be monitored at any point in the device circuitry and feedback algorithms circuits, and techniques, may be used to control components to achieve desired signal levels and system operation. The feedback algorithms may be implemented using analog or digital circuit techniques and the circuits may include a microprocessor, a digital signal processor, a field programmable gate array processor and the like.

[00366] The third block of Fig. 39 shows a rectifier circuit 2404 that may rectify the AC voltage power from the device resonator into a DC voltage. In this configuration, the output of the rectifier 2404 may be the input to a voltage clamp circuit. The voltage clamp circuit (not shown) may limit the maximum voltage at the input to the DC-to-DC converter 2408D or DC-to-AC converter 2408D. In general, it may be desirable to use a DC-to-DC/AC converter with a large input voltage dynamic range so that large variations in device position and operation may be tolerated while adequate power is delivered to the device. For example, the voltage level at the output of the rectifier may fluctuate and reach high levels as the power input and load characteristics of the device change. As the device performs different tasks it may have varying power demands. The changing power demands can cause high voltages at the output of the rectifier as the load characteristics change. Likewise as the device and the device resonator are brought closer and further away from the source, the power delivered to the device resonator may vary and cause changes in the voltage levels at the output of the rectifier. A voltage clamp circuit may prevent the voltage output from the rectifier circuit from exceeding a predetermined value which is within the operating range of the DC-to-DC/AC converter. The voltage clamp circuitry may be used to extend the operating modes and ranges of a wireless energy transfer system.

[00367] The next block of the power and control circuitry of the device is the DC-to-DC converter 2408D that may produce a stable DC output voltage. The DC-to-DC converter may be a boost converter, buck converter, boost-buck converter, single ended primary inductance converter (SEPIC), or any other DC-DC topology that fits the requirements of the particular application. If the device requires AC power, a DC-to-AC converter may be substituted for the DC-to-DC converter, or the DC-to-DC converter may be followed by a DC-to-AC converter. If the device contains a rechargeable battery, the final block of the device power and control circuitry may be a battery charge control unit which may manage the charging and maintenance of the battery in battery powered devices.

[00368] The device power and control circuitry 2304 may contain a processor 2410D, such as a microcontroller, a digital signal processor, a field programmable gate array processor, a microprocessor, or any other type of processor. The processor may be used to read or detect the state or the operating point of the power and control circuitry and the device resonator. The processor may implement algorithms to interpret and adjust the operating point of the circuits, elements, components, subsystems and resonator. The processor may be used to adjust the impedance matching, the resonator, the DC to DC converters, the DC to AC converters, the battery charging unit, the rectifier, and the like of the wirelessly powered device.

[00369] The processor may have wireless or wired data communication links to other devices or sources and may transmit or receive data that can be used to adjust the operating point of the system. Any combination of power, voltage, and current signals at a single, or over a range of frequencies, may be monitored at any point in the device circuitry. These signals may be monitored using analog or digital or combined analog and digital techniques. These monitored signals may be used in feedback loops or may be reported to the user in a variety of known ways or they may be stored and retrieved at later times. These signals may be used to alert a user of system failures, to indicate performance, or to provide audio, visual, vibrational, and the like, feedback to a user of the system.

[00370] Fig. 40 shows components of source power and control circuitry 2302 of an exemplary wireless power transfer system configured to supply power to a single or multiple devices. The source power and control circuitry 2302 of the exemplary system may be powered from an AC voltage source 2502 such as a home electrical outlet, a DC voltage source such as a battery, a USB port of a computer, a solar cell, another wireless power source, and the like. The source power and control circuitry 2302 may drive the source resonator 102S with alternating current, such as with a frequency greater than 10 kHz and less than 100 MHz. The source power and control circuitry 2302 may drive the source resonator 102S with alternating current of frequency less than less than 10 GHz. The source power and control circuitry 2302 may include a DC-to-DC converter 2408S, an AC-to-DC converter 2408S, or both an AC-to-DC converter 2408S and a DC-to-DC 2408S converter, an oscillator 2508, a power amplifier 2504, an impedance matching network 2402S, and the like.

[00371] The source power and control circuitry 2302 may be powered from multiple AC-or-DC voltage sources 2502 and may contain AC-to-DC and DC-to-DC converters 2408S to provide necessary voltage levels for the circuit components as well as DC voltages for the power amplifiers that may be used to drive the source resonator. The DC voltages may be adjustable and may be used to control the output power level of the power amplifier. The source may contain power factor correction circuitry.

[00372] The oscillator 2508 output may be used as the input to a power amplifier 2504 that drives the source resonator 102S. The oscillator frequency may be tunable and the amplitude of the oscillator signal may be varied as one means to control the output power level from the power amplifier. The frequency, amplitude, phase, waveform, and duty cycle of the oscillator signal may be controlled by analog circuitry, by digital circuitry or by a combination of analog and digital circuitry. The control circuitry may include a processor 2410S, such as a microprocessor, a digital signal processor, a field programmable gate array processor, and the like.

[00373] The impedance matching blocks 2402 of the source and device resonators may be used to tune the power and control circuits and the source and device resonators. For example, tuning of these circuits may adjust for perturbation of the quality factor Q of the source or device resonators due to extraneous objects or changes in distance between the source and device in a system. Tuning of these circuits may also be used to sense the operating environment, control power flow to one or more devices, to control power to a wireless power network, to reduce power when unsafe or failure mode conditions are detected, and the like.

[00374] Any combination of power, voltage, and current signals may be monitored at any point in the source circuitry. These signals may be monitored using analog or digital or combined analog and digital techniques. These monitored signals may be used in feedback circuits or may be reported to the user in a variety of known ways or they may be stored and retrieved at later times. These signals may be used to alert a user to system failures, to alert a user to exceeded safety thresholds, to indicate performance, or to provide audio, visual, vibrational, and the like, feedback to a user of the system.

[00375] The source power and control circuitry may contain a processor. The processor may be used to read the state or the operating point of the power and control circuitry and the source resonator. The processor may implement algorithms to interpret and adjust the operating point of the circuits, elements, components, subsystems and resonator. The processor may be used to adjust the impedance matching, the resonator, the DC-to-DC converters, the AC-to-DC converters, the oscillator, the power amplifier of the source, and the like. The processor and adjustable components of the system may be used to implement frequency and/or time power delivery multiplexing schemes. The processor may have wireless or wired data communication links to devices and other sources and may transmit or receive data that can be used to adjust the operating point of the system.

[00376] Although detailed and specific designs are shown in these block diagrams, it should be clear to those skilled in the art that many different modifications and rearrangements of the components and building blocks are possible within the spirit of the exemplary system. The division of the circuitry was outlined for illustrative purposes and it should be clear to those skilled in the art that the components of each block may be further divided into smaller blocks or merged or shared. In equivalent examples the power and control circuitry may be composed of individual discrete components or larger integrated circuits. For example, the rectifier circuitry may be composed of discrete diodes, or use diodes integrated on a single chip. A multitude of other circuits and integrated devices can be substituted in the design depending on design criteria such as power or size or cost or application. The whole of the power and control circuitry or any portion of the source or device circuitry may be integrated into one chip.

[00377] The impedance matching network of the device and or source may include a capacitor or networks of capacitors, an inductor or networks of inductors, or any combination of capacitors, inductors, diodes, switches, resistors, and the like. The components of the impedance matching network may be adjustable and variable and may be controlled to affect the efficiency and operating point of the system. The impedance matching may be performed by controlling the connection point of the resonator, adjusting the permeability of a magnetic material, controlling a bias field, adjusting the frequency of excitation, and the like. The impedance matching may use or include any number or combination of varactors, varactor arrays, switched elements, capacitor banks, switched and tunable elements, reverse bias diodes, air gap capacitors, compression capacitors, BZT electrically tuned capacitors, MEMS-tunable capacitors, voltage variable dielectrics, transformer coupled tuning circuits, and the like. The variable components may be mechanically tuned, thermally tuned, electrically tuned, piezo-electrically tuned, and the like. Elements of the impedance matching may be silicon devices, gallium nitride devices, silicon carbide devices and the like. The elements may be chosen to withstand high currents, high voltages, high powers, or any combination of current, voltage and power. The elements may be chosen to be high-Q elements.

[00378] The matching and tuning calculations of the source may be performed on an external device through a USB port that powers the device. The device may be a computer a PDA or other computational platform.

[00379] A demonstration system used a source resonator, coupled to a device resonator, to wirelessly power/recharge multiple electronic consumer devices including, but not limited to, a laptop, a DVD player, a projector, a cell-phone, a display, a television, a projector, a digital picture frame, a light, a TV/DVD player, a portable music player, a circuit breaker, a hand-held tool, a personal digital assistant, an external battery charger, a mouse, a keyboard, a camera, an active load, and the like. A variety of devices may be powered simultaneously from a single device resonator. Device resonators may be operated simultaneously as source resonators. The power supplied to a device resonator may pass through additional resonators before being delivered to its intended device resonator.

[00380] *Monitoring, Feedback and Control*

[00381] So-called port parameter measurement circuitry may measure or monitor certain power, voltage, and current, signals in the system and processors or control circuits may adjust certain settings or operating parameters based on those measurements. In addition to these port parameter measurements, the magnitude and phase of voltage and current signals, and the magnitude of the power signals, throughout the system may be accessed to measure or monitor the system performance. The measured signals referred to throughout this disclosure may be any combination of the port parameter signals, as well as voltage signals, current signals, power signals, and the like. These parameters may be measured using analog or digital signals, they may be sampled and processed, and they may be digitized or converted using a number of known analog and digital processing techniques. Measured or monitored signals may be used in feedback circuits or systems to control the operation of the resonators and/or the system. In general, we refer to these monitored or measured signals as reference signals, or port parameter measurements or signals, although they are sometimes also referred to as error signals, monitor signals, feedback signals, and the like. We will refer to the signals that are used to control circuit elements such as the voltages used to drive voltage controlled capacitors as the control signals.

[00382] In some cases the circuit elements may be adjusted to achieve a specified or predetermined impedance value for the source and device resonators. In other cases the impedance may be adjusted to achieve a desired impedance value for the source and device resonators when the device resonator is connected to a power consumer or consumers. In other cases the impedance may be adjusted to mitigate changes in the resonant frequency, or impedance or power level changes owing to movement of the source and/or device resonators, or changes in the environment (such as the movement of interacting materials or objects) in the vicinity of the resonators. In other cases the impedance of the source and device resonators may be adjusted to different impedance values.

[00383] The coupled resonators may be made of different materials and may include different circuits, components and structural designs or they may be the same. The coupled resonators may include performance monitoring and measurement circuitry, signal processing and control circuitry or a combination of measurement and control circuitry. Some or all of the high-Q magnetic resonators may include tunable impedance circuits. Some or all of the high-Q magnetic resonators may include automatically controlled tunable impedance circuits.

[00384] Fig. 41 shows a magnetic resonator with port parameter measurement circuitry 3802 configured to measure certain parameters of the resonator. The port parameter measurement circuitry may measure the input impedance of the structure, or the reflected power. Port parameter measurement circuits may be included in the source and/or device resonator designs and may be used to measure two port circuit parameters such as S-parameters (scattering parameters), Z-parameters (impedance parameters), Y-parameters (admittance parameters), T-parameters (transmission parameters), H-parameters (hybrid parameters), ABCD-parameters (chain, cascade or transmission parameters), and the like. These parameters may be used to describe the electrical behavior of linear electrical networks when various types of signals are applied.

[00385] Different parameters may be used to characterize the electrical network under different operating or coupling scenarios. For example, S-parameters may be used to measure matched and unmatched loads. In addition, the magnitude and phase of voltage and current signals within the magnetic resonators and/or within the sources and devices themselves may be monitored at a variety of points to yield system performance information. This information may be presented to users of the system via a user interface such as a light, a read-out, a beep, a noise, a vibration or the like, or it may be presented as a digital signal or it may be provided to a processor in the system and used in the automatic control of the system. This information may be logged, stored, or may be used by higher level monitoring and control systems.

[00386] Fig. 42 shows a circuit diagram of a magnetic resonator where the tunable impedance network may be realized with voltage controlled capacitors 3902 or capacitor networks. Such an implementation may be adjusted, tuned or controlled by electrical circuits and/or computer processors, such as a programmable voltage source 3908, and the like. For example, the voltage controlled capacitors may be adjusted in response to data acquired by the port parameter measurement circuitry 3802 and processed by a measurement analysis and control algorithm subsystem 3904. Reference signals may be derived from the port parameter measurement circuitry or other monitoring circuitry designed to measure the degree of deviation from a desired system operating point. The measured reference signals may include voltage, current, complex-impedance, reflection coefficient, power levels and the like, at one or several points in the system and at a single frequency or at multiple frequencies.

[00387] The reference signals may be fed to measurement analysis and control algorithm subsystem modules that may generate control signals to change the values of various components in a tunable impedance matching network. The control signals may vary the resonant frequency and/or the input impedance of the magnetic resonator, or the power level supplied by the source, or the power level drawn by the device, to achieve the desired power exchange between power supplies/generators and power drains/loads.

[00388] Adjustment algorithms may be used to adjust the frequency and/or impedance of the magnetic resonators. The algorithms may take in reference signals related to the degree of deviation from a desired operating point for the system and output correction or control signals related to that deviation that control variable or tunable elements of the system to bring the system back towards the desired operating point or points. The reference signals for the magnetic resonators may be acquired while the resonators are exchanging power in a wireless power transmission system, or they may be switched out of the circuit during system operation. Corrections to the system may be applied or performed continuously, periodically, upon a threshold crossing, digitally, using analog methods, and the like.

[00389] Fig. 43 shows an end-to-end wireless power transmission system. Both the source and the device may include port measurement circuitry 3802 and a processor 2410. The box labeled "coupler/switch" 4002 indicates that the port measurement circuitry 3802 may be connected to the resonator 102 by a directional coupler or a switch, enabling the measurement, adjustment and control of the source and device resonators to take place in conjunction with, or separate from, the power transfer functionality.

[00390] The port parameter measurement and/or processing circuitry may reside with some, any, or all resonators in a system. The port parameter measurement circuitry may utilize portions of the power transmission signal or may utilize excitation signals over a range of frequencies to measure the source/device resonator response (i.e. transmission and reflection between any two ports in the system), and may contain amplitude and/or phase information. Such measurements may be achieved with a swept single frequency signal or a multi-frequency signal. The signals used to measure and monitor the resonators and the wireless power transmission system may be generated by a processor or processors and standard input/output (I/O) circuitry including digital to analog converters (DACs), analog to digital converters (ADCs), amplifiers, signal generation chips, passive components and the like. Measurements may be achieved using test equipment such as a network analyzer or using customized circuitry. The measured reference signals may be digitized by ADCs and processed using customized algorithms running on a computer, a microprocessor, a DSP chip, an ASIC, and the like. The measured reference signals may be processed in an analog control loop.

[00391] The measurement circuitry may measure any set of two port parameters such as S-parameters, Y-parameters, Z-parameters, H-parameters, G-parameters, T-parameters, ABCD-parameters, and the like. Measurement circuitry may be used to characterize current and voltage signals at various points in the drive and resonator circuitry, the impedance and/or admittance of the source and device resonators at opposite ends of the system, i.e. looking into the source resonator matching network ("port 1" in Fig. 43) towards the device and vice versa.

[00392] The device may measure relevant signals and/or port parameters, interpret the measurement data, and adjust its matching network to optimize the impedance looking into the coupled system independently of the actions of the source. The source may measure relevant port parameters, interpret the measurement data, and adjust its matching network to optimize the impedance looking into the coupled system independently of the actions of the device.

[00393] Fig. 43 shows a block diagram of a source and device in a wireless power transmission system. The system may be configured to execute a control algorithm that actively adjusts the tuning/matching networks in either of or both the source and device resonators to optimize performance in the coupled system. Port measurement circuitry 3802S may measure signals in the source and communicate those signals to a processor 2410. A processor 2410 may use the measured signals in a performance optimization or stabilization algorithm and generate control signals based on the outputs of those algorithms. Control signals may be applied to variable circuit elements in the tuning/impedance matching circuits 2402S to adjust the source's operating characteristics, such as power in the resonator and coupling to devices. Control signals may be applied to the power supply or generator to turn the supply on or off, to increase or decrease the power level, to modulate the supply signal and the like.

[00394] The power exchanged between sources and devices may depend on a variety of factors. These factors may include the effective impedance of the sources and devices, the Q's of the sources and devices, the resonant frequencies of the sources and devices, the distances between sources and devices, the interaction of materials and objects in the vicinity of sources and devices and the like. The port measurement circuitry and processing algorithms may work in concert to adjust the resonator parameters to maximize power transfer, to hold the power transfer constant, to controllably adjust the power transfer, and the like, under both dynamic and steady state operating conditions.

[00395] Some, all or none of the sources and devices in a system implementation may include port measurement circuitry 3802S and processing 2410 capabilities. Fig. 44 shows an end-to-end wireless power transmission system in which only the source 102S contains port measurement circuitry 3802 and a processor 2410S. In this case, the device resonator 102D operating characteristics may be fixed or may be adjusted by analog control circuitry and without the need for control signals generated by a processor.

[00396] Fig. 45 shows an end-to-end wireless power transmission system. Both the source and the device may include port measurement circuitry 3802 but in the system of Fig. 45, only the source contains a processor 2410S. The source and device may be in communication with each other and the adjustment of certain system parameters may be in response to control signals that have been wirelessly communicated, such as though wireless communications circuitry 4202, between the source and the device. The wireless communication channel 4204 may be separate from the wireless power transfer channel 4208, or it may be the same. That is, the resonators 102 used for power exchange may also be used to exchange information. In some cases, information may be exchanged by modulating a component a source or device circuit and sensing that change with port parameter or other monitoring equipment.

[00397] Implementations where only the source contains a processor 2410 may be beneficial for multi-device systems where the source can handle all of the tuning and adjustment "decisions" and simply communicate the control signals back to the device(s). This implementation may make the device smaller and cheaper because it may eliminate the need for, or reduce the required functionality of, a processor in the device. A portion of or an entire data set from each port measurement at each device may be sent back to the source microprocessor for analysis, and the control instructions may be sent back to the devices. These communications may be wireless communications.

[00398] Fig. 46 shows an end-to-end wireless power transmission system. In this example, only the source contains port measurement circuitry 3802 and a processor 2410S. The source and device may be in communication, such as via wireless communication circuitry 4202, with each other and the adjustment of certain system parameters may be in response to control signals that have been wirelessly communicated between the source and the device.

[00399] Fig. 47 shows coupled electromagnetic resonators 102 whose frequency and impedance may be automatically adjusted using a processor or a computer. Resonant frequency tuning and continuous impedance adjustment of the source and device resonators may be implemented with reverse biased diodes, Schottky diodes and/or varactor elements contained within the capacitor networks shown as C1, C2, and C3 in Fig. 47. The circuit topology that was built and demonstrated and is described here is exemplary and is not meant to limit the discussion of automatic system tuning and control in any way. Other circuit topologies could be utilized with the measurement and control architectures discussed in this disclosure.

[00400] Device and source resonator impedances and resonant frequencies may be measured with a network analyzer 4402A-B, or by other means described above, and implemented with a controller, such as with Lab View 4404. The measurement circuitry or equipment may output data to a computer or a processor that implements feedback algorithms and dynamically adjusts the frequencies and impedances via a programmable DC voltage source.

[00401] In one arrangement, the reverse biased diodes (Schottky, semiconductor junction, and the like) used to realize the tunable capacitance drew very little DC current and could be reverse biased by amplifiers having large series output resistances. This implementation may enable DC control signals to be applied directly to the controllable circuit elements in the resonator circuit while maintaining a very high-Q in the magnetic resonator.

[00402] C2 biasing signals may be isolated from C1 and/or C3 biasing signals with a DC blocking capacitor as shown in Fig. 47, if the required DC biasing voltages are different. The output of the biasing amplifiers may be bypassed to circuit ground to isolate RF voltages from the biasing amplifiers, and to keep non-fundamental RF voltages from being injected into the resonator. The reverse bias voltages for some of the capacitors may instead be applied through the inductive element in the resonator itself, because the inductive element acts as a short circuit at DC.

[00403] The port parameter measurement circuitry may exchange signals with a processor (including any required ADCs and DACs) as part of a feedback or control system that is used to automatically adjust the resonant frequency, input impedance, energy stored or captured by the resonator or power delivered by a source or to a device load. The processor may also send control signals to tuning or adjustment circuitry in or attached to the magnetic resonator.

[00404] When utilizing varactors or diodes as tunable capacitors, it may be beneficial to place fixed capacitors in parallel and in series with the tunable capacitors operating at high reverse bias voltages in the tuning/matching circuits. This arrangement may yield improvements in circuit and system stability and in power handling capability by optimizing the operating voltages on the tunable capacitors.

[00405] Varactors or other reverse biased diodes may be used as a voltage controlled capacitor. Arrays of varactors may be used when higher voltage compliance or different capacitance is required than that of a single varactor component. Varactors may be arranged in an N by M array connected serially and in parallel and treated as a single two terminal component with different characteristics than the individual varactors in the array. For example, an N by N array of equal varactors where components in each row are connected in parallel and components in each column are connected in series may be used as a two terminal device with the same capacitance as any single varactor in the array but with a voltage compliance that is N times that of a single varactor in the array. Depending on the variability and differences of parameters of the individual varactors in the array additional biasing circuits composed of resistors, inductors, and the like may be needed. A schematic of a four by four array of unbiased varactors 4502 that may be suitable for magnetic resonator applications is shown in Fig. 48.

[00406] Further improvements in system performance may be realized by careful selection of the fixed value capacitor(s) that are placed in parallel and/or in series with the tunable (varactor/diode/capacitor) elements. Multiple fixed capacitors that are switched in or out of the circuit may be able to compensate for changes in resonator *Q*'s, impedances, resonant frequencies, power levels, coupling strengths, and the like, that might be encountered in test, development and operational wireless power transfer systems. Switched capacitor banks and other switched element banks may be used to assure the convergence to the operating frequencies and impedance values required by the system design.

[00407] An exemplary control algorithm for isolated and coupled magnetic resonators may be described for the circuit and system elements shown in Fig. 47. One control algorithm first adjusts each of the source and device resonator loops "in isolation", that is, with the other resonators in the system "shorted out" or "removed" from the system. For practical purposes, a resonator can be "shorted out" by making it resonant at a much lower frequency such as by maximizing the value of C1 and/or C3. This step effectively reduces the coupling between the resonators, thereby effectively reducing the system to a single resonator at a particular frequency and impedance.

[00408] Tuning a magnetic resonator in isolation includes varying the tunable elements in the tuning and matching circuits until the values measured by the port parameter measurement circuitry are at their predetermined, calculated or measured relative values. The desired values for the quantities measured by the port parameter measurement circuitry may be chosen based on the desired matching impedance, frequency, strong coupling parameter, and the like. For the exemplary algorithms disclosed below, the port parameter measurement circuitry measures S-parameters over a range of frequencies. The range of frequencies used to characterize the resonators may be a compromise between the system performance information obtained and computation/measurement speed. For the algorithms described below the frequency range may be approximately +/- 20% of the operating resonant frequency.

[00409] Each isolated resonator may be tuned as follows. First, short out the resonator not being adjusted. Next minimize C1, C2, and C3, in the resonator that is being characterized and adjusted. In most cases there will be fixed circuit elements in parallel with C1, C2, and C3, so this step does not reduce the capacitance values to zero. Next, start increasing C2 until the resonator impedance is matched to the “target” real impedance at any frequency in the range of measurement frequencies described above. The initial “target” impedance may be less than the expected operating impedance for the coupled system.

[00410] C2 may be adjusted until the initial “target” impedance is realized for a frequency in the measurement range. Then C1 and/or C3 may be adjusted until the loop is resonant at the desired operating frequency.

[00411] Each resonator may be adjusted according to the above algorithm. After tuning each resonator in isolation, a second feedback algorithm may be applied to optimize the resonant frequencies and/or input impedances for wirelessly transferring power in the coupled system.

[00412] The required adjustments to C1 and/or C2 and/or C3 in each resonator in the coupled system may be determined by measuring and processing the values of the real and imaginary parts of the input impedance from either and/or both “port(s)” shown in Fig. 43. For coupled resonators, changing the input impedance of one resonator may change the input impedance of the other resonator. Control and tracking algorithms may adjust one port to a desired operating point based on measurements at that port, and then adjust the other port based on measurements at that other port. These steps may be repeated until both sides converge to the desired operating point.

[00413] S-parameters may be measured at both the source and device ports and the following series of measurements and adjustments may be made. In the description that follows,

Z_0 is an input impedance and may be the target impedance. In some cases Z_0 is 50 ohms or is near 50 ohms. Z_1 and Z_2 are intermediate impedance values that may be the same value as Z_0 or may be different than Z_0. Re{value} means the real part of a value and Im{value} means the imaginary part of a value.

[00414] An algorithm that may be used to adjust the input impedance and resonant frequency of two coupled resonators is set forth below:

1) Adjust each resonator "in isolation" as described above.

2) Adjust source C1/C3 until, at ω_o, Re{S11} = (Z_1 +/- ε_{Re}) as follows:

- If Re{S11 @ ω_o } > ($Z_1 + \varepsilon_{Re}$), decrease C1/C3. If Re{S11 @ ω_o } < (Zo - ε_{Re}), increase C1/C3.

3) Adjust source C2 until, at ω_o, Im{S11} = (+/- ε_{Im}) as follows:

- If Im{S11 @ ω_o} > ε_{Im}, decrease C2. If Im{S11 @ ω_o} < - ε_{Im} , increase C2.

4) Adjust device C1/C3 until, at ω_0, Re{S22} = (Z_2 +/- ε_{Re}) as follows:

- If Re{S22 @ ω_o} > ($Z_2 + \varepsilon_{Re}$), decrease C1/C3. If Re{S22 @ ω_o} < (Zo - ε_{Re}), increase C1/C3.

5) Adjust device C2 until, at ω_o, Im{S22} = 0 as follows:

- If Im{S22 @ ω_o} > ε_{Im}, decrease C2. If Im{S22 @ ω_o} < -ε_{Im} , increase C2.

[00415] We have achieved a working system by repeating steps 1-4 until both (Re{S11}, Im{S11}) and (Re{S22}, Im{S22}) converge to ((Z_0 +/- ε_{Re}), (+/- ε_{Im})) at ω_o, where Z_0 is the desired matching impedance and ω_o is the desired operating frequency. Here, ε_{Im} represents the maximum deviation of the imaginary part, at ω_o, from the desired value of 0, and ε_{Re} represents the maximum deviation of the real part from the desired value of Z_0. It is understood that ε_{Im} and ε_{Re} can be adjusted to increase or decrease the number of steps to convergence at the potential cost of system performance (efficiency). It is also understood that steps 1-4 can be performed in a variety of sequences and a variety of ways other than that outlined above (i.e. first adjust the source imaginary part, then the source real part; or first adjust the device real part, then the device imaginary part, etc.) The intermediate impedances Z_1 and Z_2 may be adjusted during steps 1-4 to reduce the number of steps required for convergence. The desire or target impedance value may be complex, and may vary in time or under different operating scenarios.

[00416] Steps 1-4 may be performed in any order, in any combination and any number of times. Having described the above algorithm, variations to the steps or the described implementation may be apparent to one of ordinary skill in the art. The algorithm outlined above may be implemented with any equivalent linear network port parameter measurements (i.e., Z-parameters, Y-parameters, T-parameters, H-parameters, ABCD-parameters, etc.) or other monitor signals described above, in the same way that impedance or admittance can be alternatively used to analyze a linear circuit to derive the same result.

[00417] The resonators may need to be retuned owing to changes in the "loaded" resistances, Rs and Rd, caused by changes in the mutual inductance M (coupling) between the source and device resonators. Changes in the inductances, Ls and Ld, of the inductive elements themselves may be caused by the influence of external objects, as discussed earlier, and may also require compensation. Such variations may be mitigated by the adjustment algorithm described above.

[00418] A directional coupler or a switch may be used to connect the port parameter measurement circuitry to the source resonator and tuning/adjustment circuitry. The port parameter measurement circuitry may measure properties of the magnetic resonator while it is exchanging power in a wireless power transmission system, or it may be switched out of the circuit during system operation. The port parameter measurement circuitry may measure the parameters and the processor may control certain tunable elements of the magnetic resonator at start-up, or at certain intervals, or in response to changes in certain system operating parameters.

[00419] A wireless power transmission system may include circuitry to vary or tune the impedance and/or resonant frequency of source and device resonators. Note that while tuning circuitry is shown in both the source and device resonators, the circuitry may instead be included in only the source or the device resonators, or the circuitry may be included in only some of the source and/or device resonators. Note too that while we may refer to the circuitry as "tuning" the impedance and or resonant frequency of the resonators, this tuning operation simply means that various electrical parameters such as the inductance or capacitance of the structure are being varied. In some cases, these parameters may be varied to achieve a specific predetermined value, in other cases they may be varied in response to a control algorithm or to stabilize a target performance value that is changing. In some cases, the parameters are varied as a function of temperature, of other sources or devices in the area, of the environment, at the like.

[00420] **Applications**

[00421] For each listed application, it will be understood by one of ordinary skill-in-the-art that there are a variety of ways that the resonator structures used to enable wireless power transmission may be connected or integrated with the objects that are supplying or being powered. The resonator may be physically separate from the source and device objects. The resonator may supply or remove power from an object using traditional inductive techniques or through direct electrical connection, with a wire or cable for example. The electrical connection may be from the resonator output to the AC or DC power input port on the object. The electrical connection may be from the output power port of an object to the resonator input.

[00422] FIG. 49 shows a source resonator 4904 that is physically separated from a power supply and a device resonator 4902 that is physically separated from the device 4900, in this illustration a laptop computer. Power may be supplied to the source resonator, and power may be taken from the device resonator directly, by an electrical connection. One of ordinary skill in the art will understand from the materials incorporated by reference that the shape, size, material composition, arrangement, position and orientation of the resonators above are provided by way of non-limiting example, and that a wide variation in any and all of these parameters could be supported by the disclosed technology for a variety of applications.

[00423] Continuing with the example of the laptop, and without limitation, the device resonator may be physically connected to the device it is powering or charging. For example, as shown in FIG. 50a and FIG. 50b, the device resonator 5002 may be (a) integrated into the housing of the device 5000 or (b) it may be attached by an adapter. The resonator 5002 may (FIG. 50b-d) or may not (FIG. 50a) be visible on the device. The resonator may be affixed to the device, integrated into the device, plugged into the device, and the like.

[00424] The source resonator may be physically connected to the source supplying the power to the system. As described above for the devices and device resonators, there are a variety of ways the resonators may be attached to, connected to or integrated with the power supply. One of ordinary skill in the art will understand that there are a variety of ways the resonators may be integrated in the wireless power transmission system, and that the sources and devices may utilize similar or different integration techniques.

[00425] Continuing again with the example of the laptop computer, and without limitation, the laptop computer may be powered, charged or recharged by a wireless power transmission system. A source resonator may be used to supply wireless power and a device resonator may be used to capture the wireless power. A device resonator 5002 may be integrated into the edge of the screen (display) as illustrated in FIG. 50d, and/or into the base of the laptop as illustrated in FIG. 50c. The source resonator 5002 may be integrated into the base of the laptop and the device resonator may be integrated into the edge of the screen. The resonators may also or instead be affixed to the power source and/or the laptop. The source and device resonators may also or instead be physically separated from the power supply and the laptop and may be electrically connected by a cable. The source and device resonators may also or instead be physically separated from the power supply and the laptop and may be electrically coupled using a traditional inductive technique. One of ordinary skill in the art will understand that, while the preceding examples relate to wireless power transmission to a laptop, that the methods and systems disclosed for this application may be suitably adapted for use with other electrical or electronic devices. In general, the source resonator may be external to the source and supplying power to a device resonator that in turn supplies power the device, or the source resonator may be connected to the source and supplying power to a device resonator that in turn supplies power to a portion of the device, or the source resonator may internal to the source and supplying power to a device resonator that in turn supplies power to a portion of the device, as well as any combination of these.

[00426] A system or method disclosed herein may provide power to an electrical or electronics device, such as, and not limited to, phones, cell phones, cordless phones, smart phones, PDAs, audio devices, music players, MP3 players, radios, portable radios and players, wireless headphones, wireless headsets, computers, laptop computers, wireless keyboards, wireless mouse, televisions, displays, flat screen displays, computer displays, displays embedded in furniture, digital picture frames, electronic books, (e.g. the Kindle, e-ink books, magazines, and the like), remote control units (also referred to as controllers, game controllers, commanders, clickers, and the like, and used for the remote control of a plurality of electronics devices, such as televisions, video games, displays, computers, audio visual equipment, lights, and the like), lighting devices, cooling devices, air circulation devices, purification devices, personal hearing aids, power tools, security systems, alarms, bells, flashing lights, sirens, sensors, loudspeakers, electronic locks, electronic keypads, light switches, other electrical switches, and the like. Here the term electronic lock is used to indicate a door lock which operates electronically (e.g. with electronic combo-key, magnetic card, RFID card, and the like) which is placed on a door instead of a mechanical key-lock. Such locks are often battery operated, risking the possibility that the lock might stop working when a battery dies, leaving the user locked-out. This may be avoided where the battery is either charged or completely replaced by a wireless power transmission implementation as described herein.

[00427] Here, the term light switch (or other electrical switch) is meant to indicate any switch (e.g. on a wall of a room) in one part of the room that turns on/off a device (e.g. light fixture at the center of the ceiling) in another part of the room. To install such a switch by direct connection, one would have to run a wire all the way from the device to the switch. Once such a switch is installed at a particular spot, it may be very difficult to move. Alternately, one can envision a 'wireless switch', where "wireless" means the switching (on/off) commands are communicated wirelessly, but such a switch has traditionally required a battery for operation. In general, having too many battery operated switches around a house may be impractical, because those many batteries will need to be replaced periodically. So, a wirelessly communicating switch may be more convenient, provided it is also wirelessly powered. For example, there already exist communications wireless door-bells that are battery powered, but where one still has to replace the battery in them periodically. The remote doorbell button may be made to be completely wireless, where there may be no need to ever replace the battery again. Note that here, the term 'cordless' or 'wireless' or 'communications wireless' is used to indicate that there is a cordless or wireless communications facility between the device and another electrical component, such as the base station for a cordless phone, the computer for a wireless keyboard, and the like. One skilled in the art will recognize that any electrical or electronics device may include a wireless communications facility, and that the systems and methods described herein may be used to add wireless power transmission to the device. As described herein, power to the electrical or electronics device may be delivered from an external or internal source resonator, and to the device or portion of the device. Wireless power transmission may significantly reduce the need to charge and/or replace batteries for devices that enter the near vicinity of the source resonator and thereby may reduce the downtime, cost and disposal issues often associated with batteries.

[00428] The systems and methods described herein may provide power to lights without the need for either wired power or batteries. That is, the systems and methods described herein may provide power to lights without wired connection to any power source, and provide the energy to the light non-radiatively across mid-range distances, such as across a distance of a quarter of a meter, one meter, three meters, and the like. A 'light' as used herein may refer to the light source itself, such as an incandescent light bulb, florescent light bulb lamps, Halogen lamps, gas discharge lamps, fluorescent lamps, neon lamps, high-intensity discharge lamps, sodium vapor lamps, Mercury-vapor lamps, electroluminescent lamps, light emitting diodes (LED) lamps, and the like; the light as part of a light fixture, such as a table lamp, a floor lamp, a ceiling lamp, track lighting, recessed light fixtures, and the like; light fixtures integrated with other functions, such as a light/ceiling fan fixture, and illuminated picture frame, and the like. As such, the systems and methods described herein may reduce the complexity for installing a light, such as by minimizing the installation of electrical wiring, and allowing the user to place or mount the light with minimal regard to sources of wired power. For instance, a light may be placed anywhere in the vicinity of a source resonator, where the source resonator may be mounted in a plurality of different places with respect to the location of the light, such as on the floor of the room above, (e.g. as in the case of a ceiling light and especially when the room above is the attic); on the wall of the next room, on the ceiling of the room below, (e.g. as in the case of a floor lamp); in a component within the room or in the infrastructure of the room as described herein; and the like. For example, a light/ceiling fan combination is often installed in a master bedroom, and the master bedroom often has the attic above it. In this instance a user may more easily install the light/ceiling fan combination in the master bedroom, such as by simply mounting the light/ceiling fan combination to the ceiling, and placing a source coil (plugged into the house wired AC power) in the attic above the mounted fixture. In another example, the light may be an external light, such as a flood light or security light, and the source resonator mounted inside the structure. This way of installing lighting may be particularly beneficial to users who rent their homes, because now they may be able to mount lights and such other electrical components without the need to install new electrical wiring. The control for the light may also be communicated by near-field communications as described herein, or by traditional wireless communications methods.

[00429] The systems and methods described herein may provide power from a source resonator to a device resonator that is either embedded into the device component, or outside the device component, such that the device component may be a traditional electrical component or fixture. For instance, a ceiling lamp may be designed or retrofitted with a device resonator integrated into the fixture, or the ceiling lamp may be a traditional wired fixture, and plugged into a separate electrical facility equipped with the device resonator. In an example, the electrical facility may be a wireless junction box designed to have a device resonator for receiving wireless power, say from a source resonator placed on the floor of the room above (e.g. the attic), and which contains a number of traditional outlets that are powered from the device resonator. The wireless junction box, mounted on the ceiling, may now provide power to traditional wired electrical components on the ceiling (e.g. a ceiling light, track lighting, a ceiling fan). Thus, the ceiling lamp may now be mounted to the ceiling without the need to run wires through the infrastructure of the building. This type of device resonator to traditional outlet junction box may be used in a plurality of applications, including being designed for the interior or exterior of a building, to be made portable, made for a vehicle, and the like. Wireless power may be transferred through common building materials, such as wood, wall board, insulation, glass. brick, stone, concrete, and the like. The benefits of reduced installation cost, re-configurability, and increased application flexibility may provide the user significant benefits over traditional wired installations. The device resonator for a traditional outlet junction box may include a plurality of electrical components for facilitating the transfer of power from the device resonator to the traditional outlets, such as power source electronics which convert the specific frequencies needed to implement efficient power transfer to line voltage, power capture electronics which may convert high frequency AC to usable voltage and frequencies (AC and/or DC), controls which synchronize the capture device and the power output and which ensure consistent, safe, and maximally efficient power transfer, and the like.

[00430] The systems and methods described herein may provide advantages to lights or electrical components that operate in environments that are wet, harsh, controlled, and the like, such has outside and exposed to the rain, in a pool/sauna/shower, in a maritime application, in hermetically sealed components, in an explosive-proof room, on outside signage, a harsh industrial environment in a volatile environment (e.g. from volatile vapors or airborne organics, such as in a grain silo or bakery) , and the like. For example, a light mounted under the water level of a pool is normally difficult to wire up, and is required to be water-sealed despite the need for external wires. But a pool light using the principles disclosed herein may more easily be made water sealed, as there may be no external wires needed. In another example, an explosion proof room, such as containing volatile vapors, may not only need to be hermetically sealed, but may need to have all electrical contacts (that could create a spark) sealed. Again, the principles disclosed herein may provide a convenient way to supply sealed electrical components for such applications.

[00431] The systems and methods disclosed herein may provide power to game controller applications, such as to a remote handheld game controller. These game controllers may have been traditionally powered solely by batteries, where the game controller's use and power profile caused frequent changing of the battery, battery pack, rechargeable batteries, and the like, that may not have been ideal for the consistent use to the game controller, such as during extended game play. A device resonator may be placed into the game controller, and a source resonator, connected to a power source, may be placed in the vicinity. Further, the device resonator in the game controller may provide power directly to the game controller electronics without a battery; provide power to a battery, battery pack, rechargeable battery, and the like, which then provides power to the game controller electronics; and the like. The game controller may utilize multiple battery packs, where each battery pack is equipped with a device resonator, and thus may be constantly recharging while in the vicinity of the source resonator, whether plugged into the game controller or not. The source resonator may be resident in a main game controller facility for the game, where the main game controller facility and source resonator are supplied power from AC 'house' power; resident in an extension facility form AC power, such as in a source resonator integrated into an 'extension cord'; resident in a game chair, which is at least one of plugged into the wall AC, plugged into the main game controller facility, powered by a battery pack in the game chair; and the like. The source resonator may be placed and implemented in any of the configurations described herein.

[00432] The systems and methods disclosed herein may integrate device resonators into battery packs, such as battery packs that are interchangeable with other battery packs. For instance, some portable devices may use up electrical energy at a high rate such that a user may need to have multiple interchangeable battery packs on hand for use, or the user may operate the device out of range of a source resonator and need additional battery packs to continue operation, such as for power tools, portable lights, remote control vehicles, and the like. The use of the principles disclosed herein may not only provide a way for device resonator enabled battery packs to be recharged while in use and in range, but also for the recharging of battery packs not currently in use and placed in range of a source resonator. In this way, battery packs may always be ready to use when a user runs down the charge of a battery pack being used. For example, a user may be working with a wireless power tool, where the current requirements may be greater than can be realized through direct powering from a source resonator. In this case, despite the fact that the systems and methods described herein may be providing charging power to the in-use battery pack while in range, the battery pack may still run down, as the power usage may have exceeded the recharge rate. Further, the user may simply be moving in and out of range, or be completely out of range while using the device. However, the user may have placed additional battery packs in the vicinity of the source resonator, which have been recharged while not in use, and are now charged sufficiently for use. In another example, the user may be working with the power tool away from the vicinity of the source resonator, but leave the supplemental battery packs to charge in the vicinity of the source resonator, such as in a room with a portable source resonator or extension cord source resonator, in the user’s vehicle, in user’s tool box, and the like. In this way, the user may not have to worry about taking the time to, and/or remembering to plug in their battery packs for future use. The user may only have to change out the used battery pack for the charged battery pack and place the used one in the vicinity of the source resonator for recharging. Device resonators may be built into enclosures with known battery form factors and footprints and may replace traditional chemical batteries in known devices and applications. For example, device resonators may be built into enclosures with mechanical dimensions equivalent to AA batteries, AAA batteries, D batteries, 9V batteries, laptop batteries, cell phone batteries, and the like. The enclosures may include a smaller “button battery” in addition to the device resonator to store charge and provide extended operation, either in terms of time or distance. Other energy storage devices in addition to or instead of button batteries may be integrated with the device resonators and any associated power conversion circuitry. These new energy packs may provide similar voltage and current levels as provided by traditional batteries, but may be composed of device resonators, power conversion electronics, a small battery, and the like. These new energy packs may last longer than traditional batteries because they may be more easily recharged and may be recharging constantly when they are located in a wireless power zone. In addition, such energy packs may be lighter than traditional batteries, may be safer to use and store, may operate over wider temperature and humidity ranges, may be less harmful to the environment when thrown away, and the like. As described herein, these energy packs may last beyond the life of the product when used in wireless power zones as described herein.

[00433] The systems and methods described herein may be used to power visual displays, such as in the case of the laptop screen, but more generally to include the great variety and diversity of displays utilized in today's electrical and electronics components, such as in televisions, computer monitors, desktop monitors, laptop displays, digital photo frames, electronic books, mobile device displays (e.g. on phones, PDAs, games, navigation devices, DVD players), and the like. Displays that may be powered through one or more of the wireless power transmission systems described herein may also include embedded displays, such as embedded in electronic components (e.g. audio equipment, home appliances, automotive displays, entertainment devices, cash registers, remote controls), in furniture, in building infrastructure, in a vehicle, on the surface of an object (e.g. on the surface of a vehicle, building, clothing, signs, transportation), and the like. Displays may be very small with tiny resonant devices, such as in a smart card as described herein, or very large, such as in an advertisement sign. Displays powered using the principles disclosed herein may also be any one of a plurality of imaging technologies, such as liquid crystal display (LCD), thin film transistor LCD, passive LCD, cathode ray tube (CRT), plasma display, projector display (e.g. LCD, DLP, LCOS), surface-conduction electron-emitter display (SED), organic light-emitting diode (OLED), and the like. Source coil configurations may include attaching to a primary power source, such as building power, vehicle power, from a wireless extension cord as described herein, and the like; attached to component power, such as the base of an electrical component (e.g. the base of a computer, a cable box for a TV); an intermediate relay source coil; and the like. For example, hanging a digital display on the wall may be very appealing, such as in the case of a digital photo frame that receives its information signals wirelessly or through a portable memory device, but the need for an unsightly power cord may make it aesthetically unpleasant. However, with a device coil embedded in the digital photo frame, such as wrapped within the frame portion, may allow the digital photo frame to be hung with no wires at all. The source resonator may then be placed in the vicinity of the digital photo frame, such as in the next room on the other side of the wall, plugged directly into a traditional power outlet, from a wireless extension cord as described herein, from a central source resonator for the room, and the like.

[00434] The systems and methods described herein may provide wireless power transmission between different portions of an electronics facility. Continuing with the example of the laptop computer, and without limitation, the screen of the laptop computer may require power from the base of the laptop. In this instance, the electrical power has been traditionally routed via direct electrical connection from the base of the laptop to the screen over a hinged portion of the laptop between the screen and the base. When a wired connection is utilized, the wired connection may tend to wear out and break, the design functionality of the laptop computer may be limited by the required direct electrical connection, the design aesthetics of the laptop computer may be limited by the required direct electrical connection, and the like. However, a wireless connection may be made between the base and the screen. In this instance, the device resonator may be placed in the screen portion to power the display, and the base may be either powered by a second device resonator, by traditional wired connections, by a hybrid of resonator-battery- direct electrical connection, and the like. This may not only improve the reliability of the power connection due to the removal of the physical wired connection, but may also allow designers to improve the functional and/or aesthetic design of the hinge portion of the laptop in light of the absence of physical wires associated with the hinge. Again, the laptop computer has been used here to illustrate how the principles disclosed herein may improve the design of an electric or electronic device, and should not be taken as limiting in any way. For instance, many other electrical devices with separated physical portions could benefit from the systems and methods described herein, such as a refrigerator with electrical functions on the door, including an ice maker, a sensor system, a light, and the like; a robot with movable portions, separated by joints; a car’s power system and a component in the car’s door; and the like. The ability to provide power to a device via a device resonator from an external source resonator, or to a portion of the device via a device resonator from either external or internal source resonators, will be recognized by someone skilled in the art to be widely applicable across the range of electric and electronic devices.

[00435] The systems and methods disclosed herein may provide for a sharing of electrical power between devices, such as between charged devices and uncharged devices. For instance a charged up device or appliance may act like a source and send a predetermined amount of energy, dialed in amount of energy, requested and approved amount of energy, and the like, to a nearby device or appliance. For example, a user may have a cell phone and a digital camera that are both capable of transmitting and receiving power through embedded source and device resonators, and one of the devices, say the cell phone, is found to be low on charge. The user may then transfer charge from the digital camera to the cell phone. The source and device resonators in these devices may utilize the same physical resonator for both transmission and reception, utilize separate source and device resonators, one device may be designed to receive and transmit while the other is designed to receive only, one device may be designed to transmit only and the other to receive only, and the like.

[00436] To prevent complete draining the battery of a device it may have a setting allowing a user to specify how much of the power resource the receiving device is entitled to. It may be useful, for example, to put a limit on the amount of power available to external devices and to have the ability to shut down power transmission when battery power falls below a threshold.

[00437] The systems and methods described herein may provide wireless power transfer to a nearby electrical or electronics component in association with an electrical facility, where the source resonator is in the electrical facility and the device resonator is in the electronics component. The source resonator may also be connected to, plugged into, attached to the electrical facility, such as through a universal interface (e.g. a USB interface, PC card interface), supplemental electrical outlet, universal attachment point, and the like, of the electrical facility. For example, the source resonator may be inside the structure of a computer on a desk, or be integrated into some object, pad, and the like, that is connected to the computer, such as into one of the computer's USB interfaces. In the example of the source resonator embedded in the object, pad, and the like, and powered through a USB interface, the source resonator may then be easily added to a user's desktop without the need for being integrated into any other electronics device, thus conveniently providing a wireless energy zone around which a plurality of electric and/or electronics devices may be powered. The electrical facility may be a computer, a light fixture, a dedicated source resonator electrical facility, and the like, and the nearby components may be computer peripherals, surrounding electronics components, infrastructure devices, and the like, such as computer keyboards, computer mouse, fax machine, printer, speaker system, cell phone, audio device, intercom, music player, PDA, lights, electric pencil sharpener, fan, digital picture frame, calculator, electronic games, and the like. For example, a computer system may be the electrical facility with an integrated source resonator that utilizes a 'wireless keyboard' and 'wireless mouse', where the use of the term wireless here is meant to indicate that there is wireless communication facility between each device and the computer, and where each device must still contain a separate battery power source. As a result, batteries would need to be replaced periodically, and in a large company, may result in a substantial burden for support personnel for replacement of batteries, cost of batteries, and proper disposal of batteries. Alternatively, the systems and methods described herein may provide wireless power transmission from the main body of the computer to each of these peripheral devices, including not only power to the keyboard and mouse, but to other peripheral components such as a fax, printer, speaker system, and the like, as described herein. A source resonator integrated into the electrical facility may provide wireless power transmission to a plurality of peripheral devices, user devices, and the like, such that there is a significant reduction in the need to charge and/or replace batteries for devices in the near vicinity of the source resonator integrated electrical facility. The electrical facility may also provide tuning or auto-tuning software, algorithms, facilities, and the like, for adjusting the power transfer parameters between the electrical facility and the wirelessly powered device. For example, the electrical facility may be a computer on a user's desktop, and the source resonator may be either integrated into the computer or plugged into the computer (e.g. through a USB connection), where the computer provides a facility for providing the tuning algorithm (e.g. through a software program running on the computer).

[00438] The systems and methods disclosed herein may provide wireless power transfer to a nearby electrical or electronics component in association with a facility infrastructure component, where the source resonator is in, or mounted on, the facility infrastructure component and the device resonator is in the electronics component. For instance, the facility infrastructure component may be a piece of furniture, a fixed wall, a movable wall or partition, the ceiling, the floor, and the source resonator attached or integrated into a table or desk (e.g. just below/above the surface, on the side, integrated into a table top or table leg), a mat placed on the floor (e.g. below a desk, placed on a desk), a mat on the garage floor (e.g. to charge the car and/or devices in the car), in a parking lot/garage (e.g. on a post near where the car is parked), a television (e.g. for charging a remote control), a computer monitor (e.g. to power/charge a wireless keyboard, wireless mouse, cell phone), a chair (e.g. for powering electric blankets, medical devices, personal health monitors), a painting, office furniture, common household appliances, and the like. For example, the facility infrastructure component may be a lighting fixture in an office cubical, where the source resonator and light within the lighting fixture are both directly connected to the facility's wired electrical power. However, with the source resonator now provided in the lighting fixture, there would be no need to have any additional wired connections for those nearby electrical or electronics components that are connected to, or integrated with, a device resonator. In addition, there may be a reduced need for the replacement of batteries for devices with device resonators, as described herein.

[00439] The use of the systems and methods described herein to supply power to electrical and electronic devices from a central location, such as from a source resonator in an electrical facility, from a facility infrastructure component and the like, may minimize the electrical wiring infrastructure of the surrounding work area. For example, in an enterprise office space there are typically a great number of electrical and electronic devices that need to be powered by wired connections. With utilization of the systems and methods described herein, much of this wiring may be eliminated, saving the enterprise the cost of installation, decreasing the physical limitations associated with office walls having electrical wiring, minimizing the need for power outlets and power strips, and the like. The systems and methods described herein may save money for the enterprise through a reduction in electrical infrastructure associated with installation, re-installation (e.g., reconfiguring office space), maintenance, and the like. In another example, the principles disclosed herein may allow the wireless placement of an electrical outlet in the middle of a room. Here, the source could be placed on the ceiling of a basement below the location on the floor above where one desires to put an outlet. The device resonator could be placed on the floor of the room right above it. Installing a new lighting fixture (or any other electric device for that matter, e.g. camera, sensor, etc., in the center of the ceiling may now be substantially easier for the same reason).

[00440] In another example, the systems and methods described herein may provide power "through" walls. For instance, suppose one has an electric outlet in one room (e.g. on a wall), but one would like to have an outlet in the next room, but without the need to call an electrician, or drill through a wall, or drag a wire around the wall, or the like. Then one might put a source resonator on the wall in one room, and a device resonator outlet/pickup on the other side of the wall. This may power a flat-screen TV or stereo system or the like (e.g. one may not want to have an ugly wire climbing up the wall in the living room, but doesn't mind having a similar wire going up the wall in the next room, e.g. storage room or closet, or a room with furniture that blocks view of wires running along the wall). The systems and methods described herein may be used to transfer power from an indoor source to various electric devices outside of homes or buildings without requiring holes to be drilled through, or conduits installed in, these outside walls. In this case, devices could be wirelessly powered outside the building without the aesthetic or structural damage or risks associated with drilling holes through walls and siding. In addition, the systems and methods described herein may provide for a placement sensor to assist in placing an interior source resonator for an exterior device resonator equipped electrical component. For example, a home owner may place a security light on the outside of their home which includes a wireless device resonator, and now needs to adequately or optimally position the source resonator inside the home. A placement sensor acting between the source and device resonators may better enable that placement by indicating when placement is good, or to a degree of good, such as in a visual indication, an audio indication, a display indication, and the like. In another example, and in a similar way, the systems and methods described herein may provide for the installation of equipment on the roof of a home or building, such as radio transmitters and receivers, solar panels and the like. In the case of the solar panel, the source resonator may be associated with the panel, and power may be wirelessly transferred to a distribution panel inside the building without the need for drilling through the roof. The systems and methods described herein may allow for the mounting of electric or electrical components across the walls of vehicles (such as through the roof) without the need to drill holes, such as for automobiles, water craft, planes, trains, and the like. In this way, the vehicle's walls may be left intact without holes being drilled, thus maintaining the value of the vehicle, maintaining watertightness, eliminating the need to route wires, and the like. For example, mounting a siren or light to the roof of a police car decreases the future resale of the car, but with the systems and methods described herein, any light, horn, siren, and the like, may be attached to the roof without the need to drill a hole.

[00441] The systems and methods described herein may be used for wireless transfer of power from solar photovoltaic (PV) panels. PV panels with wireless power transfer capability may have several benefits including simpler installation, more flexible, reliable, and weatherproof design. Wireless power transfer may be used to transfer power from the PV panels to a device, house, vehicle, and the like. Solar PV panels may have a wireless source resonator allowing the PV panel to directly power a device that is enabled to receive the wireless power. For example, a solar PV panel may be mounted directly onto the roof of a vehicle, building, and the like. The energy captured by the PV panel may be wirelessly transferred directly to devices inside the vehicle or under the roof of a building. Devices that have resonators can wirelessly receive power from the PV panel. Wireless power transfer from PV panels may be used to transfer energy to a resonator that is coupled to the wired electrical system of a house, vehicle, and the like allowing traditional power distribution and powering of conventional devices without requiring any direct contact between the exterior PV panels and the internal electrical system.

[00442] With wireless power transfer significantly simpler installation of rooftop PV panels is possible because power may be transmitted wirelessly from the panel to a capture resonator in the house, eliminating all outside wiring, connectors, and conduits, and any holes through the roof or walls of the structure. Wireless power transfer used with solar cells may have a benefit in that it can reduced roof danger since it eliminates the need for electricians to work on the roof to interconnect panels, strings, and junction boxes. Installation of solar panels integrated with wireless power transfer may require less skilled labor since fewer electrical contacts need to be made. Less site specific design may be required with wireless power transfer since the technology gives the installer the ability to individually optimize and position each solar PV panel, significantly reducing the need for expensive engineering and panel layout services. There may not be need to carefully balance the solar load on every panel and no need for specialized DC wiring layout and interconnections.

[00443] For rooftop or on-wall installations of PV panels, the capture resonator may be mounted on the underside of the roof, inside the wall, or in any other easily accessible inside space within a foot or two of the solar PV panel. A diagram showing a possible general rooftop PV panel installation is shown in Figure 51. Various PV solar collectors may be mounted in top of a roof with wireless power capture coils mounted inside the building under the roof. The resonator coils in the PV panels can transfer their energy wirelessly through the roof to the wireless capture coils. The captured energy from the PV cells may be collected and coupled to the electrical system of the house to power electric and electronic devices or coupled to the power grid when more power than needed is generated. Energy is captured from the PV cells without requiring holes or wires that penetrate the roof or the walls of the building. Each PV panel may have a resonator that is coupled to a corresponding resonator on the interior of the vehicle or building. Multiple panels may utilize wireless power transfer between each other to transfer or collect power to one or a couple of designated panels that are coupled to resonators on the interior of the vehicle of house. Panels may have wireless power resonators on their sides or in their perimeter that can couple to resonators located in other like panels allowing transfer of power from panel to panel. An additional bus or connection structure may be provided that wirelessly couples the power from multiple panels on the exterior of a building or vehicle and transfers power to one or a more resonators on the interior of building or vehicle.

[00444] For example, as shown in Fig. 51, a source resonator 5102 may be coupled to a PV cell 5100 mounted on top of roof 5104 of a building. A corresponding capture resonator 5106 is placed inside the building. The solar energy captured by the PV cells can then be transferred between the source resonators 5102 outside to the device resonators 5106 inside the building without having direct holes and connections through the building.

[00445] Each solar PV panel with wireless power transfer may have its own inverter, significantly improving the economics of these solar systems by individually optimizing the power production efficiency of each panel, supporting a mix of panel sizes and types in a single installation, including single panel "pay-as-you-grow" system expansions. Reduction of installation costs may make a single panel economical for installation. Eliminating the need for panel string designs and careful positioning and orienting of multiple panels, and eliminating a single point of failure for the system.

[00446] Wireless power transfer in PV solar panels may enable more solar deployment scenarios because the weather-sealed solar PV panels eliminate the need to drill holes for wiring through sealed surfaces such as car roofs and ship decks, and eliminate the requirement that the panels be installed in fixed locations. With wireless power transfer, PV panels may be deployed temporarily, and then moved or removed, without leaving behind permanent alterations to the surrounding structures. They may be placed out in a yard on sunny days, and moved around to follow the sun, or brought inside for cleaning or storage, for example. For backyard or mobile solar PV applications, an extension cord with a wireless energy capture device may be thrown on the ground or placed near the solar unit. The capture extension cord can be completely sealed from the elements and electrically isolated, so that it may be used in any indoor or outdoor environment.

[00447] With wireless power transfer no wires or external connections may be necessary and the PV solar panels can be completely weather sealed. Significantly improved reliability and lifetime of electrical components in the solar PV power generation and transmission circuitry can be expected since the weather-sealed enclosures can protect components from UV radiation, humidity, weather, and the like. With wireless power transfer and weather-sealed enclosures it may be possible to use less expensive components since they will no longer be directly exposed to external factors and weather elements and it may reduce the cost of PV panels.

[00448] Power transfer between the PV panels and the capture resonators inside a building or a vehicle may be bidirectional. Energy may be transmitted from the house grid to the PV panels to provide power when the panels do not have enough energy to perform certain tasks such. Reverse power flow can be used to melt snow from the panels, or power motors that will position the panels in a more favorable positions with respect to the sun energy. Once the snow is melted or the panels are repositioned and the PV panels can generate their own energy the direction of power transfer can be returned to normal delivering power from the PV panels to buildings, vehicles, or devices.

[00449] PV panels with wireless power transfer may include auto-tuning on installation to ensure maximum and efficient power transfer to the wireless collector. Variations in roofing materials or variations in distances between the PV panels and the wireless power collector in different installations may affect the performance or perturb the properties of the resonators of the wireless power transfer. To reduce the installation complexity the wireless power transfer components may include a tuning capability to automatically adjust their operating point to compensate for any effects due to materials or distance. Frequency, impedance, capacitance, inductance, duty cycle, voltage levels and the like may be adjusted to ensure efficient and safe power transfer

[00450] The systems and methods described herein may be used to provide a wireless power zone on a temporary basis or in extension of traditional electrical outlets to wireless power zones, such as through the use of a wireless power extension cord. For example, a wireless power extension cord may be configured as a plug for connecting into a traditional power outlet, a long wire such as in a traditional power extension cord, and a resonant source coil on the other end (e.g. in place of, or in addition to, the traditional socket end of the extension The wireless extension cord may also be configured where there are source resonators at a plurality of locations along the wireless extension cord. This configuration may then replace any traditional extension cord where there are wireless power configured devices, such as providing wireless power to a location where there is no convenient power outlet (e.g. a location in the living room where there's no outlet), for temporary wireless power where there is no wired power infrastructure (e.g. a construction site), out into the yard where there are no outlets (e.g. for parties or for yard grooming equipment that is wirelessly powered to decrease the chances of cutting the traditional electrical cord), and the like. The wireless extension cord may also be used as a drop within a wall or structure to provide wireless power zones within the vicinity of the drop. For example, a wireless extension cord could be run within a wall of a new or renovated room to provide wireless power zones without the need for the installation of traditional electrical wiring and outlets.

[00451] The systems and methods described herein may be utilized to provide power between moving parts or rotating assemblies of a vehicle, a robot, a mechanical device, a wind turbine, or any other type of rotating device or structure with moving parts such as robot arms, construction vehicles, movable platforms and the like. Traditionally, power in such systems may have been provided by slip rings or by rotary joints for example. Using wireless power transfer as described herein, the design simplicity, reliability and longevity of these devices may be significantly improved because power can be transferred over a range of distances without any physical connections or contact points that may wear down or out with time. In particular, the preferred coaxial and parallel alignment of the source and device coils may provide wireless power transmission that is not severely modulated by the relative rotational motion of the two coils.

[00452] The systems and methods described herein may be utilized to extend power needs beyond the reach of a single source resonator by providing a series of source-device-source-device resonators. For instance, suppose an existing detached garage has no electrical power and the owner now wants to install a new power service. However, the owner may not want to run wires all over the garage, or have to break into the walls to wire electrical outlets throughout the structure. In this instance, the owner may elect to connect a source resonator to the new power service, enabling wireless power to be supplied to device resonator outlets throughout the back of the garage. The owner may then install a device-source 'relay' to supply wireless power to device resonator outlets in the front of the garage. That is, the power relay may now receive wireless power from the primary source resonator, and then supply available power to a second source resonator to supply power to a second set of device resonators in the front of the garage. This configuration may be repeated again and again to extend the effective range of the supplied wireless power.

[00453] Multiple resonators may be used to extend power needs around an energy blocking material. For instance, it may be desirable to integrate a source resonator into a computer or computer monitor such that the resonator may power devices placed around and especially in front of the monitor or computer such as keyboards, computer mice, telephones , and the like. Due to aesthetics, space constraints, and the like an energy source that may be used for the source resonator may only be located or connected to in the back of the monitor or computer. In many designs of computer or monitors metal components and metal containing circuits are used in the design and packaging which may limit and prevent power transfer from source resonator in the back of the monitor or computer to the front of the monitor or computer. An additional repeater resonator may be integrated into the base or pedestal of the monitor or computer that couples to the source resonator in the back of the monitor or computer and allows power transfer to the space in front of the monitor or computer. The intermediate resonator integrated into the base or pedestal of the monitor or computer does not require an additional power source, it captures power from the source resonator and transfers power to the front around the blocking or power shielding metal components of the monitor or computer.

[00454] The systems and methods described herein may be built-into, placed on, hung from, embedded into, integrated into, and the like, the structural portions of a space, such as a vehicle, office, home, room, building, outdoor structure, road infrastructure, and the like. For instance, one or more sources may be built into, placed on, hung from, embedded or integrated into a wall, a ceiling or ceiling panel, a floor, a divider, a doorway, a stairwell, a compartment, a road surface, a sidewalk, a ramp, a fence, an exterior structure, and the like. One or more sources may be built into an entity within or around a structure, for instance a bed, a desk, a chair, a rug, a mirror, a clock, a display, a television, an electronic device, a counter, a table, a piece of furniture, a piece of artwork, an enclosure, a compartment, a ceiling panel, a floor or door panel, a dashboard, a trunk, a wheel well, a post, a beam, a support or any like entity. For example, a source resonator may be integrated into the dashboard of a user's car so that any device that is equipped with or connected to a device resonator may be supplied with power from the dashboard source resonator. In this way, devices brought into or integrated into the car may be constantly charged or powered while in the car.

[00455] The systems and methods described herein may provide power through the walls of vehicles, such as boats, cars, trucks, busses, trains, planes, satellites and the like. For instance, a user may not want to drill through the wall of the vehicle in order to provide power to an electric device on the outside of the vehicle. A source resonator may be placed inside the vehicle and a device resonator may be placed outside the vehicle (e.g. on the opposite side of a window, wall or structure). In this way the user may achieve greater flexibility in optimizing the placement, positioning and attachment of the external device to the vehicle, (such as without regard to supplying or routing electrical connections to the device). In addition, with the electrical power supplied wirelessly, the external device may be sealed such that it is water tight, making it safe if the electric device is exposed to weather (e.g. rain), or even submerged under water. Similar techniques may be employed in a variety of applications, such as in charging or powering hybrid vehicles, navigation and communications equipment, construction equipment, remote controlled or robotic equipment and the like, where electrical risks exist because of exposed conductors. The systems and methods described herein may provide power through the walls of vacuum chambers or other enclosed spaces such as those used in semiconductor growth and processing, material coating systems, aquariums, hazardous materials handling systems and the like. Power may be provided to translation stages, robotic arms, rotating stages, manipulation and collection devices, cleaning devices and the like.

[00456] The systems and methods described herein may provide wireless power to a kitchen environment, such as to counter-top appliances, including mixers, coffee makers, toasters, toaster ovens, grills, griddles, electric skillets, electric pots, electric woks, waffle makers, blenders, food processors, crock pots, warming trays, induction cooktops, lights, computers, displays, and the like. This technology may improve the mobility and/or positioning flexibility of devices, reduce the number of power cords stored on and strewn across the counter-top, improve the washability of the devices, and the like. For example, an electric skillet may traditionally have separate portions, such as one that is submersible for washing and one that is not submersible because it includes an external electrical connection (e.g. a cord or a socket for a removable cord). However, with a device resonator integrated into the unit, all electrical connections may be sealed, and so the entire device may now be submersed for cleaning. In addition, the absence of an external cord may eliminate the need for an available electrical wall outlet, and there is no longer a need for a power cord to be placed across the counter or for the location of the electric griddle to be limited to the location of an available electrical wall outlet.

[00457] The systems and methods described herein may provide continuous power/charging to devices equipped with a device resonator because the device doesn't leave the proximity of a source resonator, such as fixed electrical devices, personal computers, intercom systems, security systems, household robots, lighting, remote control units, televisions, cordless phones, and the like. For example, a household robot (e.g. ROOMBA) could be powered/charged via wireless power, and thus work arbitrarily long without recharging. In this way, the power supply design for the household robot may be changed to take advantage of this continuous source of wireless power, such as to design the robot to only use power from the source resonator without the need for batteries, use power from the source resonator to recharge the robot's batteries, use the power from the source resonator to trickle charge the robot's batteries, use the power from the source resonator to charge a capacitive energy storage unit, and the like. Similar optimizations of the power supplies and power circuits may be enabled, designed, and realized, for any and all of the devices disclosed herein.

[00458] The systems and methods described herein may be able to provide wireless power to electrically heated blankets, heating pads/patches, and the like. These electrically heated devices may find a variety of indoor and outdoor uses. For example, hand and foot warmers supplied to outdoor workers such as guards, policemen, construction workers and the like might be remotely powered from a source resonator associated with or built into a nearby vehicle, building, utility pole, traffic light, portable power unit, and the like.

[00459] The systems and methods described herein may be used to power a portable information device that contains a device resonator and that may be powered up when the information device is near an information source containing a source resonator. For instance, the information device may be a card (e.g. credit card, smart card, electronic card, and the like) carried in a user's pocket, wallet, purse, vehicle, bike, and the like. The portable information device may be powered up when it is in the vicinity of an information source that then transmits information to the portable information device that may contain electronic logic, electronic processors, memory, a display, an LCD display, LEDs, RFID tags, and the like. For example, the portable information device may be a credit card with a display that "turns on" when it is near an information source, and provide the user with some information such as, “You just received a coupon for 50% off your next Coca Cola purchase”. The information device may store information such as coupon or discount information that could be used on subsequent purchases. The portable information device may be programmed by the user to contain tasks, calendar appointments, to-do lists, alarms and reminders, and the like. The information device may receive up-to-date price information and inform the user of the location and price of previously selected or identified items.

[00460] The systems and methods described herein may provide wireless power transmission to directly power or recharge the batteries in sensors, such as environmental sensors, security sensors, agriculture sensors, appliance sensors, food spoilage sensors, power sensors, and the like, which may be mounted internal to a structure, external to a structure, buried underground, installed in walls, and the like. For example, this capability may replace the need to dig out old sensors to physically replace the battery, or to bury a new sensor because the old sensor is out of power and no longer operational. These sensors may be charged up periodically through the use of a portable sensor source resonator charging unit. For instance, a truck carrying a source resonator equipped power source, say providing ~kW of power, may provide enough power to a ~mW sensor in a few minutes to extend the duration of operation of the sensor for more than a year. Sensors may also be directly powered, such as powering sensors that are in places where it is difficult to connect to them with a wire but they are still within the vicinity of a source resonator, such as devices outside of a house (security camera), on the other side of a wall, on an electric lock on a door, and the like. In another example, sensors that may need to be otherwise supplied with a wired power connection may be powered through the systems and methods described herein. For example, a ground fault interrupter breaker combines residual current and over-current protection in one device for installation into a service panel. However, the sensor traditionally has to be independently wired for power, and this may complicate the installation. However, with the systems and methods described herein the sensor may be powered with a device resonator, where a single source resonator is provided within the service panel, thus simplifying the installation and wiring configuration within the service panel. In addition, the single source resonator may power device resonators mounted on either side of the source resonator mounted within the service panel, throughout the service panel, to additional nearby service panels, and the like. The systems and methods described herein may be employed to provide wireless power to any electrical component associated with electrical panels, electrical rooms, power distribution and the like, such as in electric switchboards, distribution boards, circuit breakers, transformers, backup batteries, fire alarm control panels, and the like. Through the use of the systems and methods described herein, it may be easier to install, maintain, and modify electrical distribution and protection components and system installations.

[00461] In another example, sensors that are powered by batteries may run continuously, without the need to change the batteries, because wireless power may be supplied to periodically or continuously recharge or trickle charge the battery. In such applications, even low levels of power may adequately recharge or maintain the charge in batteries, significantly extending their lifetime and usefulness. In some cases, the battery life may be extended to be longer than the lifetime of the device it is powering, making it essentially a battery that “lasts forever”.

[00462] The systems and methods described herein may be used for charging implanted medical device batteries, such as in an artificial heart, pacemaker, heart pump, insulin pump, implanted coils for nerve or acupressure/acupuncture point stimulation, and the like. For instance, it may not be convenient or safe to have wires sticking out of a patient because the wires may be a constant source of possible infection and may generally be very unpleasant for the patient. The systems and methods described herein may also be used to charge or power medical devices in or on a patient from an external source, such as from a bed or a hospital wall or ceiling with a source resonator. Such medical devices may be easier to attach, read, use and monitor the patient. The systems and methods described herein may ease the need for attaching wires to the patient and the patient’s bed or bedside, making it more convenient for the patient to move around and get up out of bed without the risk of inadvertently disconnecting a medical device. This may, for example, be usefully employed with patients that have multiple sensors monitoring them, such as for measuring pulse, blood pressure, glucose, and the like. For medical and monitoring devices that utilize batteries, the batteries may need to be replaced quite often, perhaps multiple times a week. This may present risks associated with people forgetting to replace batteries, not noticing that the devices or monitors are not working because the batteries have died, infection associated with improper cleaning of the battery covers and compartments, and the like.

[00463] The systems and methods described herein may reduce the risk and complexity of medical device implantation procedures. Today many implantable medical devices such as ventricular assist devices, pacemakers, defibrillators and the like, require surgical implantation due to their device form factor, which is heavily influenced by the volume and shape of the long-life battery that is integrated in the device. In one aspect, there is described herein a non-invasive method of recharging the batteries so that the battery size may be dramatically reduced, and the entire device may be implanted, such as via a catheter. A catheter implantable device may include an integrated capture or device coil. A catheter implantable capture or device coil may be designed so that it may be wired internally, such as after implantation. The capture or device coil may be deployed via a catheter as a rolled up flexible coil (e.g. rolled up like two scrolls, easily unrolled internally with a simple spreader mechanism). The power source coil may be worn in a vest or article of clothing that is tailored to fit in such a way that places the source in proper position, may be placed in a chair cushion or bed cushion, may be integrated into a bed or piece of furniture, and the like.

[00464] The systems and methods described herein may enable patients to have a 'sensor vest', sensor patch, and the like, that may include at least one of a plurality of medical sensors and a device resonator that may be powered or charged when it is in the vicinity of a source resonator. Traditionally, this type of medical monitoring facility may have required batteries, thus making the vest, patch, and the like, heavy, and potentially impractical. But using the principles disclosed herein, no batteries (or a lighter rechargeable battery) may be required, thus making such a device more convenient and practical, especially in the case where such a medical device could be held in place without straps, such as by adhesive, in the absence of batteries or with substantially lighter batteries. A medical facility may be able to read the sensor data remotely with the aim of anticipating (e.g. a few minutes ahead of) a stroke, a heart-attack, or the like. When the vest is used by a person in a location remote from the medical facility, such as in their home, the vest may then be integrated with a cell-phone or communications device to call an ambulance in case of an accident or a medical event. The systems and methods described herein may be of particular value in the instance when the vest is to be used by an elderly person, where traditional non-wireless recharging practices (e.g. replacing batteries, plugging in at night, and the like) may not be followed as required. The systems and methods described herein may also be used for charging devices that are used by or that aid handicapped or disabled people who may have difficulty replacing or recharging batteries, or reliably supplying power to devices they enjoy or rely on.

[00465] The systems and methods described herein may be used for the charging and powering of artificial limbs. Artificial limbs have become very capable in terms of replacing the functionality of original limbs, such as arms, legs, hands and feet. However, an electrically powered artificial limb may require substantial power, (such as 10-20W) which may translate into a substantial battery. In that case, the amputee may be left with a choice between a light battery that doesn't last very long, and a heavy battery that lasts much longer, but is more difficult to 'carry' around. The systems and methods described herein may enable the artificial limb to be powered with a device resonator, where the source resonator is either carried by the user and attached to a part of the body that may more easily support the weight (such as on a belt around the waist, for example) or located in an external location where the user will spend an adequate amount of time to keep the device charged or powered, such as at their desk, in their car, in their bed, and the like.

[00466] The systems and methods described herein may be used for charging and powering of electrically powered exo-skeletons, such as those used in industrial and military applications, and for elderly/weak/sick people. An electrically powered exo-skeleton may provide up to a 10-to-20 times increase in "strength" to a person, enabling the person to perform physically strenuous tasks repeatedly without much fatigue. However, exo-skeletons may require more than 100W of power under certain use scenarios, so battery powered operation may be limited to 30 minutes or less. The delivery of wireless power as described herein may provide a user of an exo-skeleton with a continuous supply of power both for powering the structural movements of the exo-skeleton and for powering various monitors and sensors distributed throughout the structure. For instance, an exo-skeleton with an embedded device resonator(s) may be supplied with power from a local source resonator. For an industrial exo-skeleton, the source resonator may be placed in the walls of the facility. For a military exo-skeleton, the source resonator may be carried by an armored vehicle. For an exo-skeleton employed to assist a caretaker of the elderly, the source resonator(s) may be installed or placed in or the room(s) of a person's home.

[00467] The systems and methods described herein may be used for the powering/charging of portable medical equipment, such as oxygen systems, ventilators, defibrillators, medication pumps, monitors, and equipment in ambulances or mobile medical units, and the like. Being able to transport a patient from an accident scene to the hospital, or to move patients in their beds to other rooms or areas, and bring all the equipment that is attached with them and have it powered the whole time offers great benefits to the patients' health and eventual well-being. Certainly one can understand the risks and problems caused by medical devices that stop working because their battery dies or because they must be unplugged while a patient is transported or moved in any way. For example, an emergency medical team on the scene of an automotive accident might need to utilize portable medical equipment in the emergency care of patients in the field. Such portable medical equipment must be properly maintained so that there is sufficient battery life to power the equipment for the duration of the emergency. However, it is too often the case that the equipment is not properly maintained so that batteries are not fully charged and in some cases, necessary equipment is not available to the first responders. The systems and methods described herein may provide for wireless power to portable medical equipment (and associated sensor inputs on the patient) in such a way that the charging and maintaining of batteries and power packs is provided automatically and without human intervention. Such a system also benefits from the improved mobility of a patient unencumbered by a variety of power cords attached to the many medical monitors and devices used in their treatment.

[00468] The systems and methods described herein may be used to for the powering/charging of personal hearing aids. Personal hearing aids need to be small and light to fit into or around the ear of a person. The size and weight restrictions limit the size of batteries that can be used. Likewise, the size and weight restrictions of the device make battery replacement difficult due to the delicacy of the components. The dimensions of the devices and hygiene concerns make it difficult to integrate additional charging ports to allow recharging of the batteries. The systems and methods described herein may be integrated into the hearing aid and may reduce the size of the necessary batteries which may allow even smaller hearing aids. Using the principles disclosed herein, the batteries of the hearing aid may be recharged without requiring external connections or charging ports. Charging and device circuitry and a small rechargeable battery may be integrated into a form factor of a conventional hearing aid battery allowing retrofit into existing hearing aids. The hearing aid may be recharged while it is used and worn by a person. The energy source may be integrated into a pad or a cup allowing recharging when the hearing is placed on such a structure. The charging source may be integrated into a hearing aid dryer box allowing wireless recharging while the hearing aid is drying or being sterilized. The source and device resonator may be used to also heat the device reducing or eliminating the need for an additional heating element. Portable charging cases powered by batteries or AC adaptors may be used as storage and charging stations.

[00469] The source resonator for the medical systems described above may be in the main body of some or all of the medical equipment, with device resonators on the patient's sensors and devices; the source resonator may be in the ambulance with device resonators on the patient's sensors and the main body of some or all of the equipment; a primary source resonator may be in the ambulance for transferring wireless power to a device resonator on the medical equipment while the medical equipment is in the ambulance and a second source resonator is in the main body of the medical equipment and a second device resonator on the patient sensors when the equipment is away from the ambulance; and the like. The systems and methods described herein may significantly improve the ease with which medical personnel are able to transport patients from one location to another, where power wires and the need to replace or manually charge associated batteries may now be reduced.

[00470] The systems and methods described herein may be used for the charging of devices inside a military vehicle or facility, such as a tank, armored carrier, mobile shelter, and the like. For instance, when soldiers come back into a vehicle after "action" or a mission, they may typically start charging their electronic devices. If their electronic devices were equipped with device resonators, and there was a source resonator inside the vehicle, (e.g. integrated in the seats or on the ceiling of the vehicle), their devices would start charging immediately. In fact, the same vehicle could provide power to soldiers/robots (e.g. packbot from iRobot) standing outside or walking beside the vehicle. This capability may be useful in minimizing accidental battery-swapping with someone else (this may be a significant issue, as soldiers tend to trust only their own batteries); in enabling quicker exits from a vehicle under attack; in powering or charging laptops or other electronic devices inside a tank, as too many wires inside the tank may present a hazard in terms of reduced ability to move around fast in case of "trouble" and/or decreased visibility; and the like. The systems and methods described herein may provide a significant improvement in association with powering portable power equipment in a military environment.

[00471] The systems and methods described herein may provide wireless powering or charging capabilities to mobile vehicles such as golf carts or other types of carts, all-terrain vehicles, electric bikes, scooters, cars, mowers, bobcats and other vehicles typically used for construction and landscaping and the like. The systems and methods described herein may provide wireless powering or charging capabilities to miniature mobile vehicles, such as mini-helicopters, airborne drones, remote control planes, remote control boats, remote controlled or robotic rovers, remote controlled or robotic lawn mowers or equipment, bomb detection robots, and the like. For instance, mini-helicopter flying above a military vehicle to increase its field of view can fly for a few minutes on standard batteries. If these mini-helicopters were fitted with a device resonator, and the control vehicle had a source resonator, the mini-helicopter might be able to fly indefinitely. The systems and methods described herein may provide an effective alternative to recharging or replacing the batteries for use in miniature mobile vehicles. In addition, the systems and methods described herein may provide power/charging to even smaller devices, such as microelectromechanical systems (MEMS), nano-robots, nano devices, and the like. In addition, the systems and methods described herein may be implemented by installing a source device in a mobile vehicle or flying device to enable it to serve as an in-field or in-flight re-charger, that may position itself autonomously in proximity to a mobile vehicle that is equipped with a device resonator.

[00472] The systems and methods described herein may be used to provide power networks for temporary facilities, such as military camps, oil drilling setups, remote filming locations, and the like, where electrical power is required, such as for power generators, and where power cables are typically run around the temporary facility. There are many instances when it is necessary to set up temporary facilities that require power. The systems and methods described herein may enable a more efficient way to rapidly set up and tear down these facilities, and may reduce the number of wires that must be run throughout the faculties to supply power. For instance, when Special Forces move into an area, they may erect tents and drag many wires around the camp to provide the required electricity. Instead, the systems and methods described herein may enable an army vehicle, outfitted with a power supply and a source resonator, to park in the center of the camp, and provide all the power to nearby tents where the device resonator may be integrated into the tents, or some other piece of equipment associated with each tent or area. A series of source-device-source-device resonators may be used to extend the power to tents that are farther away. That is, the tents closest to the vehicle could then provide power to tents behind them. The systems and methods described herein may provide a significant improvement to the efficiency with which temporary installations may be set up and torn down, thus improving the mobility of the associated facility.

[00473] The systems and methods described herein may be used in vehicles, such as for replacing wires, installing new equipment, powering devices brought into the vehicle, charging the battery of a vehicle (e.g. for a traditional gas powered engine, for a hybrid car, for an electric car, and the like), powering devices mounted to the interior or exterior of the vehicle, powering devices in the vicinity of the vehicle, and the like. For example, the systems and methods described herein may be used to replace wires such as those are used to power lights, fans and sensors distributed throughout a vehicle. As an example, a typical car may have 50kg of wires associated with it, and the use of the systems and methods described herein may enable the elimination of a substantial amount of this wiring. The performance of larger and more weight sensitive vehicles such as airplanes or satellites could benefit greatly from having the number of cables that must be run throughout the vehicle reduced. The systems and methods described herein may allow the accommodation of removable or supplemental portions of a vehicle with electric and electrical devices without the need for electrical harnessing. For example, a motorcycle may have removable side boxes that act as a temporary trunk space for when the cyclist is going on a long trip. These side boxes may have exterior lights, interior lights, sensors, auto equipment, and the like, and if not for being equipped with the systems and methods described herein might require electrical connections and harnessing.

[00474] An in-vehicle wireless power transmission system may charge or power one or more mobile devices used in a car: mobile phone handset, Bluetooth headset, blue tooth hands free speaker phone, GPS, MP3 player, wireless audio transceiver for streaming MP3 audio through car stereo via FM, Bluetooth, and the like. The in vehicle wireless power source may utilize source resonators that are arranged in any of several possible configurations including charging pad on dash, charging pad otherwise mounted on floor, or between seat and center console, charging "cup" or receptacle that fits in cup holder or on dash, and the like.

[00475] The wireless power transmission source may utilize a rechargeable battery system such that said supply battery gets charged whenever the vehicle power is on such that when the vehicle is turned off the wireless supply can draw power from the supply battery and can continue to wirelessly charge or power mobile devices that are still in the car.

[00476] The plug-in electric cars, hybrid cars, and the like, of the future need to be charged, and the user may need to plug in to an electrical supply when they get home or to a charging station. Based on a single over-night recharging, the user may be able to drive up to 50 miles the next day. Therefore, in the instance of a hybrid car, if a person drives less than 50 miles on most days, they will be driving mostly on electricity. However, it would be beneficial if they didn't have to remember to plug in the car at night. That is, it would be nice to simply drive into a garage, and have the car take care of its own charging. To this end, a source resonator may be built into the garage floor and/or garage side-wall, and the device resonator may be built into the bottom (or side) of the car. Even a few kW transfer may be sufficient to recharge the car over-night. The in-vehicle device resonator may measure magnetic field properties to provide feedback to assist in vehicle (or any similar device) alignment to a stationary resonating source. The vehicle may use this positional feedback to automatically position itself to achieve optimum alignment, thus optimum power transmission efficiency. Another method may be to use the positional feedback to help the human operator to properly position the vehicle or device, such as by making LED's light up, providing noises, and the like when it is well positioned. In such cases where the amount of power being transmitted could present a safety hazard to a person or animal that intrudes into the active field volume, the source or receiver device may be equipped with an active light curtain or some other external device capable of sensing intrusion into the active field volume, and capable of shutting off the source device and alert a human operator. In addition, the source device may be equipped with self-sensing capability such that it may detect that its expected power transmission rate has been interrupted by an intruding element, and in such case shut off the source device and alert a human operator. Physical or mechanical structures such as hinged doors or inflatable bladder shields may be incorporated as a physical barrier to prevent unwanted intrusions. Sensors such as optical, magnetic, capacitive, inductive, and the like may also be used to detect foreign structures or interference between the source and device resonators. The shape of the source resonator may be shaped such to prevent water or debris accumulation. The source resonator may be placed in a cone shaped enclosure or may have an enclosure with an angled top to allow water and debris to roll off. The source of the system may use battery power of the vehicle or its own battery power to transmit its presence to the source to initiate power transmission.

[00477] The source resonator may be mounted on an embedded or hanging post, on a wall, on a stand, and the like for coupling to a device resonator mounted on the bumper, hood, body panel, and the like, of an electric vehicle. The source resonator may be enclosed or embedded into a flexible enclosure such as a pillow, a pad, a bellows, a spring loaded enclosure and the like so that the electric vehicle may make contact with the structure containing the source coil without damaging the car in any way. The structure containing the source may prevent objects from getting between the source and device resonators. Because the wireless power transfer may be relatively insensitive to misalignments between the source and device coils, a variety of flexible source structures and parking procedures may be appropriate for this application.

[00478] The systems and methods described herein may be used to trickle charge batteries of electric, hybrid or combustion engine vehicles. Vehicles may require small amounts of power to maintain or replenish battery power. The power may be transferred wirelessly from a source to a device resonator that may be incorporated into the front grill, roof, bottom, or other parts of the vehicle. The device resonator may be designed to fit into a shape of a logo on the front of a vehicle or around the grill as not to obstruct air flow through the radiator. The device or source resonator may have additional modes of operation that allow the resonator to be used as a heating element which can be used to melt of snow or ice from the vehicle.

[00479] An electric vehicle or hybrid vehicle may require multiple device resonators, such as to increase the ease with which the vehicle may come in proximity with a source resonator for charging (i.e. the greater the number and varied position of device resonators are, the greater the chances that the vehicle can pull in and interface with a diversity of charging stations), to increase the amount of power that can be delivered in a period of time (e.g. additional device resonators may be required to keep the local heating due to charging currents to acceptable levels), to aid in automatic parking/docking the vehicle with the charging station, and the like. For example, the vehicle may have multiple resonators (or a single resonator) with a feedback system that provides guidance to either the driver or an automated parking/docking facility in the parking of the vehicle for optimized charging conditions (i.e., the optimum positioning of the vehicle's device resonator to the charging station's source resonator may provide greater power transfer efficiency). An automated parking/docking facility may allow for the automatic parking of the vehicle based on how well the vehicle is coupled.

[00480] The power transmission system may be used to power devices and peripherals of a vehicle. Power to peripherals may be provided while a vehicle is charging, or while not charging, or power may be delivered to conventional vehicles that do not need charging. For example, power may be transferred wirelessly to conventional non-electric cars to power air conditioning, refrigeration units, heaters, lights, and the like while parked to avoid running the engine which may be important to avoid exhaust build up in garage parking lots or loading docks. Power may for example be wirelessly transferred to a bus while it is parked to allow powering of lights, peripherals, passenger devices, and the like avoiding the use of onboard engines or power sources. Power may be wirelessly transferred to an airplane while parked on the tarmac or in a hanger to power instrumentation, climate control, de-icing equipment, and the like without having to use onboard engines or power sources.

[00481] Wireless power transmission on vehicles may be used to enable the concept of Vehicle to Grid (V2G). Vehicle to grid is based on utilizing electric vehicles and plug-in hybrid electric vehicles (PHEV) as distributed energy storage devices, charged at night when the electric grid is underutilized, and available to discharge back into the grid during episodes of peak demand that occur during the day. The wireless power transmission system on a vehicle and the respective infrastructure may be implemented in such a way as to enable bidirectional energy flow—so that energy can flow back into the grid from the vehicle—without requiring a plug in connection. Vast fleets of vehicles, parked at factories, offices, parking lots, can be viewed as "peaking power capacity" by the smart grid. Wireless power transmission on vehicles can make such a V2G vision a reality. By simplifying the process of connecting a vehicle to the grid, (i.e. by simply parking it in a wireless charging enabled parking spot), it becomes much more likely that a certain number of vehicles will be "dispatchable" when the grid needs to tap their power. Without wireless charging, electric and PHEV owners will likely charge their vehicles at home, and park them at work in conventional parking spots. Who will want to plug their vehicle in at work, if they do not need charging? With wireless charging systems capable of handling 3 kW, 100,000 vehicles can provide 300 Megawatts back to the grid—using energy generated the night before by cost effective base load generating capacity. It is the streamlined ergonomics of the cordless self charging PHEV and electric vehicles that make it a viable V2G energy source.

[00482] The systems and methods described herein may be used to power sensors on the vehicle, such as sensors in tires to measure air-pressure, or to run peripheral devices in the vehicle, such as cell phones, GPS devices, navigation devices, game players, audio or video players, DVD players, wireless routers, communications equipment, anti-theft devices, radar devices, and the like. For example, source resonators described herein may be built into the main compartment of the car in order to supply power to a variety of devices located both inside and outside of the main compartment of the car. Where the vehicle is a motorcycle or the like, devices described herein may be integrated into the body of the motorcycle, such as under the seat, and device resonators may be provided in a user's helmet, such as for communications, entertainment, signaling, and the like, or device resonators may be provided in the user's jacket, such as for displaying signals to other drivers for safety, and the like.

[00483] The systems and methods described herein may be used in conjunction with transportation infrastructure, such as roads, trains, planes, shipping, and the like. For example, source resonators may be built into roads, parking lots, rail-lines, and the like. Source resonators may be built into traffic lights, signs, and the like. For example, with source resonators embedded into a road, and device resonators built into vehicles, the vehicles may be provided power as they drive along the road or as they are parked in lots or on the side of the road. The systems and methods described herein may provide an effective way for electrical systems in vehicles to be powered and/or charged while the vehicle traverses a road network, or a portion of a road network. In this way, the systems and methods described herein may contribute to the powering/charging of autonomous vehicles, automatic guided vehicles, and the like. The systems and methods described herein may provide power to vehicles in places where they typically idle or stop, such as in the vicinity of traffic lights or signs, on highway ramps, or in parking lots.

[00484] The systems and methods described herein may be used in an industrial environment, such as inside a factory for powering machinery, powering/charging robots, powering and/or charging wireless sensors on robot arms, powering/charging tools and the like. For example, using the systems and methods described herein to supply power to devices on the arms of robots may help eliminate direct wire connections across the joints of the robot arm. In this way, the wearing out of such direct wire connections may be reduced, and the reliability of the robot increased. In this case, the device resonator may be out on the arm of the robot, and the source resonator may be at the base of the robot, in a central location near the robot, integrated into the industrial facility in which the robot is providing service, and the like. The use of the systems and methods described herein may help eliminate wiring otherwise associated with power distribution within the industrial facility, and thus benefit the overall reliability of the facility.

[00485] The systems and methods described herein may be used for underground applications, such as drilling, mining, digging, and the like. For example, electrical components and sensors associated with drilling or excavation may utilize the systems and methods described herein to eliminate cabling associated with a digging mechanism, a drilling bit, and the like, thus eliminating or minimizing cabling near the excavation point. In another example, the systems and methods described herein may be used to provide power to excavation equipment in a mining application where the power requirements for the equipment may be high and the distances large, but where there are no people to be subjected to the associated required fields. For instance, the excavation area may have device resonator powered digging equipment that has high power requirements and may be digging relatively far from the source resonator. As a result the source resonator may need to provide high field intensities to satisfy these requirements, but personnel are far enough away to be outside these high intensity fields. This high power, no personnel, scenario may be applicable to a plurality of industrial applications.

[00486] The systems and methods described herein may also use the near-field non-radiative resonant scheme for information transfer rather than, or in addition to, power transfer. For instance, information being transferred by near-field non-radiative resonance techniques may not be susceptible to eavesdropping and so may provide an increased level of security compared to traditional wireless communication schemes. In addition, information being transferred by near-field non-radiative resonance techniques may not interfere with the EM radiative spectrum and so may not be a source of EM interference, thereby allowing communications in an extended frequency range and well within the limits set by any regulatory bodies. Communication services may be provided between remote, inaccessible or hard-to-reach places such as between remote sensors, between sections of a device or vehicle, in tunnels, caves and wells (e.g. oil wells, other drill sites) and between underwater or underground devices, and the like. Communications services may be provided in places where magnetic fields experience less loss than electric fields.

[00487] The systems and methods described herein may enable the simultaneous transmission of power and communication signals between sources and devices in wireless power transmission systems, or it may enable the transmission of power and communication signals during different time periods or at different frequencies. The performance characteristics of the resonator may be controllably varied to preferentially support or limit the efficiency or range of either energy or information transfer. The performance characteristics of the resonators may be controlled to improve the security by reducing the range of information transfer, for example. The performance characteristics of the resonators may be varied continuously, periodically, or according to a predetermined, computed or automatically adjusted algorithm. For example, the power and information transfer enabled by the systems and methods described herein may be provided in a time multiplexed or frequency multiplexed manner. A source and device may signal each other by tuning, changing, varying, dithering, and the like, the resonator impedance which may affect the reflected impedance of other resonators that can be detected. The information transferred as described herein may include information regarding device identification, device power requirements, handshaking protocols, and the like.

[00488] The source and device may sense, transmit, process and utilize position and location information on any other sources and/or devices in a power network. The source and device may capture or use information such as elevation, tilt, latitude and longitude, and the like from a variety of sensors and sources that may be built into the source and device or may be part of a component the source or device connect. The positioning and orientation information may include sources such as global positioning sensors (GPS), compasses, accelerometers, pressure sensors, atmospheric barometric sensors, positioning systems which use Wi-Fi or cellular network signals, and the like. The source and device may use the position and location information to find nearby wireless power transmission sources. A source may broadcast or communicate with a central station or database identifying its location. A device may obtain the source location information from the central station or database or from the local broadcast and guide a user or an operator to the source with the aid of visual, vibrational, or auditory signals. Sources and devices may be nodes in a power network, in a communications network, in a sensor network, in a navigational network, and the like or in kind of combined functionality network.

[00489] The position and location information may also be used to optimize or coordinate power delivery. Additional information about the relative position of a source and a device may be used to optimize magnetic field direction and resonator alignment. The orientation of a device and a source which may be obtained from accelerometers and magnetic sensors, and the like, for example, may be used to identify the orientation of resonators and the most favorable direction of a magnetic field such that the magnetic flux is not blocked by the device circuitry. With such information a source with the most favorable orientation, or a combination of sources, may be used. Likewise, position and orientation information may be used to move or provide feedback to a user or operator of a device to place a device in a favorable orientation or location to maximize power transmission efficiency, minimize losses, and the like.

[00490] The source and device may include power metering and measuring circuitry and capability. The power metering may be used to track how much power was delivered to a device or how much power was transferred by a source. The power metering and power usage information may be used in fee based power delivery arrangements for billing purposes. Power metering may be also be used to enable power delivery policies to ensure power is distributed to multiple devices according to specific criteria. For example, the power metering may be used to categorize devices based on the amount of power they received and priority in power delivery may be given to those having received the least power. Power metering may be used to provide tiered delivery services such as “guaranteed power” and “best effort power” which may be billed at separate rates. Power metering may be used to institute and enforce hierarchical power delivery structures and may enable priority devices to demand and receive- more power under certain circumstances or use scenarios.

[00491] Power metering may be used to optimize power delivery efficiency and minimize absorption and radiation losses. Information related to the power received by devices may be used by a source in conjunction with information about the power output of the source to identify unfavorable operating environments or frequencies. For example, a source may compare the amount of power which was received by the devices and the amount of power which it transmitted to determine if the transmission losses may be unusually or unacceptably large. Large transmission losses may be due to an unauthorized device receiving power from the source and the source and other devices may initiate frequency hopping of the resonance frequency or other defensive measures to prevent or deter unauthorized use. Large transmission losses may be due to absorption losses for example, and the device and source may tune to alternate resonance frequencies to minimize such losses. Large transmission losses may also indicate the presence of unwanted or unknown objects or materials and the source may turn down or off its power level until the unwanted or unknown object is removed or identified, at which point the source may resume powering remote devices.

[00492] The source and device may include authentication capability. Authentication may be used to ensure that only compatible sources and devices are able to transmit and receive power. Authentication may be used to ensure that only authentic devices that are of a specific manufacturer and not clones or devices and sources from other manufacturers, or only devices that are part of a specific subscription or plan, are able to receive power from a source. Authentication may be based on cryptographic request and respond protocols or it may be based on the unique signatures of perturbations of specific devices allowing them to be used and authenticated based on properties similar to physically unclonable functions. Authentication may be performed locally between each source and device with local communication or it may be used with third person authentication methods where the source and device authenticate with communications to a central authority. Authentication protocols may use position information to alert a local source or sources of a genuine device.

[00493] The source and device may use frequency hopping techniques to prevent unauthorized use of a wireless power source. The source may continuously adjust or change the resonant frequency of power delivery. The changes in frequency may be performed in a pseudorandom or predetermined manner that is known, reproducible, or communicated to authorized device but difficult to predict. The rate of frequency hopping and the number of various frequencies used may be large and frequent enough to ensure that unauthorized use is difficult or impractical. Frequency hopping may be implemented by tuning the impedance network, tuning any of the driving circuits, using a plurality of resonators tuned or tunable to multiple resonant frequencies, and the like.

[00494] The source may have a user notification capability to show the status of the source as to whether it is coupled to a device resonator and transmitting power, if it is in standby mode, or if the source resonator is detuned or perturbed by an external object. The notification capability may include visual, auditory, and vibrational methods. The notification may be as simple as three color lights, one for each state, and optionally a speaker to provide notification in case of an error in operation. Alternatively, the notification capability may involve an interactive display that shows the status of the source and optionally provides instructions on how to fix or solve any errors or problems identified.

[00495] As another example, wireless power transfer may be used to improve the safety of electronic explosive detonators. Explosive devices are detonated with an electronic detonator, electric detonator, or shock tube detonator. The electronic detonator utilizes stored electrical energy (usually in a capacitor) to activate the igniter charge, with a low energy trigger signal transmitted conductively or by radio. The electric detonator utilizes a high energy conductive trigger signal to provide both the signal and the energy required to activate the igniter charge. A shock tube sends a controlled explosion through a hollow tube coated with explosive from the generator to the igniter charge. There are safety issues associated with the electric and electronic detonators, as there are cases of stray electromagnetic energy causing unintended activation. Wireless power transfer via sharply resonant magnetic coupling can improve the safety of such systems.

[00496] Using the wireless power transfer methods disclosed herein, one can build an electronic detonation system that has no locally stored energy, thus reducing the risk of unintended activation. A wireless power source can be placed in proximity (within a few meters) of the detonator. The detonator can be equipped with a resonant capture coil. The activation energy can be transferred when the wireless power source has been triggered. The triggering of the wireless power source can be initiated by any number of mechanisms: radio, magnetic near field radio, conductive signaling, ultrasonics, laser light. Wireless power transfer based on resonant magnetic coupling also has the benefit of being able to transfer power through materials such as rock, soil, concrete, water, and other dense materials. The use of very high Q coils as receivers and sources, having very narrow band response and sharply tuned to proprietary frequencies, further ensure that the detonator circuits cannot capture stray EMI and activate unintentionally.

[00497] The resonator of a wirelessly powered device may be external, or outside of the device, and wired to the battery of the device. The battery of the device may be modified to include appropriate rectification and control circuitry to receive the alternating currents of the device resonator. This can enable configurations with larger external coils, such as might be built into a battery door of a keyboard or mouse, or digital still camera, or even larger coils that are attached to the device but wired back to the battery/converter with ribbon cable. The battery door can be modified to provide interconnection from the external coil to the battery/converter (which will need an exposed contact that can touch the battery door contacts.

[00498] While the invention has been described in connection with certain preferred embodiments, other embodiments will be understood by one of ordinary skill in the art and are intended to fall within the scope of this disclosure, which is to be interpreted in the broadest sense allowable by law.

[00499] All documents referenced herein are hereby incorporated by reference.

ABSTRACT

[00500] In embodiments of the present invention improved capabilities are described for a method and system comprising a source resonator optionally coupled to an energy source and a second resonator located a distance from the source resonator, where the source resonator and the second resonator are coupled to provide near-field wireless energy transfer among the source resonator and the second resonator and where the field of at least one of the source resonator and the second resonator is shaped to avoid a loss-inducing object.

WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS

CROSS-REFERENCE TO RELATED APPLICATIONS

[0001] This application is a continuation-in-part of the following U.S. patent application, U.S. 12/567716 filed September 25, 2009 which claims the benefit of the following U.S. provisional applications, U.S. App. No. 61/100,721 filed September 27,2008; U.S. App. No. 61/108,743 filed October 27, 2008; U.S. App. No. 61/147,386 filed January 26, 2009; U.S. App. No. 61/152,086 filed February 12, 2009; U.S. App. No. 61/178,508 filed May 15, 2009; U.S. App. No. 61/182,768 filed June 1, 2009; U.S. App. No. 61/121,159 filed December 9, 2008; U.S. App. No. 61/142,977 filed January 7, 2009; U.S. App. No. 61/142,885 filed January 6, 2009; U.S. App. No. 61/142,796 filed January 6, 2009; U.S. App. No. 61/142,889 filed January 6, 2009; U.S. App. No. 61/142,880 filed January 6, 2009; U.S. App. No. 61/142,818 filed January 6, 2009; U.S. App. No. 61/142,887 filed January 6, 2009; U.S. App. No. 61/156,764 filed March 2, 2009; U.S. App. No. 61/143,058 filed January 7, 2009; U.S. App. No. 61/152,390 filed February 13, 2009; U.S. App. No. 61/163,695 filed March 26, 2009; U.S. App. No. 61/172,633 filed April 24, 2009; U.S. App. No. 61/169,240 filed April 14, 2009, U.S. App. No. 61/173,747 filed April 29, 2009.

[0002] Each of the foregoing applications is incorporated herein by reference in its entirety.

BACKGROUND

[0003] Field:

[0004] This disclosure relates to wireless energy transfer, also referred to as wireless power transmission.

[0005] Description of the Related Art:

[0006] Energy or power may be transferred wirelessly using a variety of known radiative, or far-field, and non-radiative, or near-field, techniques. For example, radiative wireless information transfer using low-directionality antennas, such as those used in radio and cellular communications systems and home computer networks, may be considered wireless energy transfer. However, this type of radiative transfer is very inefficient because only a tiny

portion of the supplied or radiated power, namely, that portion in the direction of, and overlapping with, the receiver is picked up. The vast majority of the power is radiated away in all the other directions and lost in free space. Such inefficient power transfer may be acceptable for data transmission, but is not practical for transferring useful amounts of electrical energy for the purpose of doing work, such as for powering or charging electrical devices. One way to improve the transfer efficiency of some radiative energy transfer schemes is to use directional antennas to confine and preferentially direct the radiated energy towards a receiver. However, these directed radiation schemes may require an uninterruptible line-of-sight and potentially complicated tracking and steering mechanisms in the case of mobile transmitters and/or receivers. In addition, such schemes may pose hazards to objects or people that cross or intersect the beam when modest to high amounts of power are being transmitted. A known non-radiative, or near-field, wireless energy transfer scheme, often referred to as either induction or traditional induction, does not (intentionally) radiate power, but uses an oscillating current passing through a primary coil, to generate an oscillating magnetic near-field that induces currents in a near-by receiving or secondary coil. Traditional induction schemes have demonstrated the transmission of modest to large amounts of power, however only over very short distances, and with very small offset tolerances between the primary power supply unit and the secondary receiver unit. Electric transformers and proximity chargers are examples of devices that utilize this known short range, near-field energy transfer scheme.

[0007] Therefore a need exists for a wireless power transfer scheme that is capable of transferring useful amounts of electrical power over mid-range distances or alignment offsets. Such a wireless power transfer scheme should enable useful energy transfer over greater distances and alignment offsets than those realized with traditional induction schemes, but without the limitations and risks inherent in radiative transmission schemes.

SUMMARY

[0008] There is disclosed herein a non-radiative or near-field wireless energy transfer scheme that is capable of transmitting useful amounts of power over mid-range distances and alignment offsets. This inventive technique uses coupled electromagnetic resonators with long-lived oscillatory resonant modes to transfer power from a power supply to a power drain. The technique is general and may be applied to a wide range of resonators, even where the specific

examples disclosed herein relate to electromagnetic resonators. If the resonators are designed such that the energy stored by the electric field is primarily confined within the structure and that the energy stored by the magnetic field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant magnetic near-field. These types of resonators may be referred to as magnetic resonators. If the resonators are designed such that the energy stored by the magnetic field is primarily confined within the structure and that the energy stored by the electric field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant electric near-field. These types of resonators may be referred to as electric resonators. Either type of resonator may also be referred to as an electromagnetic resonator. Both types of resonators are disclosed herein.

[0009] The omni-directional but stationary (non-lossy) nature of the near-fields of the resonators we disclose enables efficient wireless energy transfer over mid-range distances, over a wide range of directions and resonator orientations, suitable for charging, powering, or simultaneously powering and charging a variety of electronic devices. As a result, a system may have a wide variety of possible applications where a first resonator, connected to a power source, is in one location, and a second resonator, potentially connected to electrical/electronic devices, batteries, powering or charging circuits, and the like, is at a second location, and where the distance from the first resonator to the second resonator is on the order of centimeters to meters. For example, a first resonator connected to the wired electricity grid could be placed on the ceiling of a room, while other resonators connected to devices, such as robots, vehicles, computers, communication devices, medical devices, and the like, move about within the room, and where these devices are constantly or intermittently receiving power wirelessly from the source resonator. From this one example, one can imagine many applications where the systems and methods disclosed herein could provide wireless power across mid-range distances, including consumer electronics, industrial applications, infrastructure power and lighting, transportation vehicles, electronic games, military applications, and the like.

[0010] Energy exchange between two electromagnetic resonators can be optimized when the resonators are tuned to substantially the same frequency and when the losses in the system are minimal. Wireless energy transfer systems may be designed so that the "coupling-time" between resonators is much shorter than the resonators' "loss-times". Therefore, the systems and methods described herein may utilize high quality factor (high-Q) resonators with

low intrinsic-loss rates. In addition, the systems and methods described herein may use sub-wavelength resonators with near-fields that extend significantly longer than the characteristic sizes of the resonators, so that the near-fields of the resonators that exchange energy overlap at mid-range distances. This is a regime of operation that has not been practiced before and that differs significantly from traditional induction designs.

[0011] It is important to appreciate the difference between the high-Q magnetic resonator scheme disclosed here and the known close-range or proximity inductive schemes, namely, that those known schemes do not conventionally utilize high-Q resonators. Using coupled-mode theory (CMT), (see, for example, *Waves and Fields in Optoelectronics*, H.A. Haus, Prentice Hall, 1984), one may show that a high-Q resonator-coupling mechanism can enable orders of magnitude more efficient power delivery between resonators spaced by mid-range distances than is enabled by traditional inductive schemes. Coupled high-Q resonators have demonstrated efficient energy transfer over mid-range distances and improved efficiencies and offset tolerances in short range energy transfer applications.

[0012] The systems and methods described herein may provide for near-field wireless energy transfer via strongly coupled high-Q resonators, a technique with the potential to transfer power levels from picowatts to kilowatts, safely, and over distances much larger than have been achieved using traditional induction techniques. Efficient energy transfer may be realized for a variety of general systems of strongly coupled resonators, such as systems of strongly coupled acoustic resonators, nuclear resonators, mechanical resonators, and the like, as originally described by researchers at M.I.T. in their publications, "Efficient wireless non-radiative mid-range energy transfer", *Annals of Physics*, vol. 323, Issue 1, p. 34 (2008) and "Wireless Power Transfer via Strongly Coupled Magnetic Resonances", *Science*, vol. 317, no. 5834, p. 83, (2007). Disclosed herein are electromagnetic resonators and systems of coupled electromagnetic resonators, also referred to more specifically as coupled magnetic resonators and coupled electric resonators, with operating frequencies below 10 GHz.

[0013] This disclosure describes wireless energy transfer technologies, also referred to as wireless power transmission technologies. Throughout this disclosure, we may use the terms wireless energy transfer, wireless power transfer, wireless power transmission, and the like, interchangeably. We may refer to supplying energy or power from a source, an AC or DC source, a battery, a source resonator, a power supply, a generator, a solar panel, and thermal

collector, and the like, to a device, a remote device, to multiple remote devices, to a device resonator or resonators, and the like. We may describe intermediate resonators that extend the range of the wireless energy transfer system by allowing energy to hop, transfer through, be temporarily stored, be partially dissipated, or for the transfer to be mediated in any way, from a source resonator to any combination of other device and intermediate resonators, so that energy transfer networks, or strings, or extended paths may be realized. Device resonators may receive energy from a source resonator, convert a portion of that energy to electric power for powering or charging a device, and simultaneously pass a portion of the received energy onto other device or mobile device resonators. Energy may be transferred from a source resonator to multiple device resonators, significantly extending the distance over which energy may be wirelessly transferred. The wireless power transmission systems may be implemented using a variety of system architectures and resonator designs. The systems may include a single source or multiple sources transmitting power to a single device or multiple devices. The resonators may be designed to be source or device resonators, or they may be designed to be repeaters. In some cases, a resonator may be a device and source resonator simultaneously, or it may be switched from operating as a source to operating as a device or a repeater. One skilled in the art will understand that a variety of system architectures may be supported by the wide range of resonator designs and functionalities described in this application.

[0014] In the wireless energy transfer systems we describe, remote devices may be powered directly, using the wirelessly supplied power or energy, or the devices may be coupled to an energy storage unit such as a battery, a super-capacitor, an ultra-capacitor, or the like (or other kind of power drain), where the energy storage unit may be charged or re-charged wirelessly, and/or where the wireless power transfer mechanism is simply supplementary to the main power source of the device. The devices may be powered by hybrid battery/energy storage devices such as batteries with integrated storage capacitors and the like. Furthermore, novel battery and energy storage devices may be designed to take advantage of the operational improvements enabled by wireless power transmission systems.

[0015] Other power management scenarios include using wirelessly supplied power to recharge batteries or charge energy storage units while the devices they power are turned off, in an idle state, in a sleep mode, and the like. Batteries or energy storage units may be charged or recharged at high (fast) or low (slow) rates. Batteries or energy storage units may be trickle

charged or float charged. Multiple devices may be charged or powered simultaneously in parallel or power delivery to multiple devices may be serialized such that one or more devices receive power for a period of time after which other power delivery is switched to other devices. Multiple devices may share power from one or more sources with one or more other devices either simultaneously, or in a time multiplexed manner, or in a frequency multiplexed manner, or in a spatially multiplexed manner, or in an orientation multiplexed manner, or in any combination of time and frequency and spatial and orientation multiplexing. Multiple devices may share power with each other, with at least one device being reconfigured continuously, intermittently, periodically, occasionally, or temporarily, to operate as wireless power sources. It would be understood by one of ordinary skill in the art that there are a variety of ways to power and/or charge devices, and the variety of ways could be applied to the technologies and applications described herein.

[0016] Wireless energy transfer has a variety of possible applications including for example, placing a source (e.g. one connected to the wired electricity grid) on the ceiling, under the floor, or in the walls of a room, while devices such as robots, vehicles, computers, PDAs or similar are placed or move freely within the room. Other applications may include powering or recharging electric-engine vehicles, such as buses and/or hybrid cars and medical devices, such as wearable or implantable devices. Additional example applications include the ability to power or recharge autonomous electronics (e.g. laptops, cell-phones, portable music players, house-hold robots, GPS navigation systems, displays, etc), sensors, industrial and manufacturing equipment, medical devices and monitors, home appliances and tools (e.g. lights, fans, drills, saws, heaters, displays, televisions, counter-top appliances, etc.), military devices, heated or illuminated clothing, communications and navigation equipment, including equipment built into vehicles, clothing and protective-wear such as helmets, body armor and vests, and the like, and the ability to transmit power to physically isolated devices such as to implanted medical devices, to hidden, buried, implanted or embedded sensors or tags, to and/or from roof-top solar panels to indoor distribution panels, and the like.

[0017] In one aspect, disclosed herein is a system including a source resonator having a Q-factor Q_1 and a characteristic size $_{x1}$, coupled to a power generator with direct electrical connections; and a second resonator having a Q-factor Q_2 and a characteristic size x_2, coupled to a load with direct electrical connections, and located a distance D from the source resonator,

wherein the source resonator and the second resonator are coupled to exchange energy wirelessly among the source resonator and the second resonator in order to transmit power from the power generator to the load, and wherein $\sqrt{Q_1 Q_2}$ is greater than 100.

[0018] Q_1 may be greater than 100 and Q_2 may be less than 100. Q_1 may be greater than 100 and Q_2 may be greater than 100. A useful energy exchange may be maintained over an operating distance from 0 to D, where D is larger than the smaller of x_1 and x_2. At least one of the source resonator and the second resonator may be a coil of at least one turn of a conducting material connected to a first network of capacitors. The first network of capacitors may include at least one tunable capacitor. The direct electrical connections of at least one of the source resonator to the ground terminal of the power generator and the second resonator to the ground terminal of the load may be made at a point on an axis of electrical symmetry of the first network of capacitors. The first network of capacitors may include at least one tunable butterfly-type capacitor, wherein the direct electrical connection to the ground terminal is made on a center terminal of the at least one tunable butterfly-type capacitor. The direct electrical connection of at least one of the source resonator to the power generator and the second resonator to the load may be made via a second network of capacitors, wherein the first network of capacitors and the second network of capacitors form an impedance matching network. The impedance matching network may be designed to match the coil to a characteristic impedance of the power generator or the load at a driving frequency of the power generator.

[0019] At least one of the first network of capacitors and the second network of capacitors may include at least one tunable capacitor. The first network of capacitors and the second network of capacitors may be adjustable to change an impedance of the impedance matching network at a driving frequency of the power generator. The first network of capacitors and the second network of capacitors may be adjustable to match the coil to the characteristic impedance of the power generator or the load at a driving frequency of the power generator. At least one of the first network of capacitors and the second network of capacitors may include at least one fixed capacitor that reduces a voltage across the at least one tunable capacitor. The direct electrical connections of at least one of the source resonator to the power generator and the second resonator to the load may be configured to substantially preserve a resonant mode. At least one of the source resonator and the second resonator may be a tunable resonator. The source resonator may be physically separated from the power generator and the second resonator

may be physically separated from the load. The second resonator may be coupled to a power conversion circuit to deliver DC power to the load. The second resonator may be coupled to a power conversion circuit to deliver AC power to the load. The second resonator may be coupled to a power conversion circuit to deliver both AC and DC power to the load. The second resonator may be coupled to a power conversion circuit to deliver power to a plurality of loads.

[0020] In another aspect, a system disclosed herein includes a source resonator having a *Q*-factor Q_1 and a characteristic size x_1, and a second resonator having a *Q*-factor Q_2 and a characteristic size x_2, and located a distance D from the source resonator; wherein the source resonator and the second resonator are coupled to exchange energy wirelessly among the source resonator and the second resonator; and wherein $\sqrt{Q_1 Q_2}$ is greater than 100, and wherein at least one of the resonators is enclosed in a low loss tangent material.

[0021] In another aspect, a system disclosed herein includes a source resonator having a *Q*-factor Q_1 and a characteristic size x_1, and a second resonator having a *Q*-factor Q_2 and a characteristic size x_2, and located a distance D from the source resonator; wherein the source resonator and the second resonator are coupled to exchange energy wirelessly among the source resonator and the second resonator, and wherein $\sqrt{Q_1 Q_2}$ is greater than 100; and wherein at least one of the resonators includes a coil of a plurality of turns of a conducting material connected to a network of capacitors, wherein the plurality of turns are in a common plane, and wherein a characteristic thickness of the at least one of the resonators is much less than a characteristic size of the at least one of the resonators.

[0022] In embodiments, the present invention may provide for a method and system comprising a source resonator optionally coupled to an energy source and a second resonator located a distance from the source resonator, where the source resonator and the second resonator are coupled to provide near-field wireless energy transfer among the source resonator and the second resonator and where the field of at least one of the source resonator and the second resonator is shaped to avoid a loss-inducing object. In embodiments, at least one of the source resonator and the second resonator may have a quality factor, Q>100. The source resonator Q may be greater than 100 and the second resonator Q may be greater than 100. The square root of the source resonator Q times the second resonator Q may be greater than 100. In embodiments, there may be more than one source resonator, more than one second resonator, more than three resonators, and the like.

[0023] Throughout this disclosure we may refer to the certain circuit components such as capacitors, inductors, resistors, diodes, switches and the like as circuit components or elements. We may also refer to series and parallel combinations of these components as elements, networks, topologies, circuits, and the like. We may describe combinations of capacitors, diodes, varactors, transistors, and/or switches as adjustable impedance networks, tuning networks, matching networks, adjusting elements, and the like. We may also refer to "self-resonant" objects that have both capacitance, and inductance distributed (or partially distributed, as opposed to solely lumped) throughout the entire object. It would be understood by one of ordinary skill in the art that adjusting and controlling variable components within a circuit or network may adjust the performance of that circuit or network and that those adjustments may be described generally as tuning, adjusting, matching, correcting, and the like. Other methods to tune or adjust the operating point of the wireless power transfer system may be used alone, or in addition to adjusting tunable components such as inductors and capacitors, or banks of inductors and capacitors.

[0024] Unless otherwise defined, all technical and scientific terms used herein have the same meaning as commonly understood by one of ordinary skill in the art to which this disclosure belongs. In case of conflict with publications, patent applications, patents, and other references mentioned or incorporated herein by reference, the present specification, including definitions, will control.

[0025] Any of the features described above may be used, alone or in combination, without departing from the scope of this disclosure. Other features, objects, and advantages of the systems and methods disclosed herein will be apparent from the following detailed description and figures.

BRIEF DESCRIPTION OF FIGURES

[0026] Fig. 1 (a) and (b) depict exemplary wireless power systems containing a source resonator 1 and device resonator 2 separated by a distance D.

[0027] Fig. 2 shows an exemplary resonator labeled according to the labeling convention described in this disclosure. Note that there are no extraneous objects or additional resonators shown in the vicinity of resonator 1.

[0028] Fig. 3 shows an exemplary resonator in the presence of a “loading” object, labeled according to the labeling convention described in this disclosure.

[0029] Fig. 4 shows an exemplary resonator in the presence of a “perturbing” object, labeled according to the labeling convention described in this disclosure.

[0030] Fig. 5 shows a plot of efficiency, η, vs. strong coupling factor, $U=\kappa/\sqrt{\Gamma_s\Gamma_d}=k\sqrt{Q_sQ_d}$.

[0031] Fig. 6 (a) shows a circuit diagram of one example of a resonator (b) shows a diagram of one example of a capacitively-loaded inductor loop magnetic resonator, (c) shows a drawing of a self-resonant coil with distributed capacitance and inductance, (d) shows a simplified drawing of the electric and magnetic field lines associated with an exemplary magnetic resonator of the current disclosure, and (e) shows a diagram of one example of an electric resonator.

[0032] Fig. 7 shows a plot of the “quality factor”, Q (solid line), as a function of frequency, of an exemplary resonator that may be used for wireless power transmission at MHz frequencies. The absorptive Q (dashed line) increases with frequency, while the radiative Q (dotted line) decreases with frequency, thus leading the overall Q to peak at a particular frequency.

[0033] Fig. 8 shows a drawing of a resonator structure with its characteristic size, thickness and width indicated.

[0034] Fig. 9 (a) and (b) show drawings of exemplary inductive loop elements.

[0035] Fig. 10 (a) and (b) show two examples of trace structures formed on printed circuit boards and used to realize the inductive element in magnetic resonator structures.

[0036] Fig. 11 (a) shows a perspective view diagram of a planar magnetic resonator, (b) shows a perspective view diagram of a two planar magnetic resonator with various geometries, and c) shows is a perspective view diagram of a two planar magnetic resonators separated by a distance D.

[0037] Fig. 12 is a perspective view of an example of a planar magnetic resonator.

[0038] Fig. 13 is a perspective view of a planar magnetic resonator arrangement with a circular resonator coil.

[0039] Fig. 14 is a perspective view of an active area of a planar magnetic resonator.

[0040] Fig. 15 is a perspective view of an application of the wireless power transfer system with a source at the center of a table powering several devices placed around the source.

[0041] Fig. 16(a) shows a 3D finite element model of a copper and magnetic material structure driven by a square loop of current around the choke point at its center. In this example, a structure may be composed of two boxes made of a conducting material such as copper, covered by a layer of magnetic material, and connected by a block of magnetic material. The inside of the two conducting boxes in this example would be shielded from AC electromagnetic fields generated outside the boxes and may house lossy objects that might lower the Q of the resonator or sensitive components that might be adversely affected by the AC electromagnetic fields. Also shown are the calculated magnetic field streamlines generated by this structure, indicating that the magnetic field lines tend to follow the lower reluctance path in the magnetic material. Fig. 16(b) shows interaction, as indicated by the calculated magnetic field streamlines, between two identical structures as shown in (a). Because of symmetry, and to reduce computational complexity, only one half of the system is modeled (but the computation assumes the symmetrical arrangement of the other half).

[0042] Fig. 17 shows an equivalent circuit representation of a magnetic resonator including a conducting wire wrapped N times around a structure, possibly containing magnetically permeable material. The inductance is realized using conducting loops wrapped around a structure comprising a magnetic material and the resistors represent loss mechanisms in the system (R_{wire} for resistive losses in the loop, R_{μ} denoting the equivalent series resistance of the structure surrounded by the loop). Losses may be minimized to realize high-Q resonators.

[0043] Fig. 18 shows a Finite Element Method (FEM) simulation of two high conductivity surfaces above and below a disk composed of lossy dielectric material, in an external magnetic field of frequency 6.78 MHz. Note that the magnetic field was uniform before the disk and conducting materials were introduced to the simulated environment. This simulation is performed in cylindrical coordinates. The image is azimuthally symmetric around the r= 0 axis. The lossy dielectric disk has $\epsilon_r = 1$ and σ= 10 S/m.

[0044] Fig. 19 shows a drawing of a magnetic resonator with a lossy object in its vicinity completely covered by a high-conductivity surface.

[0045] Fig. 20 shows a drawing of a magnetic resonator with a lossy object in its vicinity partially covered by a high-conductivity surface.

[0046] Fig. 21 shows a drawing of a magnetic resonator with a lossy object in its vicinity placed on top of a high-conductivity surface.

[0047] Fig. 22 shows a diagram of a completely wireless projector.

[0048] Fig. 23 shows the magnitude of the electric and magnetic fields along a line that contains the diameter of the circular loop inductor and along the axis of the loop inductor.

[0049] Fig. 24 shows a drawing of a magnetic resonator and its enclosure along with a necessary but lossy object placed either (a) in the corner of the enclosure, as far away from the resonator structure as possible or (b) in the center of the surface enclosed by the inductive element in the magnetic resonator.

[0050] Fig. 25 shows a drawing of a magnetic resonator with a high-conductivity surface above it and a lossy object, which may be brought into the vicinity of the resonator, but above the high-conductivity sheet.

[0051] Fig. 26(a) shows an axially symmetric FEM simulation of a thin conducting (copper) cylinder or disk (20 cm in diameter, 2 cm in height) exposed to an initially uniform, externally applied magnetic field (gray flux lines) along the z-axis. The axis of symmetry is at r=0. The magnetic streamlines shown originate at $z = -\infty$, where they are spaced from r=3 cm to r=10 cm in intervals of 1 cm. The axes scales are in meters. Fig. 26 (b) shows the same structure and externally applied field as in (a), except that the conducting cylinder has been modified to include a 0.25 mm layer of magnetic material (not visible) with $\mu_r^{'}$=40, on its outside surface. Note that the magnetic streamlines are deflected away from the cylinder significantly less than in (a).

[0052] Fig. 27 shows an axi-symmetric view of a variation based on the system shown in Fig. 26. Only one surface of the lossy material is covered by a layered structure of copper and magnetic materials. The inductor loop is placed on the side of the copper and magnetic material structure opposite to the lossy material as shown.

[0053] Fig. 28 (a) depicts a general topology of a matching circuit including an indirect coupling to a high-Q inductive element.

[0054] Fig. 28 (b) shows a block diagram of a magnetic resonator that includes a conductor loop inductor and a tunable impedance network. Physical electrical connections to this resonator may be made to the terminal connections.

[0055] Fig. 28 (c) depicts a general topology of a matching circuit directly coupled to a high-Q inductive element.

[0056] Fig. 28 (d) depicts a general topology of a symmetric matching circuit directly coupled to a high-Q inductive element and driven anti-symmetrically (balanced drive).

[0057] Fig. 28 (e) depicts a general topology of a matching circuit directly coupled to a high-Q inductive element and connected to ground at a point of symmetry of the main resonator (unbalanced drive).

[0058] Figs. 29(a) and 29(b) depict two topologies of matching circuits transformer-coupled (i.e. indirectly or inductively) to a high-Q inductive element. The highlighted portion of the Smith chart in (c) depicts the complex impedances (arising from L and R of the inductive element) that may be matched to an arbitrary real impedance Z_0 by the topology of Fig. 31(b) in the case $\omega L_2=1/\omega C_2$.

[0059] Figs. 30(a),(b),(c),(d),(e),(f) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in series with Z_0. The topologies shown in Figs. 30(a),(b),(c) are driven with a common-mode signal at the input terminals, while the topologies shown in Figs 30(d),(e),(f) are symmetric and receive a balanced drive. The highlighted portion of the Smith chart in 30(g) depicts the complex impedances that may be matched by these topologies. Figs. 30(h),(i),(j),(k),(l),(m) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in series with Z_0.

[0060] Figs. 31(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in series with Z_0. They are connected to ground at the center point of a capacitor and receive an unbalanced drive. The highlighted portion of the Smith chart in Fig. 31(d) depicts the complex impedances that may be matched by these topologies. Figs. 31(e),(f),(g) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in series with Z_0.

[0061] Figs. 32(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in series with Z_0. They are connected to ground by tapping at the center point of the inductor loop and receive an unbalanced drive. The highlighted portion of the Smith chart in (d) depicts the complex impedances that may be matched by these topologies, (e),(f),(g) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in series with Z_0.

[0062] Figs. 33(a),(b),(c),(d),(e),(f) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in parallel with Z_o. The topologies shown in Figs. 33(a),(b),(c) are driven with a common-mode signal at the input terminals, while the topologies shown in Figs 33(d),(e),(f) are symmetric and receive a balanced drive. The highlighted portion of the Smith chart in Fig. 33(g) depicts the complex impedances that may be matched by these topologies. Figs. 33(h),(i),(j),(k),(l),(m) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in parallel with Z_o.

[0063] Figs. 34(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in parallel with Z_0. They are connected to ground at the center point of a capacitor and receive an unbalanced drive. The highlighted portion of the Smith chart in (d) depicts the complex impedances that may be matched by these topologies. Figs. 34(e),(f),(g) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in parallel with Z_0.

[0064] Figs. 35(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in parallel with Z_0. They are connected to ground by tapping at the center point of the inductor loop and receive an unbalanced drive. The highlighted portion of the Smith chart in Figs. 35(d),(e), and (f) depict the complex impedances that may be matched by these topologies.

[0065] Figs. 36(a),(b),(c),(d) depict four topologies of networks of fixed and variable capacitors designed to produce an overall variable capacitance with finer tuning resolution and some with reduced voltage on the variable capacitor.

[0066] Figs. 37(a) and 37(b) depict two topologies of networks of fixed capacitors and a variable inductor designed to produce an overall variable capacitance.

[0067] Fig. 38 depicts a high level block diagram of a wireless power transmission system.

[0068] Fig. 39 depicts a block diagram of an exemplary wirelessly powered device.

[0069] Fig. 40 depicts a block diagram of the source of an exemplary wireless power transfer system.

[0070] Fig. 41 shows an equivalent circuit diagram of a magnetic resonator. The slash through the capacitor symbol indicates that the represented capacitor may be fixed or variable.

The port parameter measurement circuitry may be configured to measure certain electrical signals and may measure the magnitude and phase of signals.

[0071] Fig. 42 shows a circuit diagram of a magnetic resonator where the tunable impedance network is realized with voltage controlled capacitors. Such an implementation may be adjusted, tuned or controlled by electrical circuits including programmable or controllable voltage sources and/or computer processors. The voltage controlled capacitors may be adjusted in response to data measured by the port parameter measurement circuitry and processed by measurement analysis and control algorithms and hardware. The voltage controlled capacitors may be a switched bank of capacitors.

[0072] Fig. 43 shows an end-to-end wireless power transmission system. In this example, both the source and the device contain port measurement circuitry and a processor. The box labeled "coupler/switch" indicates that the port measurement circuitry may be connected to the resonator by a directional coupler or a switch, enabling the measurement, adjustment and control of the source and device resonators to take place in conjunction with, or separate from, the power transfer functionality.

[0073] Fig. 44 shows an end-to-end wireless power transmission system. In this example, only the source contains port measurement circuitry and a processor. In this case, the device resonator operating characteristics may be fixed or may be adjusted by analog control circuitry and without the need for control signals generated by a processor.

[0074] Fig. 45 shows an end-to-end wireless power transmission system. In this example, both the source and the device contain port measurement circuitry but only the source contains a processor. Data from the device is transmitted through a wireless communication channel, which could be implemented either with a separate antenna, or through some modulation of the source drive signal.

[0075] Fig. 46 shows an end-to-end wireless power transmission system. In this example, only the source contains port measurement circuitry and a processor. Data from the device is transmitted through a wireless communication channel, which could be implemented either with a separate antenna, or through some modulation of the source drive signal.

[0076] Fig. 47 shows coupled magnetic resonators whose frequency and impedance may be automatically adjusted using algorithms implemented using a processor or a computer.

[0077] Fig. 48 shows a varactor array.

[0078] Fig. 49 shows a device (laptop computer) being wirelessly powered or charged by a source, where both the source and device resonator are physically separated from, but electrically connected to, the source and device.

[0079] Fig. 50 (a) is an illustration of a wirelessly powered or charged laptop application where the device resonator is inside the laptop case and is not visible.

[0080] Fig. 50 (b) is an illustration of a wirelessly powered or charged laptop application where the resonator is underneath the laptop base and is electrically connected to the laptop power input by an electrical cable.

[0081] Fig. 50 (c) is an illustration of a wirelessly powered or charged laptop application where the resonator is attached to the laptop base.

[0082] Fig. 50 (d) is an illustration of a wirelessly powered or charged laptop application where the resonator is attached to the laptop display.

[0083] Fig. 51 is a diagram of rooftop PV panels with wireless power transfer.

DETAILED DESCRIPTION

[0084] As described above, this disclosure relates to coupled electromagnetic resonators with long-lived oscillatory resonant modes that may wirelessly transfer power from a power supply to a power drain. However, the technique is not restricted to electromagnetic resonators, but is general and may be applied to a wide variety of resonators and resonant objects. Therefore, we first describe the general technique, and then disclose electromagnetic examples for wireless energy transfer.

[0085] **Resonators**

[0086] A resonator may be defined as a system that can store energy in at least two different forms, and where the stored energy is oscillating between the two forms. The resonance has a specific oscillation mode with a resonant (modal) frequency, *f*, and a resonant (modal) field. The angular resonant frequency, *ω*, may be defined as $\omega = 2\pi f$, the resonant wavelength, λ, may be defined as $\lambda = c/f$, where *c* is the speed of light, and the resonant period, *T*, may be defined as $T = 1/f = 2\pi/\omega$. In the absence of loss mechanisms, coupling mechanisms or external energy supplying or draining mechanisms, the total resonator stored energy, *W*, would stay fixed

and the two forms of energy would oscillate, wherein one would be maximum when the other is minimum and vice versa.

[0087] In the absence of extraneous materials or objects, the energy in the resonator 102 shown in Fig. 1 may decay or be lost by intrinsic losses. The resonator fields then obey the following linear equation:

$$\frac{da(t)}{dt} = -i(\omega - i\Gamma)a(t),$$

where the variable *a(t)* is the resonant field amplitude, defined so that the energy contained within the resonator is given by $|a(t)|^2$. Γ is the intrinsic energy decay or loss rate (e.g. due to absorption and radiation losses).

[0088] The Quality Factor, or *Q*-factor, or *Q*, of the resonator, which characterizes the energy decay, is inversely proportional to these energy losses. It may be defined as $Q=\omega*W/P$, where *P* is the time-averaged power lost at steady state. That is, a resonator 102 with a high-*Q* has relatively low intrinsic losses and can store energy for a relatively long time. Since the resonator loses energy at its intrinsic decay rate, 2Γ, its *Q*, also referred to as its intrinsic *Q*, is given by $Q = \omega/2\Gamma$. The quality factor also represents the number of oscillation periods, *T*, it takes for the energy in the resonator to decay by a factor of *e*.

[0089] As described above, we define the quality factor or *Q* of the resonator as that due only to intrinsic loss mechanisms. A subscript index such as Q_1, indicates the resonator (resonator 1 in this case) to which the *Q* refers. Fig. 2 shows an electromagnetic resonator 102 labeled according to this convention. Note that in this figure, there are no extraneous objects or additional resonators in the vicinity of resonator 1.

[0090] Extraneous objects and/or additional resonators in the vicinity of a first resonator may perturb or load the first resonator, thereby perturbing or loading the *Q* of the first resonator, depending on a variety of factors such as the distance between the resonator and object or other resonator, the material composition of the object or other resonator, the structure of the first resonator, the power in the first resonator, and the like. Unintended external energy losses or coupling mechanisms to extraneous materials and objects in the vicinity of the resonators may be referred to as "perturbing" the *Q* of a resonator, and may be indicated by a subscript within rounded parentheses, (). Intended external energy losses, associated with energy transfer via coupling to other resonators and to generators and loads in the wireless energy transfer system

may be referred to as "loading" the Q of the resonator, and may be indicated by a subscript within square brackets, [].

[0091] The Q of a resonator 102 connected or coupled to a power generator, g, or load 302 , l, may be called the "loaded quality factor" or the "loaded Q" and may be denoted by $Q_{[g]}$ or $Q_{[l]}$, as illustrated in Fig. 3. In general, there may be more than one generator or load 302 connected to a resonator 102. However, we do not list those generators or loads separately but rather use "g" and "l" to refer to the equivalent circuit loading imposed by the combinations of generators and loads. In general descriptions, we may use the subscript "l" to refer to either generators or loads connected to the resonators.

[0092] In some of the discussion herein, we define the "loading quality factor" or the "loading Q" due to a power generator or load connected to the resonator, as $\delta Q_{[l]}$, where, $1/\delta Q_{[l]} \equiv 1/Q_{[l]} - 1/Q$. Note that the larger the loading Q, $\delta Q_{[l]}$, of a generator or load, the less the loaded Q, $Q_{[l]}$, deviates from the unloaded Q of the resonator.

[0093] The Q of a resonator in the presence of an extraneous object 402, p, that is not intended to be part of the energy transfer system may be called the "perturbed quality factor" or the "perturbed Q" and may be denoted by $Q_{(p)}$, as illustrated in Fig. 4. In general, there may be many extraneous objects, denoted as $p1$, $p2$, etc., or a set of extraneous objects $\{p\}$, that perturb the Q of the resonator 102. In this case, the perturbed Q may be denoted $Q_{(p1+p2+...)}$ or $Q_{(\{p\})}$. For example, $Q_{1(brick+wood)}$ may denote the perturbed quality factor of a first resonator in a system for wireless power exchange in the presence of a brick and a piece of wood, and $Q_{2(\{office\})}$ may denote the perturbed quality factor of a second resonator in a system for wireless power exchange in an office environment.

[0094] In some of the discussion herein, we define the "perturbing quality factor" or the "perturbing Q" due to an extraneous object, p, as $\delta Q_{(p)}$, where $1/\delta Q_{(p)} \equiv 1/Q_{(p)} - 1/Q$. As stated before, the perturbing quality factor may be due to multiple extraneous objects, $p1$, $p2$, etc. or a set of extraneous objects, $\{p\}$. The larger the perturbing Q, $\delta Q_{(p)}$, of an object, the less the perturbed Q, $Q_{(p)}$, deviates from the unperturbed Q of the resonator.

[0095] In some of the discussion herein, we also define $\Theta_{(p)} \equiv Q_{(p)} / Q$ and call it the "quality factor insensitivity" or the "Q-insensitivity" of the resonator in the presence of an

extraneous object. A subscript index, such as $\Theta_{1(p)}$, indicates the resonator to which the perturbed and unperturbed quality factors are referring, namely, $\Theta_{1(p)} \equiv Q_{1(p)} / Q_1$.

[0096] Note that the quality factor, Q, may also be characterized as "unperturbed", when necessary to distinguish it from the perturbed quality factor, $Q_{(p)}$, and "unloaded", when necessary to distinguish it from the loaded quality factor, $Q_{[l]}$. Similarly, the perturbed quality factor, $Q_{(p)}$, may also be characterized as "unloaded", when necessary to distinguish them from the loaded perturbed quality factor, $Q_{(p)[l]}$.

[0097] Coupled Resonators

[0098] Resonators having substantially the same resonant frequency, coupled through any portion of their near-fields may interact and exchange energy. There are a variety of physical pictures and models that may be employed to understand, design, optimize and characterize this energy exchange. One way to describe and model the energy exchange between two coupled resonators is using coupled mode theory (CMT).

[0099] In coupled mode theory, the resonator fields obey the following set of linear equations:

$$\frac{da_m(t)}{dt} = -i(\omega_m - i\Gamma_m)a_m(t) + i\sum_{n \neq m} \kappa_{mn} a_n(t)$$

where the indices denote different resonators and and κ_{mn} are the coupling coefficients between the resonators. For a reciprocal system, the coupling coefficients may obey the relation $\kappa_{mn} = \kappa_{nm}$. Note that, for the purposes of the present specification, far-field radiation interference effects will be ignored and thus the coupling coefficients will be considered real. Furthermore, since in all subsequent calculations of system performance in this specification the coupling coefficients appear only with their square, κ_{mn}^2, we use κ_{mn} to denote the absolute value of the real coupling coefficients.

[00100] Note that the coupling coefficient, κ_{mn}, from the CMT described above is related to the so-called coupling factor, k_{mn}, between resonators m and n by $k_{mn} = 2\kappa_{mn} / \sqrt{\omega_m \omega_n}$.

We define a "strong-coupling factor", U_{mn}, as the ratio of the coupling and loss rates between resonators *m* and *n*, by $U_{mn} = \kappa_{mn} / \sqrt{\Gamma_m \Gamma_n} = k_{mn}\sqrt{Q_m Q_n}$.

[00101] The quality factor of a resonator *m,* in the presence of a similar frequency resonator *n* or additional resonators, may be loaded by that resonator *n* or additional resonators, in a fashion similar to the resonator being loaded by a connected power generating or consuming device. The fact that resonator *m* may be loaded by resonator *n* and vice versa is simply a different way to see that the resonators are coupled.

[00102] The loaded *Q'*s of the resonators in these cases may be denoted as $Q_{m[n]}$ and $Q_{n[m]}$. For multiple resonators or loading supplies or devices, the total loading of a resonator may be determined by modeling each load as a resistive loss, and adding the multiple loads in the appropriate parallel and/or series combination to determine the equivalent load of the ensemble.

[00103] In some of the discussion herein, we define the "loading quality factor" or the "loading Q_m" of resonator *m* due to resonator *n* as $\delta Q_{m[n]}$, where $1/\delta Q_{m[n]} \equiv 1/Q_{m[n]} - 1/Q_m$. Note that resonator *n* is also loaded by resonator *m* and its "loading Q_n" is given by $1/\delta Q_{n[m]} \equiv 1/Q_{n[m]} - 1/Q_n$.

[00104] When one or more of the resonators are connected to power generators or loads, the set of linear equations is modified to:

$$\frac{da_m(t)}{dt} = -i(\omega_m - i\Gamma_m)a_m(t) + i\sum_{n\neq m}\kappa_{mn}a_n(t) - \kappa_m a_m(t) + \sqrt{2\kappa_m}s_{+m}(t)$$
$$s_{-m}(t) = \sqrt{2\kappa_m}a_m(t) - s_{+m}(t),$$

where $s_{+m}(t)$ and $s_{-m}(t)$ are respectively the amplitudes of the fields coming from a generator into the resonator *m* and going out of the resonator *m* either back towards the generator or into a load, defined so that the power they carry is given by $|s_{+m}(t)|^2$ and $|s_{-m}(t)|^2$. The loading coefficients κ_m relate to the rate at which energy is exchanged between the resonator *m* and the generator or load connected to it.

[00105] Note that the loading coefficient, κ_m, from the CMT described above is related to the loading quality factor, $\delta Q_{m[l]}$, defined earlier, by $\delta Q_{m[l]} = \omega_m / 2\kappa_m$.

[00106] We define a “strong-loading factor”, $U_{m[l]}$, as the ratio of the loading and loss rates of resonator m, $U_{m[l]} = \kappa_m / \Gamma_m = Q_m / \delta Q_{m[l]}$.

[00107] Fig. 1(a) shows an example of two coupled resonators 1000, a first resonator 102S, configured as a source resonator and a second resonator 102D, configured as a device resonator. Energy may be transferred over a distance *D* between the resonators. The source resonator 102S may be driven by a power supply or generator (not shown). Work may be extracted from the device resonator 102D by a power consuming drain or load (e.g. a load resistor, not shown). Let us use the subscripts “s” for the source, “d” for the device, “g” for the generator, and “l” for the load, and, since in this example there are only two resonators and $\kappa_{sd}=\kappa_{ds}$, let us drop the indices on κ_{sd}, k_{sd}, and U_{sd}, and denote them as κ, k, and U, respectively.

[00108] The power generator may be constantly driving the source resonator at a constant driving frequency, *f*, corresponding to an angular driving frequency, ω, where $\omega = 2\pi f$.

[00109] In this case, the efficiency, $\eta = |s_{-d}|^2 / |s_{+s}|^2$, of the power transmission from the generator to the load (via the source and device resonators) is maximized under the following conditions: The source resonant frequency, the device resonant frequency and the generator driving frequency have to be matched, namely

$$\omega_s = \omega_d = \omega .$$

Furthermore, the loading Q of the source resonator due to the generator, $\delta Q_{s[g]}$, has to be matched (equal) to the loaded Q of the source resonator due to the device resonator and the load, $Q_{s[dl]}$, and inversely the loading Q of the device resonator due to the load, $\delta Q_{d[l]}$, has to be matched (equal) to the loaded Q of the device resonator due to the source resonator and the generator, $Q_{d[sg]}$, namely

$$\delta Q_{s[g]} = Q_{s[dl]} \text{ and } \delta Q_{d[l]} = Q_{d[sg]} .$$

These equations determine the optimal loading rates of the source resonator by the generator and of the device resonator by the load as

$$U_{d[l]} = \kappa_d / \Gamma_d = Q_d / \delta Q_{d[l]} = \sqrt{1+U^2} = \sqrt{1+\left(\kappa / \sqrt{\Gamma_s \Gamma_d}\right)^2} = Q_s / \delta Q_{s[g]} = \kappa_s / \Gamma_s = U_{s[g]} .$$

Note that the above frequency matching and Q matching conditions are together known as "impedance matching" in electrical engineering.

[00110] Under the above conditions, the maximized efficiency is a monotonically increasing function of only the strong-coupling factor, $U=\kappa/\sqrt{\Gamma_s\Gamma_d}=k\sqrt{Q_sQ_d}$, between the source and device resonators and is given by, $\eta=U^2/\left(1+\sqrt{1+U^2}\right)^2$, as shown in Fig. 5. Note that the coupling efficiency, η, is greater than 1% when U is greater than 0.2, is greater than 10% when U is greater than 0.7, is greater than 17% when U is greater than 1, is greater than 52% when U is greater than 3, is greater than 80% when U is greater than 9, is greater than 90% when U is greater than 19, and is greater than 95% when U is greater than 45. In some applications, the regime of operation where $U>1$ may be referred to as the "strong-coupling" regime.

[00111] Since a large $U=\kappa/\sqrt{\Gamma_s\Gamma_d}=\left(2\kappa/\sqrt{\omega_s\omega_d}\right)\sqrt{Q_sQ_d}$ is desired in certain circumstances, resonators may be used that are high-Q. The Q of each resonator may be high. The geometric mean of the resonator Q's, $\sqrt{Q_sQ_d}$ may also or instead be high.

[00112] The coupling factor, k, is a number between $0\leq k\leq 1$, and it may be independent (or nearly independent) of the resonant frequencies of the source and device resonators, rather it may determined mostly by their relative geometry and the physical decay-law of the field mediating their coupling. In contrast, the coupling coefficient, $\kappa=k\sqrt{\omega_s\omega_d}/2$, may be a strong function of the resonant frequencies. The resonant frequencies of the resonators may be chosen preferably to achieve a high Q rather than to achieve a low Γ, as these two goals may be achievable at two separate resonant frequency regimes.

[00113] A high-Q resonator may be defined as one with $Q>100$. Two coupled resonators may be referred to as a system of high-Q resonators when each resonator has a Q greater than 100, $Q_s>100$ and $Q_d>100$. In other implementationss, two coupled resonators may be referred to as a system of high-Q resonators when the geometric mean of the resonator Q's is greater than 100, $\sqrt{Q_sQ_d}>100$.

[00114] The resonators may be named or numbered. They may be referred to as source resonators, device resonators, first resonators, second resonators, repeater resonators, and the like. It is to be understood that while two resonators are shown in Fig. 1, and in many of the examples below, other implementations may include three (3) or more resonators. For example, a single source resonator 102S may transfer energy to multiple device resonators 102D or multiple devices. Energy may be transferred from a first device to a second, and then from the second device to the third, and so forth. Multiple sources may transfer energy to a single device or to multiple devices connected to a single device resonator or to multiple devices connected to multiple device resonators. Resonators 102 may serve alternately or simultaneously as sources, devices, or they may be used to relay power from a source in one location to a device in another location. Intermediate electromagnetic resonators 102 may be used to extend the distance range of wireless energy transfer systems. Multiple resonators 102 may be daisy chained together, exchanging energy over extended distances and with a wide range of sources and devices. High power levels may be split between multiple sources 102S, transferred to multiple devices and recombined at a distant location.

[00115] The analysis of a single source and a single device resonator may be extended to multiple source resonators and/or multiple device resonators and/or multiple intermediate resonators. In such an analysis, the conclusion may be that large strong-coupling factors, U_{mn}, between at least some or all of the multiple resonators is preferred for a high system efficiency in the wireless energy transfer. Again, implementations may use source, device and intermediate resonators that have a high Q. The Q of each resonator may be high. The geometric mean $\sqrt{Q_m Q_n}$ of the Q's for pairs of resonators m and n, for which a large U_{mn} is desired, may also or instead be high.

[00116] Note that since the strong-coupling factor of two resonators may be determined by the relative magnitudes of the loss mechanisms of each resonator and the coupling mechanism between the two resonators, the strength of any or all of these mechanisms may be perturbed in the presence of extraneous objects in the vicinity of the resonators as described above.

[00117] Continuing the conventions for labeling from the previous sections, we describe k as the coupling factor in the absence of extraneous objects or materials. We denote the coupling factor in the presence of an extraneous object, p, as $k_{(p)}$, and call it the "perturbed coupling factor" or the "perturbed k". Note that the coupling factor, k, may also be characterized as "unperturbed", when necessary to distinguish from the perturbed coupling factor $k_{(p)}$.

[00118] We define $\delta k_{(p)} \equiv k_{(p)} - k$ and we call it the "perturbation on the coupling factor" or the "perturbation on k" due to an extraneous object, p.

[00119] We also define $\beta_{(p)} \equiv k_{(p)}/k$ and we call it the "coupling factor insensitivity" or the "k-insensitivity". Lower indices, such as $\beta_{12(p)}$, indicate the resonators to which the perturbed and unperturbed coupling factor is referred to, namely $\beta_{12(p)} \equiv k_{12(p)}/k_{12}$.

[00120] Similarly, we describe U as the strong-coupling factor in the absence of extraneous objects. We denote the strong-coupling factor in the presence of an extraneous object, p, as $U_{(p)}$, $U_{(p)} = k_{(p)}\sqrt{Q_{1(p)}Q_{2(p)}}$, and call it the "perturbed strong-coupling factor" or the "perturbed U". Note that the strong-coupling factor U may also be characterized as "unperturbed", when necessary to distinguish from the perturbed strong-coupling factor $U_{(p)}$. Note that the strong-coupling factor U may also be characterized as "unperturbed", when necessary to distinguish from the perturbed strong-coupling factor $U_{(p)}$.

[00121] We define $\delta U_{(p)} \equiv U_{(p)} - U$ and call it the "perturbation on the strong-coupling factor" or the "perturbation on U" due to an extraneous object, p.

[00122] We also define $\Xi_{(p)} \equiv U_{(p)}/U$ and call it the "strong-coupling factor insensitivity" or the "U-insensitivity". Lower indices, such as $\Xi_{12(p)}$, indicate the resonators to which the perturbed and unperturbed coupling factor refers, namely $\Xi_{12(p)} \equiv U_{12(p)}/U_{12}$.

[00123] The efficiency of the energy exchange in a perturbed system may be given by the same formula giving the efficiency of the unperturbed system, where all parameters such as strong-coupling factors, coupling factors, and quality factors are replaced by their perturbed equivalents. For example, in a system of wireless energy transfer including one source and one device resonator, the optimal efficiency may calculated as $\eta_{(p)} = \left[U_{(p)} \Big/ \left(1 + \sqrt{1 + U_{(p)}^{\;2}} \right) \right]^2$.

Therefore, in a system of wireless energy exchange which is perturbed by extraneous objects, large perturbed strong-coupling factors, $U_{mn(p)}$, between at least some or all of the multiple resonators may be desired for a high system efficiency in the wireless energy transfer. Source, device and/or intermediate resonators may have a high $Q_{(p)}$.

[00124] Some extraneous perturbations may sometimes be detrimental for the perturbed strong-coupling factors (via large perturbations on the coupling factors or the quality factors). Therefore, techniques may be used to reduce the effect of extraneous perturbations on the system and preserve large strong-coupling factor insensitivites.

[00125] **Efficiency of Energy Exchange**

[00126] The so-called "useful" energy in a useful energy exchange is the energy or power that must be delivered to a device (or devices) in order to power or charge the device. The transfer efficiency that corresponds to a useful energy exchange may be system or application dependent. For example, high power vehicle charging applications that transfer kilowatts of power may need to be at least 80% efficient in order to supply useful amounts of power resulting in a useful energy exchange sufficient to recharge a vehicle battery, without significantly heating up various components of the transfer system. In some consumer electronics applications, a useful energy exchange may include any energy transfer efficiencies greater than 10%, or any other amount acceptable to keep rechargeable batteries "topped off" and running for long periods of time. For some wireless sensor applications, transfer efficiencies that are much less than 1% may be adequate for powering multiple low power sensors from a single source located a significant distance from the sensors. For still other applications, where wired power transfer is either impossible or impractical, a wide range of transfer efficiencies may be acceptable for a useful energy exchange and may be said to supply useful power to devices in those applications. In general, an operating distance is any distance over which a useful energy exchange is or can be maintained according to the principles disclosed herein.

[00127] A useful energy exchange for a wireless energy transfer in a powering or recharging application may be efficient, highly efficient, or efficient enough, as long as the wasted energy levels, heat dissipation, and associated field strengths are within tolerable limits. The tolerable limits may depend on the application, the environment and the system location. Wireless energy transfer for powering or recharging applications may be efficient, highly efficient, or efficient enough, as long as the desired system performance may be attained for the reasonable cost restrictions, weight restrictions, size restrictions, and the like. Efficient energy transfer may be determined relative to that which could be achieved using traditional inductive techniques that are not high-Q systems. Then, the energy transfer may be defined as being efficient, highly efficient, or efficient enough, if more energy is delivered than could be delivered by similarly sized coil structures in traditional inductive schemes over similar distances or alignment offsets.

[00128] Note that, even though certain frequency and Q matching conditions may optimize the system efficiency of energy transfer, these conditions may not need to be exactly met in order to have efficient enough energy transfer for a useful energy exchange. Efficient energy exchange may be realized so long as the relative offset of the resonant frequencies ($|\omega_m - \omega_n| / \sqrt{\omega_m \omega_n}$) is less than approximately the maximum among $1/Q_{m(p)}$, $1/Q_{n(p)}$ and $k_{mn(p)}$. The Q matching condition may be less critical than the frequency matching condition for efficient energy exchange. The degree by which the strong-loading factors, $U_{m[l]}$, of the resonators due to generators and/or loads may be away from their optimal values and still have efficient enough energy exchange depends on the particular system, whether all or some of the generators and/or loads are Q-mismatched and so on.

[00129] Therefore, the resonant frequencies of the resonators may not be exactly matched, but may be matched within the above tolerances. The strong-loading factors of at least some of the resonators due to generators and/or loads may not be exactly matched to their optimal value. The voltage levels, current levels, impedance values, material parameters, and the like may not be at the exact values described in the disclosure but will be within some acceptable tolerance of those values. The system optimization may include cost, size, weight, complexity, and the like, considerations, in addition to efficiency, Q, frequency, strong coupling factor, and the like, considerations. Some system performance parameters, specifications, and designs may be far from optimal in order to optimize other system performance parameters, specifications and designs.

[00130] In some applications, at least some of the system parameters may be varying in time, for example because components, such as sources or devices, may be mobile or aging or because the loads may be variable or because the perturbations or the environmental conditions are changing etc. In these cases, in order to achieve acceptable matching conditions, at least some of the system parameters may need to be dynamically adjustable or tunable. All the system parameters may be dynamically adjustable or tunable to achieve approximately the optimal operating conditions. However, based on the discussion above, efficient enough energy exchange may be realized even if some system parameters are not variable. In some examples, at least some of the devices may not be dynamically adjusted. In some examples, at least some of the sources may not be dynamically adjusted. In some examples, at least some of the intermediate resonators may not be dynamically adjusted. In some examples, none of the system parameters may be dynamically adjusted.

[00131] __**Electromagnetic Resonators**__

[00132] The resonators used to exchange energy may be electromagnetic resonators. In such resonators, the intrinsic energy decay rates, Γ_m, are given by the absorption (or resistive) losses and the radiation losses of the resonator.

[00133] The resonator may be constructed such that the energy stored by the electric field is primarily confined within the structure and that the energy stored by the magnetic field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant magnetic near-field. These types of resonators may be referred to as magnetic resonators.

[00134] The resonator may be constructed such that the energy stored by the magnetic field is primarily confined within the structure and that the energy stored by the electric field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant electric near-field. These types of resonators may be referred to as electric resonators.

[00135] Note that the total electric and magnetic energies stored by the resonator have to be equal, but their localizations may be quite different. In some cases, the ratio of the average electric field energy to the average magnetic field energy specified at a distance from a resonator may be used to characterize or describe the resonator.

[00136] Electromagnetic resonators may include an inductive element, a distributed inductance, or a combination of inductances with inductance, L, and a capacitive element, a distributed capacitance, or a combination of capacitances, with capacitance, C. A minimal circuit model of an electromagnetic resonator 102 is shown in Fig. 6a. The resonator may include an inductive element 108 and a capacitive element 104. Provided with initial energy, such as electric field energy stored in the capacitor 104, the system will oscillate as the capacitor discharges transferring energy into magnetic field energy stored in the inductor 108 which in turn transfers energy back into electric field energy stored in the capacitor 104.

[00137] The resonators 102 shown in Figs. 6(b)(c)(d) may be referred to as magnetic resonators. Magnetic resonators may be preferred for wireless energy transfer applications in populated environments because most everyday materials including animals, plants, and humans are non-magnetic (i.e., $\mu_r \approx 1$), so their interaction with magnetic fields is minimal and due primarily to eddy currents induced by the time-variation of the magnetic fields, which is a second-order effect. This characteristic is important both for safety reasons and because it reduces the potential for interactions with extraneous environmental objects and materials that could alter system performance.

[00138] Fig. 6d shows a simplified drawing of some of the electric and magnetic field lines associated with an exemplary magnetic resonator 102B. The magnetic resonator 102B may include a loop of conductor acting as an inductive element 108 and a capacitive element 104 at the ends of the conductor loop. Note that this drawing depicts most of the energy in the region surrounding the resonator being stored in the magnetic field, and most of the energy in the resonator (between the capacitor plates) stored in the electric field. Some electric field, owing to fringing fields, free charges, and the time varying magnetic field, may be stored in the region around the resonator, but the magnetic resonator may be designed to confine the electric fields to be close to or within the resonator itself, as much as possible.

[00139] The inductor 108 and capacitor 104 of an electromagnetic resonator 102 may be bulk circuit elements, or the inductance and capacitance may be distributed and may result from the way the conductors are formed, shaped, or positioned, in the structure. For example, the inductor 108 may be realized by shaping a conductor to enclose a surface area, as shown in Figs. 6(b)(c)(d). This type of resonator 102 may be referred to as a capacitively-loaded loop inductor. Note that we may use the terms “loop” or “coil” to indicate generally a conducting structure (wire, tube, strip, etc.), enclosing a surface of any shape and dimension, with any number of turns. In Fig. 6b, the enclosed surface area is circular, but the surface may be any of a wide variety of other shapes and sizes and may be designed to achieve certain system performance specifications. As an example to indicate how inductance scales with physical dimensions, the inductance for a length of circular conductor arranged to form a circular single-turn loop is approximately,

$$L = \mu_0 x(\ln\frac{8x}{a} - 2),$$

where μ_0 is the magnetic permeability of free space, x, is the radius of the enclosed circular surface area and, a, is the radius of the conductor used to form the inductor loop. A more precise value of the inductance of the loop may be calculated analytically or numerically.

[00140] The inductance for other cross-section conductors, arranged to form other enclosed surface shapes, areas, sizes, and the like, and of any number of wire turns, may be calculated analytically, numerically or it may be determined by measurement. The inductance may be realized using inductor elements, distributed inductance, networks, arrays, series and parallel combinations of inductors and inductances, and the like. The inductance may be fixed or variable and may be used to vary impedance matching as well as resonant frequency operating conditions.

[00141] There are a variety of ways to realize the capacitance required to achieve the desired resonant frequency for a resonator structure. Capacitor plates 110 may be formed and utilized as shown in Fig. 6b, or the capacitance may be distributed and be realized between adjacent windings of a multi-loop conductor 114, as shown in Fig. 6c. The capacitance may be realized using capacitor elements, distributed capacitance, networks, arrays, series and parallel combinations of capacitances, and the like. The capacitance may be fixed or variable and may be used to vary impedance matching as well as resonant frequency operating conditions.

[00142] It is to be understood that the inductance and capacitance in an electromagnetic resonator 102 may be lumped, distributed, or a combination of lumped and distributed inductance and capacitance and that electromagnetic resonators may be realized by combinations of the various elements, techniques and effects described herein.

[00143] Electromagnetic resonators 102 may be include inductors, inductances, capacitors, capacitances, as well as additional circuit elements such as resistors, diodes, switches, amplifiers, diodes, transistors, transformers, conductors, connectors and the like.

[00144] **<u>Resonant Frequency of an Electromagnetic Resonator</u>**

[00145] An electromagnetic resonator 102 may have a characteristic, natural, or resonant frequency determined by its physical properties. This resonant frequency is the frequency at which the energy stored by the resonator oscillates between that stored by the electric field, W_E, ($W_E=q^2/2C$, where q is the charge on the capacitor, C) and that stored by the magnetic field, W_B, ($W_B = Li^2/2$, where i is the current through the inductor, L) of the resonator. In the absence of any losses in the system, energy would continually be exchanged between the electric field in the capacitor 104 and the magnetic field in the inductor 108. The frequency at which this energy is exchanged may be called the characteristic frequency, the natural frequency, or the resonant frequency of the resonator, and is given by ω,

$$\omega = 2\pi f = \sqrt{\frac{1}{LC}}.$$

[00146] The resonant frequency of the resonator may be changed by tuning the inductance, L, and/or the capacitance, C, of the resonator. The resonator frequency may be design to operate at the so-called ISM (Industrial, Scientific and Medical) frequencies as specified by the FCC. The resonator frequency may be chosen to meet certain field limit specifications, specific absorption rate (SAR) limit specifications, electromagnetic compatibility (EMC) specifications, electromagnetic interference (EMI) specifications, component size, cost or performance specifications, and the like.

[00147] **<u>Quality Factor of an Electromagnetic Resonator</u>**

[00148] The energy in the resonators 102 shown in Fig. 6 may decay or be lost by intrinsic losses including absorptive losses (also called ohmic or resistive losses) and/or radiative losses. The Quality Factor, or *Q*, of the resonator, which characterizes the energy decay, is inversely proportional to these losses. Absorptive losses may be caused by the finite conductivity of the conductor used to form the inductor as well as by losses in other elements, components, connectors, and the like, in the resonator. An inductor formed from low loss materials may be referred to as a "high-*Q* inductive element" and elements, components, connectors and the like with low losses may be referred to as having "high resistive *Q*'s". In general, the total absorptive loss for a resonator may be calculated as the appropriate series and/or parallel combination of resistive losses for the various elements and components that make up the resonator. That is, in the absence of any significant radiative or component/connection losses, the *Q* of the resonator may be given by, Q_{abs},

$$Q_{abs} = \frac{\omega L}{R_{abs}},$$

where ω, is the resonant frequency, L, is the total inductance of the resonator and the resistance for the conductor used to form the inductor, for example, may be given by $R_{abs} = l\rho/A$, (l is the length of the wire, ρ is the resistivity of the conductor material, and A is the cross-sectional area over which current flows in the wire). For alternating currents, the cross-sectional area over which current flows may be less than the physical cross-sectional area of the conductor owing to the skin effect. Therefore, high-Q magnetic resonators may be composed of conductors with high conductivity, relatively large surface areas and/or with specially designed profiles (e.g. Litz wire) to minimize proximity effects and reduce the AC resistance.

[00149] The magnetic resonator structures may include high-Q inductive elements composed of high conductivity wire, coated wire, Litz wire, ribbon, strapping or plates, tubing, paint, gels, traces, and the like. The magnetic resonators may be self-resonant, or they may include external coupled elements such as capacitors, inductors, switches, diodes, transistors, transformers, and the like. The magnetic resonators may include distributed and lumped capacitance and inductance. In general, the Q of the resonators will be determined by the Q's of all the individual components of the resonator.

[00150] Because Q is proportional to inductance, L, resonators may be designed to increase L, within certain other constraints. One way to increase L, for example, is to use more than one turn of the conductor to form the inductor in the resonator. Design techniques and trade-offs may depend on the application, and a wide variety of structures, conductors, components, and resonant frequencies may be chosen in the design of high-Q magnetic resonators.

[00151] In the absence of significant absorption losses, the Q of the resonator may be determined primarily by the radiation losses, and given by, $Q_{rad} = \omega L/R_{rad}$, where R_{rad} is the radiative loss of the resonator and may depend on the size of the resonator relative to the frequency, ω, or wavelength, λ, of operation. For the magnetic resonators discussed above, radiative losses may scale as $R_{rad} \sim (x/\lambda)^4$ (characteristic of magnetic dipole radiation), where x is a characteristic dimension of the resonator, such as the radius of the inductive element shown in Fig. 6b, and where $\lambda = c/f$, where c is the speed of light and f is as defined above. The size of the magnetic resonator may be much less than the wavelength of operation so radiation losses may be very small. Such structures may be referred to as sub-wavelength resonators. Radiation may be a loss mechanism for non-radiative wireless energy transfer systems and designs may be chosen to reduce or minimize R_{rad}. Note that a high-Q_{rad} may be desirable for non-radiative wireless energy transfer schemes.

[00152] Note too that the design of resonators for non-radiative wireless energy transfer differs from antennas designed for communication or far-field energy transmission purposes. Specifically, capacitively-loaded conductive loops may be used as resonant antennas (for example in cell phones), but those operate in the far-field regime where the radiation Q's are intentionally designed to be small to make the antenna efficient at radiating energy. Such designs are not appropriate for the efficient near-field wireless energy transfer technique disclosed in this application.

[00153] The quality factor of a resonator including both radiative and absorption losses is $Q = \omega L / (R_{abs} + R_{rad})$. Note that there may be a maximum Q value for a particular resonator and that resonators may be designed with special consideration given to the size of the resonator, the materials and elements used to construct the resonator, the operating frequency, the connection mechanisms, and the like, in order to achieve a high-Q resonator. Fig. 7 shows a plot of Q of an exemplary magnetic resonator (in this case a coil with a diameter of 60 cm made of copper pipe with an outside diameter (OD) of 4 cm) that may be used for wireless power transmission at MHz frequencies. The absorptive Q (dashed line) 702 increases with frequency, while the radiative Q (dotted line) 704 decreases with frequency, thus leading the overall Q to peak 708 at a particular frequency. Note that the Q of this exemplary resonator is greater than 100 over a wide frequency range. Magnetic resonators may be designed to have high-Q over a range of frequencies and system operating frequency may set to any frequency in that range.

[00154] When the resonator is being described in terms of loss rates, the Q may be defined using the intrinsic decay rate, 2Γ, as described previously. The intrinsic decay rate is the rate at which an uncoupled and undriven resonator loses energy. For the magnetic resonators described above, the intrinsic loss rate may be given by $\Gamma = (R_{abs} + R_{rad}) / 2L$, and the quality factor, Q, of the resonator is given by $Q = \omega / 2\Gamma$.

[00155] Note that a quality factor related only to a specific loss mechanism may be denoted as $Q_{mechanism}$, if the resonator is not specified, or as $Q_{1,mechanism}$, if the resonator is specified (e.g. resonator 1). For example, $Q_{1,rad}$ is the quality factor for resonator 1 related to its radiation losses.

[00156] **Electromagnetic Resonator Near-Fields**

[00157] The high-Q electromagnetic resonators used in the near-field wireless energy transfer system disclosed here may be sub-wavelength objects. That is, the physical dimensions of the resonator may be much smaller than the wavelength corresponding to the resonant frequency. Sub-wavelength magnetic resonators may have most of the energy in the region surrounding the resonator stored in their magnetic near-fields, and these fields may also be described as stationary or non-propagating because they do not radiate away from the resonator. The extent of the near-field in the area surrounding the resonator is typically set by the wavelength, so it may extend well beyond the resonator itself for a sub-wavelength resonator. The limiting surface, where the field behavior changes from near-field behavior to far-field behavior may be called the "radiation caustic".

[00158] The strength of the near-field is reduced the farther one gets away from the resonator. While the field strength of the resonator near-fields decays away from the resonator, the fields may still interact with objects brought into the general vicinity of the resonator. The degree to which the fields interact depends on a variety of factors, some of which may be controlled and designed, and some of which may not. The wireless energy transfer schemes described herein may be realized when the distance between coupled resonators is such that one resonator lies within the radiation caustic of the other.

[00159] The near-field profiles of the electromagnetic resonators may be similar to those commonly associated with dipole resonators or oscillators. Such field profiles may be described as omni-directional, meaning the magnitudes of the fields are non-zero in all directions away from the object.

[00160] <u>Characteristic Size of An Electromagnetic Resonator</u>

[00161] Spatially separated and/or offset magnetic resonators of sufficient Q may achieve efficient wireless energy transfer over distances that are much larger than have been seen in the prior art, even if the sizes and shapes of the resonator structures are different. Such resonators may also be operated to achieve more efficient energy transfer than was achievable with previous techniques over shorter range distances. We describe such resonators as being capable of mid-range energy transfer.

[00162] Mid-range distances may be defined as distances that are larger than the characteristic dimension of the smallest of the resonators involved in the transfer, where the distance is measured from the center of one resonator structure to the center of a spatially separated second resonator structure. In this definition, two-dimensional resonators are spatially separated when the areas circumscribed by their inductive elements do not intersect and three-dimensional resonators are spatially separated when their volumes do not intersect. A two-dimensional resonator is spatially separated from a three-dimensional resonator when the area circumscribed by the former is outside the volume of the latter.

[00163] Fig. 8 shows some example resonators with their characteristic dimensions labeled. It is to be understood that the characteristic sizes 802 of resonators 102 may be defined in terms of the size of the conductor and the area circumscribed or enclosed by the inductive element in a magnetic resonator and the length of the conductor forming the capacitive element of an electric resonator. Then, the characteristic size 802 of a resonator 102, x_{char}, may be equal to the radius of the smallest sphere that can fit around the inductive or capacitive element of the magnetic or electric resonator respectively, and the center of the resonator structure is the center of the sphere. The characteristic thickness 804, t_{char}, of a resonator 102 may be the smallest possible height of the highest point of the inductive or capacitive element in the magnetic or capacitive resonator respectively, measured from a flat surface on which it is placed. The characteristic width 808 of a resonator 102, w_{char}, may be the radius of the smallest possible circle through which the inductive or capacitive element of the magnetic or electric resonator respectively, may pass while traveling in a straight line. For example, the characteristic width 808 of a cylindrical resonator may be the radius of the cylinder.

[00164] In this inventive wireless energy transfer technique, energy may be exchanged efficiently over a wide range of distances, but the technique is distinguished by the ability to exchange useful energy for powering or recharging devices over mid-range distances and between resonators with different physical dimensions, components and orientations. Note that while k may be small in these circumstances, strong coupling and efficient energy transfer may be realized by using high-Q resonators to achieve a high U, $U = k\sqrt{Q_s Q_d}$. That is, increases in Q may be used to at least partially overcome decreases in k, to maintain useful energy transfer efficiencies.

[00165] Note too that while the near-field of a single resonator may be described as omni-directional, the efficiency of the energy exchange between two resonators may depend on the relative position and orientation of the resonators. That is, the efficiency of the energy exchange may be maximized for particular relative orientations of the resonators. The sensitivity of the transfer efficiency to the relative position and orientation of two uncompensated resonators may be captured in the calculation of either k or κ. While coupling may be achieved between resonators that are offset and/or rotated relative to each other, the efficiency of the exchange may depend on the details of the positioning and on any feedback, tuning, and compensation techniques implemented during operation.

[00166] **High-Q Magnetic Resonators**

[00167] In the near-field regime of a sub-wavelength capacitively-loaded loop magnetic resonator ($x \ll \lambda$), the resistances associated with a circular conducting loop inductor composed of N turns of wire whose radius is larger than the skin depth, are approximately $R_{abs} = \sqrt{\mu_o \rho \omega / 2} \cdot Nx / a$ and $R_{rad} = \pi / 6 \cdot \eta_o N^2 (\omega x / c)^4$, where ρ is the resistivity of the conductor material and $\eta_o \approx 120\pi\ \Omega$ is the impedance of free space. . The inductance, L, for such a N-turn loop is approximately N^2 times the inductance of a single-turn loop given previously. The quality factor of such a resonator, $Q = \omega L / (R_{abs} + R_{rad})$, is highest for a particular frequency determined by the system parameters (Fig. 4). As described previously, at lower frequencies the Q is determined primarily by absorption losses and at higher frequencies the Q is determined primarily by radiation losses.

[00168] Note that the formulas given above are approximate and intended to illustrate the functional dependence of R_{abs}, R_{rad} and L on the physical parameters of the structure. More accurate numerical calculations of these parameters that take into account deviations from the strict quasi-static limit, for example a non-uniform current/charge distribution along the conductor, may be useful for the precise design of a resonator structure.

[00169] Note that the absorptive losses may be minimized by using low loss conductors to form the inductive elements. The loss of the conductors may be minimized by using large surface area conductors such as conductive tubing, strapping, strips, machined objects, plates, and the like, by using specially designed conductors such as Litz wire, braided wires, wires of any cross-section, and other conductors with low proximity losses, in which case the frequency scaled behavior described above may be different, and by using low resistivity materials such as high-purity copper and silver, for example. One advantage of using conductive tubing as the conductor at higher operating frequencies is that it may be cheaper and lighter than a similar diameter solid conductor, and may have similar resistance because most of the current is traveling along the outer surface of the conductor owing to the skin effect.

[00170] To get a rough estimate of achievable resonator designs made from copper wire or copper tubing and appropriate for operation in the microwave regime, one may calculate the optimum Q and resonant frequency for a resonator composed of one circular inductive element ($N=1$) of copper wire ($\rho=1.69\cdot 10^{-8}\ \Omega m$) with various cross sections. Then for an inductive element with characteristic size $x=1\ cm$ and conductor diameter a=1 mm, appropriate for a cell phone for example, the quality factor peaks at Q=1225 when $f=380$ MHz. For $x=30\ cm$ and $a=2\ mm$, an inductive element size that might be appropriate for a laptop or a household robot, $Q=1103$ at $f=17$ MHz. For a larger source inductive element that might be located in the ceiling for example, x=1 m and a=4 mm, Q may be as high as Q=1315 at $f=5$ MHz. Note that a number of practical examples yield expected quality factors of $Q\approx 1000\text{-}1500$ at $\lambda/x\approx 50\text{-}80$. Measurements of a wider variety of coil shapes, sizes, materials and operating frequencies than described above show that Q's >100 may be realized for a variety of magnetic resonator structures using commonly available materials.

[00171] As described above, the rate for energy transfer between two resonators of characteristic size x_1 and x_2, and separated by a distance D between their centers, may be given by κ. To give an example of how the defined parameters scale, consider the cell phone, laptop, and ceiling resonator examples from above, at three (3) distances; D/x=10, 8, 6. In the examples considered here, the source and device resonators are the same size, $x_1=x_2$, and shape, and are oriented as shown in Fig. 1(b). In the cell phone example, $\omega/2\kappa$ =3033, 1553, 655 respectively. In the laptop example, $\omega/2\kappa$ =7131, 3651, 1540 respectively and for the ceiling resonator example, $\omega/2\kappa$ =6481, 3318, 1400. The corresponding coupling-to-loss ratios peak at the frequency where the inductive element Q peaks and are κ/Γ =0.4, 0.79, 1.97 and 0.15, 0.3, 0.72 and 0.2, 0.4, 0.94 for the three inductive element sizes and distances described above. An example using different sized inductive elements is that of an x_1=1 m inductor (e.g. source in the ceiling) and an x_2=30 cm inductor (e.g. household robot on the floor) at a distance D=3 m apart (e.g. room height). In this example, the strong-coupling figure of merit, $U=\kappa/\sqrt{\Gamma_1\Gamma_2}$ =0.88, for an efficiency of approximately 14%, at the optimal operating frequency of f=6.4 MHz. Here, the optimal system operating frequency lies between the peaks of the individual resonator Q's.

[00172] Inductive elements may be formed for use in high-Q magnetic resonators. We have demonstrated a variety of high-*Q* magnetic resonators based on copper conductors that are formed into inductive elements that enclose a surface. Inductive elements may be formed using a variety of conductors arranged in a variety of shapes, enclosing any size or shaped area, and they may be single turn or multiple turn elements. Drawings of exemplary inductive elements 900A-B are shown in Fig. 9. The inductive elements may be formed to enclose a circle, a rectangle, a square, a triangle, a shape with rounded corners, a shape that follows the contour of a particular structure or device, a shape that follows, fills, or utilizes, a dedicated space within a structure or device, and the like. The designs may be optimized for size, cost, weight, appearance, performance, and the like.

[00173] These conductors may be bent or formed into the desired size, shape, and number of turns. However, it may be difficult to accurately reproduce conductor shapes and sizes using manual techniques. In addition, it may be difficult to maintain uniform or desired center-to-center spacings between the conductor segments in adjacent turns of the inductive elements. Accurate or uniform spacing may be important in determining the self capacitance of the structure as well as any proximity effect induced increases in AC resistance, for example.

[00174] Molds may be used to replicate inductor elements for high-*Q* resonator designs. In addition, molds may be used to accurately shape conductors into any kind of shape without creating kinks, buckles or other potentially deleterious effects in the conductor. Molds may be used to form the inductor elements and then the inductor elements may be removed from the forms. Once removed, these inductive elements may be built into enclosures or devices that may house the high-Q magnetic resonator. The formed elements may also or instead remain in the mold used to form them.

[00175] The molds may be formed using standard CNC (computer numerical control) routing or milling tools or any other known techniques for cutting or forming grooves in blocks. The molds may also or instead be formed using machining techniques, injection molding techniques, casting techniques, pouring techniques, vacuum techniques, thermoforming techniques, cut-in-place techniques, compression forming techniques and the like.

[00176] The formed element may be removed from the mold or it may remain in the mold. The mold may be altered with the inductive element inside. The mold may be covered, machined, attached, painted and the like. The mold and conductor combination may be integrated into another housing, structure or device. The grooves cut into the molds may be any dimension and may be designed to form conducting tubing, wire, strapping, strips, blocks, and the like into the desired inductor shapes and sizes.

[00177] The inductive elements used in magnetic resonators may contain more than one loop and may spiral inward or outward or up or down or in some combination of directions. In general, the magnetic resonators may have a variety of shapes, sizes and number of turns and they may be composed of a variety of conducing materials.

[00178] The magnetic resonators may be free standing or they may be enclosed in an enclosure, container, sleeve or housing. The magnetic resonators may include the form used to make the inductive element. These various forms and enclosures may be composed of almost any kind of material. Low loss materials such as Teflon, REXOLITE, styrene, and the like may be preferable for some applications. These enclosures may contain fixtures that hold the inductive elements.

[00179] Magnetic resonators may be composed of self-resonant coils of copper wire or copper tubing. Magnetic resonators composed of self resonant conductive wire coils may include a wire of length *l,* and cross section radius *a,* wound into a helical coil of radius *x*, height *h,* and number of turns *N*, which may for example be characterized as $N = \sqrt{l^2 - h^2} / 2\pi x$.

[00180] A magnetic resonator structure may be configured so that *x* is about 30 cm, *h* is about 20 cm, *a* is about 3 mm and *N* is about 5.25, and, during operation, a power source coupled to the magnetic resonator may drive the resonator at a resonant frequency, *f*, where *f* is about 10.6 MHz. Where *x* is about 30 cm, *h* is about 20 cm, *a* is about 1 cm and *N* is about 4, the resonator may be driven at a frequency, *f*, where *f* is about 13.4 MHz. Where *x* is about 10 cm, *h* is about 3 cm, *a* is about 2 mm and *N* is about 6, the resonator may be driven at a frequency, *f*, where *f* is about 21.4 MHz.

[00181] High-Q inductive elements may be designed using printed circuit board traces. Printed circuit board traces may have a variety of advantages compared to mechanically formed inductive elements including that they may be accurately reproduced and easily integrated using established printed circuit board fabrication techniques, that their AC resistance may be lowered using custom designed conductor traces, and that the cost of mass-producing them may be significantly reduced.

[00182] High-Q inductive elements may be fabricated using standard PCB techniques on any PCB material such as FR-4 (epoxy E-glass), multi-functional epoxy, high performance epoxy, bismalaimide triazine/epoxy, polyimide, Cyanate Ester, polytetraflouroethylene (Teflon), FR-2, FR-3, CEM-1, CEM-2, Rogers, Resolute, and the like. The conductor traces may be formed on printed circuit board materials with lower loss tangents.

[00183] The conducting traces may be composed of copper, silver, gold, aluminum, nickel and the like, and they may be composed of paints, inks, or other cured materials. The circuit board may be flexible and it may be a flex-circuit. The conducting traces may be formed by chemical deposition, etching, lithography, spray deposition, cutting, and the like. The conducting traces may be applied to form the desired patterns and they may be formed using crystal and structure growth techniques.

[00184] The dimensions of the conducting traces, as well as the number of layers containing conducting traces, the position, size and shape of those traces and the architecture for interconnecting them may be designed to achieve or optimize certain system specifications such as resonator Q, $Q_{(p)}$, resonator size, resonator material and fabrication costs, U, $U_{(p)}$, and the like.

[00185] As an example, a three-turn high-Q inductive element 1001A was fabricated on a four-layer printed circuit board using the rectangular copper trace pattern as shown in Fig. 10(a). The copper trace is shown in black and the PCB in white. The width and thickness of the copper traces in this example was approximately 1 cm (400 mils) and 43 μm (1.7 mils) respectively. The edge-to-edge spacing between turns of the conducting trace on a single layer was approximately 0.75 cm (300 mils) and each board layer thickness was approximately 100 μm (4 mils). The pattern shown in Fig. 10(a) was repeated on each layer of the board and the conductors were connected in parallel. The outer dimensions of the 3-loop structure were approximately 30 cm by 20 cm. The measured inductance of this PCB loop was 5.3 μH. A magnetic resonator using this inductor element and tunable capacitors had a quality factor, Q, of 550 at its designed resonance frequency of 6.78 MHz. The resonant frequency could be tuned by changing the inductance and capacitance values in the magnetic resonator.

[00186] As another example, a two-turn inductor 1001B was fabricated on a four-layer printed circuit board using the rectangular copper trace pattern shown in Fig. 10(b). The copper trace is shown in black and the PCB in white. The width and height of the copper traces in this example were approximately 0.75 cm (300 mils) and 43 μ m (1.7 mils) respectively. The edge-to-edge spacing between turns of the conducting trace on a single layer was approximately 0.635 cm (250 mils) and each board layer thickness was approximately 100 μ m (4 mils). The pattern shown in Fig. 10(b) was repeated on each layer of the board and the conductors were connected in parallel. The outer dimensions of the two-loop structure were approximately 7.62 cm by 26.7 cm. The measured inductance of this PCB loop was 1.3 μ H. Stacking two boards together with a vertical separation of approximately 0.635 cm (250 mils) and connecting the two boards in series produced a PCB inductor with an inductance of approximately 3.4 μ H. A magnetic resonator using this stacked inductor loop and tunable capacitors had a quality factor, Q, of 390 at its designed resonance frequency of 6.78 MHz. The resonant frequency could be tuned by changing the inductance and capacitance values in the magnetic resonator.

[00187] The inductive elements may be formed using magnetic materials of any size, shape thickness, and the like, and of materials with a wide range of permeability and loss values. These magnetic materials may be solid blocks, they may enclose hollow volumes, they may be formed from many smaller pieces of magnetic material tiled and or stacked together, and they may be integrated with conducting sheets or enclosures made from highly conducting materials. Wires may be wrapped around the magnetic materials to generate the magnetic near-field. These wires may be wrapped around one or more than one axis of the structure. Multiple wires may be wrapped around the magnetic materials and combined in parallel, or in series, or via a switch to form customized near-field patterns.

[00188] The magnetic resonator may include 15 turns of Litz wire wound around a 19.2 cm x 10 cm x 5 mm tiled block of 3F3 ferrite material. The Litz wire may be wound around the ferrite material in any direction or combination of directions to achieve the desire resonator performance. The number of turns of wire, the spacing between the turns, the type of wire, the size and shape of the magnetic materials and the type of magnetic material are all design parameters that may be varied or optimized for different application scenarios.

[00189] **High-*Q* Magnetic resonators using magnetic material structures**

[00190] It may be possible to use magnetic materials assembled to form an open magnetic circuit, albeit one with an air gap on the order of the size of the whole structure, to realize a magnetic resonator structure. In these structures, high conductivity materials are wound around a structure made from magnetic material to form the inductive element of the magnetic resonator. Capacitive elements may be connected to the high conductivity materials, with the resonant frequency then determined as described above. These magnetic resonators have their dipole moment in the plane of the two dimensional resonator structures, rather than perpendicular to it, as is the case for the capacitively-loaded inductor loop resonators.

[00191] A diagram of a single planar resonator structure is shown in Fig. 11(a). The planar resonator structure is constructed of a core of magnetic material 1121, such as ferrite with a loop or loops of conducting material 1122 wrapped around the core 1121. The structure may be used as the source resonator that transfers power and the device resonator that captures energy. When used as a source, the ends of the conductor may be coupled to a power source. Alternating electrical current flowing through the conductor loops excites alternating magnetic fields. When the structure is being used to receive power, the ends of the conductor may be coupled to a power drain or load. Changing magnetic fields induce an electromotive force in the loop or loops of the conductor wound around the core magnetic material. The dipole moment of these types of structures is in the plane of the structures and is, for example, directed along the Y axis for the structure in Figure 11(a). Two such structures have strong coupling when placed substantially in the same plane (i.e. the X,Y plane of Figure 11). The structures of Figure 11(a) have the most favorable orientation when the resonators are aligned in the same plane along their Y axis.

[00192] The geometry and the coupling orientations of the described planar resonators may be preferable for some applications. The planar or flat resonator shape may be easier to integrate into many electronic devices that are relatively flat and planar. The planar resonators may be integrated into the whole back or side of a device without requiring a change in geometry of the device. Due to the flat shape of many devices, the natural position of the devices when placed on a surface is to lay with their largest dimension being parallel to the surface they are placed on. A planar resonator integrated into a flat device is naturally parallel to the plane of the surface and is in a favorable coupling orientation relative to the resonators of other devices or planar resonator sources placed on a flat surface.

[00193] As mentioned, the geometry of the planar resonators may allow easier integration into devices. Their low profile may allow a resonator to be integrated into or as part of a complete side of a device. When a whole side of a device is covered by the resonator, magnetic flux can flow through the resonator core without being obstructed by lossy material that may be part of the device or device circuitry.

[00194] The core of the planar resonator structure may be of a variety of shapes and thicknesses and may be flat or planar such that the minimum dimension does not exceed 30% of the largest dimension of the structure. The core may have complex geometries and may have indentations, notches, ridges, and the like. Geometric enhancements may be used to reduce the coupling dependence on orientation and they may be used to facilitate integration into devices, packaging, packages, enclosures, covers, skins, and the like. Two exemplary variations of core geometries are shown in Figure 11(b). For example, the planar core 1131 may be shaped such that the ends are substantially wider than the middle of the structure to create an indentation for the conductor winding. The core material may be of varying thickness with ends that are thicker and wider than the middle. The core material 1132 may have any number of notches or cutouts 1133 of various depths, width, and shapes to accommodate conductor loops, housing, packaging, and the like.

[00195] The shape and dimensions of the core may be further dictated by the dimensions and characteristics of the device that they are integrated into. The core material may curve to follow the contours of the device, or may require non-symmetric notches or cutouts to allow clearance for parts of the device. The core structure may be a single monolithic piece of magnetic material or may be composed of a plurality of tiles, blocks, or pieces that are arranged together to form the larger structure. The different layers, tiles, blocks, or pieces of the structure may be of similar or may be of different materials. It may be desirable to use materials with different magnetic permeability in different locations of the structure. Core structures with different magnetic permeability may be useful for guiding the magnetic flux, improving coupling, and affecting the shape or extent of the active area of a system.

[00196] The conductor of the planar resonator structure may be wound at least once around the core. In certain circumstances, it may be preferred to wind at least three loops. The conductor can be any good conductor including conducting wire, Litz wire, conducting tubing, sheets, strips, gels, inks, traces and the like.

[00197] The size, shape, or dimensions of the active area of source may be further enhanced, altered, or modified with the use of materials that block, shield, or guide magnetic fields. To create non-symmetric active area around a source once side of the source may be covered with a magnetic shield to reduce the strength of the magnetic fields in a specific direction. The shield may be a conductor or a layered combination of conductor and magnetic material which can be used to guide magnetic fields away from a specific direction. Structures composed of layers of conductors and magnetic materials may be used to reduce energy losses that may occur due to shielding of the source.

[00198] The plurality of planar resonators may be integrated or combined into one planar resonator structure. A conductor or conductors may be wound around a core structure such that the loops formed by the two conductors are not coaxial. An example of such a structure is shown in Figure 12 where two conductors 1201,1202 are wrapped around a planar rectangular core 1203 at orthogonal angles. The core may be rectangular or it may have various geometries with several extensions or protrusions. The protrusions may be useful for wrapping of a conductor, reducing the weight, size, or mass of the core, or may be used to enhance the directionality or omni-directionality of the resonator. A multi wrapped planar resonator with four protrusions is shown by the inner structure 1310 in Figure 13, where four conductors 1301, 1302, 1303, 1304 are wrapped around the core. The core may have extensions 1305,1306,1307,1308 with one or more conductor loops. A single conductor may be wrapped around a core to form loops that are not coaxial. The four conductor loops of Figure 13, for example, may be formed with one continuous piece of conductor, or using two conductors where a single conductor is used to make all coaxial loops.

[00199] Non-uniform or asymmetric field profiles around the resonator comprising a plurality of conductor loops may be generated by driving some conductor loops with non-identical parameters. Some conductor loops of a source resonator with a plurality of conductor loops may be driven by a power source with a different frequency, voltage, power level, duty cycle, and the like all of which may be used to affect the strength of the magnetic field generated by each conductor.

[00200] The planar resonator structures may be combined with a capacitively-loaded inductor resonator coil to provide an omni-directional active area all around, including above and below the source while maintaining a flat resonator structure. As shown in Figure 13, an additional resonator loop coil 1309 comprising of a loop or loops of a conductor, may be placed in a common plane as the planar resonator structure 1310. The outer resonator coil provides an active area that is substantially above and below the source. The resonator coil can be arranged with any number of planar resonator structures and arrangements described herein.

[00201] The planar resonator structures may be enclosed in magnetically permeable packaging or integrated into other devices. The planar profile of the resonators within a single, common plane allows packaging and integration into flat devices. A diagram illustrating the application of the resonators is shown in Figure 14. A flat source 1411 comprising one or more planar resonators 1414 each with one or more conductor loops may transfer power to devices 1412,1413 that are integrated with other planar resonators 1415,1416 and placed within an active area 1417 of the source. The devices may comprise a plurality of planar resonators such that regardless of the orientation of the device with respect to the source the active area of the source does not change. In addition to invariance to rotational misalignment, a flat device comprising of planar resonators may be turned upside down without substantially affecting the active area since the planar resonator is still in the plane of the source.

[00202] Another diagram illustrating a possible use of a power transfer system using the planar resonator structures is shown in Figure 15. A planar source 1521 placed on top of a surface 1525 may create an active area that covers a substantial surface area creating an "energized surface" area. Devices such as computers 1524, mobile handsets 1522, games, and other electronics 1523 that are coupled to their respective planar device resonators may receive energy from the source when placed within the active area of the source, which may be anywhere on top of the surface. Several devices with different dimensions may be placed in the active area and used normally while charging or being powered from the source without having strict placement or alignment constraints. The source may be placed under the surface of a table, countertop, desk, cabinet, and the like, allowing it to be completely hidden while energizing the top surface of the table, countertop, desk, cabinet and the like, creating an active area on the surface that is much larger than the source.

[00203] The source may include a display or other visual, auditory, or vibration indicators to show the direction of charging devices or what devices are being charged, error or problems with charging, power levels, charging time, and the like.

[00204] The source resonators and circuitry may be integrated into any number of other devices. The source may be integrated into devices such as clocks, keyboards, monitors, picture frames, and the like. For example, a keyboard integrated with the planar resonators and appropriate power and control circuitry may be used as a source for devices placed around the keyboard such as computer mice, webcams, mobile handsets, and the like without occupying any additional desk space.

[00205] While the planar resonator structures have been described in the context of mobile devices it should be clear to those skilled in the art that a flat planar source for wireless power transfer with an active area that extends beyond its physical dimensions has many other consumer and industrial applications. The structures and configuration may be useful for a large number of applications where electronic or electric devices and a power source are typically located, positioned, or manipulated in substantially the same plane and alignment. Some of the possible application scenarios include devices on walls, floor, ceilings or any other substantially planar surfaces.

[00206] Flat source resonators may be integrated into a picture frame or hung on a wall thereby providing an active area within the plane of the wall where other electronic devices such as digital picture frames, televisions, lights, and the like can be mounted and powered without wires. Planar resonators may be integrated into a floor resulting in an energized floor or active area on the floor on which devices can be placed to receive power. Audio speakers, lamps, heaters, and the like can be placed within the active are and receive power wirelessly.

[00207] The planar resonator may have additional components coupled to the conductor. Components such as capacitors, inductors, resistors, diodes, and the like may be coupled to the conductor and may be used to adjust or tune the resonant frequency and the impedance matching for the resonators.

[00208] A planar resonator structure of the type described above and shown in Fig. 11(a), may be created, for example, with a quality factor, Q, of 100 or higher and even Q of 1,000 or higher. Energy may be wirelessly transferred from one planar resonator structure to another over a distance larger than the characteristic size of the resonators, as shown in Fig. 11(c).

[00209] In addition to utilizing magnetic materials to realize a structure with properties similar to the inductive element in the magnetic resonators, it may be possible to use a combination of good conductor materials and magnetic material to realize such inductive structures. Fig. 16(a) shows a magnetic resonator structure 1602 that may include one or more enclosures made of high-conductivity materials (the inside of which would be shielded from AC electromagnetic fields generated outside) surrounded by at least one layer of magnetic material and linked by blocks of magnetic material 1604.

A structure may include a high-conductivity sheet of material covered on one side by a layer of magnetic material. The layered structure may instead be applied conformally to an electronic device, so that parts of the device may be covered by the high-conductivity and magnetic material layers, while other parts that need to be easily accessed (such as buttons or screens) may be left uncovered. The structure may also or instead include only layers or bulk pieces of magnetic material. Thus, a magnetic resonator may be incorporated into an existing device without significantly interfering with its existing functions and with little or no need for extensive redesign. Moreover, the layers of good conductor and/or magnetic material may be made thin enough (of the order of a millimeter or less) that they would add little extra weight and volume to the completed device. An oscillating current applied to a length of conductor wound around the structure, as shown by the square loop in the center of the structure in Figure 16 may be used to excite the electromagnetic fields associated with this structure.

[00210] <u>Quality factor of the structure</u>

[00211] A structure of the type described above may be created with a quality factor, Q, of the order of 1,000 or higher. This high-Q is possible even if the losses in the magnetic material are high, if the fraction of magnetic energy within the magnetic material is small compared to the total magnetic energy associated with the object. For structures composed of layers conducting materials and magnetic materials, the losses in the conducting materials may be reduced by the presence of the magnetic materials as described previously. In structures where the magnetic material layer's thickness is of the order of 1/100 of the largest dimension of the system (e.g., the magnetic material may be of the order of 1 mm thick, while the area of the structure is of the order of 10 cm x 10 cm), and the relative permeability is of the order of 1,000, it is possible to make the fraction of magnetic energy contained within the magnetic material only a few hundredths of the total magnetic energy associated with the object or resonator. To see how that comes about, note that the expression for the magnetic energy contained in a volume is $U_m = \int_V d\mathbf{r}\mathbf{B}(\mathbf{r})^2 / (2\mu_r\mu_0)$, so as long as $\boldsymbol{B}$ (rather than $\boldsymbol{H}$) is the main field conserved across the magnetic material-air interface (which is typically the case in open magnetic circuits), the fraction of magnetic energy contained in the high-μ_r region may be significantly reduced compared to what it is in air.

[00212] If the fraction of magnetic energy in the magnetic material is denoted by *frac*, and the loss tangent of the material is $tan\delta$, then the Q of the resonator, assuming the magnetic material is the only source of losses, is $Q=1/(frac \times tan\delta)$. Thus, even for loss tangents as high as 0.1, it is possible to achieve *Q's* of the order of 1,000 for these types of resonator structures.

[00213] If the structure is driven with N turns of wire wound around it, the losses in the excitation inductor loop can be ignored if N is sufficiently high. Fig. 17 shows an equivalent circuit 1700 schematic for these structures and the scaling of the loss mechanisms and inductance with the number of turns, N, wound around a structure made of conducting and magnetic material. If proximity effects can be neglected (by using an appropriate winding, or a wire designed to minimize proximity effects, such as Litz wire and the like), the resistance 1702 due to the wire in the looped conductor scales linearly with the length of the loop, which is in turn proportional to the number of turns. On the other hand, both the equivalent resistance 1708 and equivalent inductance 1704 of these special structures are proportional to the square of the magnetic field inside the structure. Since this magnetic field is proportional to N, the equivalent resistance 1708 and equivalent inductance 1704 are both proportional to N^2. Thus, for large enough N, the resistance 1702 of the wire is much smaller than the equivalent resistance 1708 of the magnetic structure, and the Q of the resonator asymptotes to $Q_{max} = \omega L_\mu / R_\mu$.

[00214] Fig. 16 (a) shows a drawing of a copper and magnetic material structure 1602 driven by a square loop of current around the narrowed segment at the center of the structure 1604 and the magnetic field streamlines generated by this structure 1608. This exemplary structure includes two 20 cm x 8 cm x 2 cm hollow regions enclosed with copper and then completely covered with a 2 mm layer of magnetic material having the properties $\mu_r' = 1{,}400$, $\mu_r'' = 5$, and $\sigma = 0.5$ S/m. These two parallelepipeds are spaced 4 cm apart and are connected by a 2 cm x 4 cm x 2 cm block of the same magnetic material. The excitation loop is wound around the center of this block. At a frequency of 300 kHz, this structure has a calculated Q of 890. The conductor and magnetic material structure may be shaped to optimize certain system parameters. For example, the size of the structure enclosed by the excitation loop may be small to reduce the resistance of the excitation loop, or it may be large to mitigate losses in the magnetic material associated with large magnetic fields. Note that the magnetic streamlines and *Q's* associated with the same structure composed of magnetic material only would be similar to the layer conductor and magnetic material design shown here.

[00215] **<u>Electromagnetic Resonators Interacting with Other Objects</u>**

[00216] For electromagnetic resonators, extrinsic loss mechanisms that perturb the intrinsic Q may include absorption losses inside the materials of nearby extraneous objects and radiation losses related to scattering of the resonant fields from nearby extraneous objects. Absorption losses may be associated with materials that, over the frequency range of interest, have non-zero, but finite, conductivity, σ , (or equivalently a non-zero and finite imaginary part of the dielectric permittivity), such that electromagnetic fields can penetrate it and induce currents in it, which then dissipate energy through resistive losses. An object may be described as lossy if it at least partly includes lossy materials.

[00217] Consider an object including a homogeneous isotropic material of conductivity, σ and magnetic permeability, μ . The penetration depth of electromagnetic fields inside this object is given by the skin depth, $\delta = \sqrt{2/\omega\mu\sigma}$. The power dissipated inside the object, P_d , can be determined from $P_d = \int_V d\mathbf{r}\sigma\,|\,\mathbf{E}\,|^2 = \int_V d\mathbf{r}\,|\,\mathbf{J}\,|^2\,/\sigma$ where we made use of Ohm's law, $\mathbf{J} = \sigma\mathbf{E}$, and where $\mathbf{E}$ is the electric field and $\mathbf{J}$ is the current density.

[00218] If over the frequency range of interest, the conductivity, σ , of the material that composes the object is low enough that the material's skin depth, δ, may be considered long, (i.e. δ is longer than the objects' characteristic size, or δ is longer than the characteristic size of the portion of the object that is lossy) then the electromagnetic fields, $\mathbf{E}$ and $\mathbf{H}$, where $\mathbf{H}$ is the magnetic field, may penetrate significantly into the object. Then, these finite-valued fields may give rise to a dissipated power that scales as $P_d \sim \sigma V_{ol} \left\langle |\mathbf{E}|^2 \right\rangle$, where V_{ol} is the volume of the object that is lossy and $\left\langle |\,\mathbf{E}\,|^2 \right\rangle$ is the spatial average of the electric-field squared, in the volume under consideration. Therefore, in the low-conductivity limit, the dissipated power scales proportionally to the conductivity and goes to zero in the limit of a non-conducting (purely dielectric) material.

[00219] If over the frequency range of interest, the conductivity, σ , of the material that composes the object is high enough that the material's skin depth may be considered short, then the electromagnetic fields, $\mathbf{E}$ and $\mathbf{H}$, may penetrate only a short distance into the object (namely they stay close to the 'skin' of the material, where δ is smaller than the characteristic thickness of the portion of the object that is lossy). In this case, the currents induced inside the material may be concentrated very close to the material surface, approximately within a skin depth, and their magnitude may be approximated by the product of a surface current density (mostly determined by the shape of the incident electromagnetic fields and, as long as the thickness of the conductor is much larger than the skin-depth, independent of frequency and conductivity to first order) $K(x,y)$ (where x and y are coordinates parameterizing the surface) and a function decaying exponentially into the surface: $\exp(-z/\delta)/\delta$ (where z denotes the coordinate locally normal to the surface): $J(x,y,z)=K(x,y)\exp(-z/\delta)/\delta$. Then, the dissipated power, P_d, may be estimated by,

$$P_d = \int_V d\mathrm{r}\,|\mathbf{J}(\mathrm{r})|^2/\sigma \simeq \left(\int_S \mathbf{dxdy}\,|\mathbf{K}(\mathbf{x},\mathbf{y})|^2\right)\left(\int_0^\infty \mathbf{dz}\exp(\mathbf{2z}/\delta)/(\sigma\delta^2)\right)=\sqrt{\mu\omega/8\sigma}\left(\int_S dxdy\,|K(x,y)|^2\right)$$

[00220] Therefore, in the high-conductivity limit, the dissipated power scales inverse proportionally to the square-root of the conductivity and goes to zero in the limit of a perfectly-conducting material.

[00221] If over the frequency range of interest, the conductivity, σ, of the material that composes the object is finite, then the material's skin depth, δ, may penetrate some distance into the object and some amount of power may be dissipated inside the object, depending also on the size of the object and the strength of the electromagnetic fields. This description can be generalized to also describe the general case of an object including multiple different materials with different properties and conductivities, such as an object with an arbitrary inhomogeneous and anisotropic distribution of the conductivity inside the object.

[00222] Note that the magnitude of the loss mechanisms described above may depend on the location and orientation of the extraneous objects relative to the resonator fields as well as the material composition of the extraneous objects. For example, high-conductivity materials may shift the resonant frequency of a resonator and detune it from other resonant objects. This frequency shift may be fixed by applying a feedback mechanism to a resonator that corrects its frequency, such as through changes in the inductance and/or capacitance of the resonator. These changes may be realized using variable capacitors and inductors, in some cases achieved by changes in the geometry of components in the resonators. Other novel tuning mechanisms, described below, may also be used to change the resonator frequency.

[00223] Where external losses are high, the perturbed Q may be low and steps may be taken to limit the absorption of resonator energy inside such extraneous objects and materials. Because of the functional dependence of the dissipated power on the strength of the electric and magnetic fields, one might optimize system performance by designing a system so that the desired coupling is achieved with shorter evanescent resonant field tails at the source resonator and longer at the device resonator, so that the perturbed Q of the source in the presence of other objects is optimized (or vice versa if the perturbed Q of the device needs to be optimized).

[00224] Note that many common extraneous materials and objects such as people, animals, plants, building materials, and the like, may have low conductivities and therefore may have little impact on the wireless energy transfer scheme disclosed here. An important fact related to the magnetic resonator designs we describe is that their electric fields may be confined primarily within the resonator structure itself, so it should be possible to operate within the commonly accepted guidelines for human safety while providing wireless power exchange over mid range distances.

[00225] **<u>Electromagnetic Resonators with Reduced Interactions</u>**

[00226] One frequency range of interest for near-field wireless power transmission is between 10 kHz and 100 MHz. In this frequency range, a large variety of ordinary non-metallic materials, such as for example several types of wood and plastic may have relatively low conductivity, such that only small amounts of power may be dissipated inside them. In addition, materials with low loss tangents, $\tan\Delta$, where $\tan\Delta = \varepsilon'' / \varepsilon'$, and ε'' and ε' are the imaginary and real parts of the permittivity respectively, may also have only small amounts of power dissipated inside them. Metallic materials, such as copper, silver, gold, and the like, with relatively high conductivity, may also have little power dissipated in them, because electromagnetic fields are not able to significantly penetrate these materials, as discussed earlier. These very high and very low conductivity materials, and low loss tangent materials and objects may have a negligible impact on the losses of a magnetic resonator.

[00227] However, in the frequency range of interest, there are materials and objects such as some electronic circuits and some lower-conductivity metals, which may have moderate (in general inhomogeneous and anisotropic) conductivity, and/or moderate to high loss tangents, and which may have relatively high dissipative losses. Relatively larger amounts of power may be dissipated inside them. These materials and objects may dissipate enough energy to reduce $Q_{(p)}$ by non-trivial amounts, and may be referred to as "lossy objects".

[00228] One way to reduce the impact of lossy materials on the $Q_{(p)}$ of a resonator is to use high-conductivity materials to shape the resonator fields such that they avoid the lossy objects. The process of using high-conductivity materials to tailor electromagnetic fields so that they avoid lossy objects in their vicinity may be understood by visualizing high-conductivity materials as materials that deflect or reshape the fields. This picture is qualitatively correct as long as the thickness of the conductor is larger than the skin-depth because the boundary conditions for electromagnetic fields at the surface of a good conductor force the electric field to be nearly completely perpendicular to, and the magnetic field to be nearly completely tangential to, the conductor surface. Therefore, a perpendicular magnetic field or a tangential electric field will be "deflected away" from the conducting surface. Furthermore, even a tangential magnetic field or a perpendicular electric field may be forced to decrease in magnitude on one side and/or in particular locations of the conducting surface, depending on the relative position of the sources of the fields and the conductive surface.

[00229] As an example, Fig. 18 shows a finite element method (FEM) simulation of two high conductivity surfaces 1802 above and below a lossy dielectric material 1804 in an external, initially uniform, magnetic field of frequency f= 6.78 MHz. The system is azimuthally symmetric around the r=0 axis. In this simulation, the lossy dielectric material 1804 is sandwiched between two conductors 1802, shown as the white lines at approximately z = ±0.01m. In the absence of the conducting surfaces above and below the dielectric disk, the magnetic field (represented by the drawn magnetic field lines) would have remained essentially uniform (field lines straight and parallel with the z-axis), indicating that the magnetic field would have passed straight through the lossy dielectric material. In this case, power would have been dissipated in the lossy dielectric disk. In the presence of conducting surfaces, however, this simulation shows the magnetic field is reshaped. The magnetic field is forced to be tangential to surface of the conductor and so is deflected around those conducting surfaces 1802, minimizing the amount of power that may be dissipated in the lossy dielectric material 1804 behind or between the conducting surfaces. As used herein, an axis of electrical symmetry refers to any axis about which a fixed or time-varying electrical or magnetic field is substantially symmetric during an exchange of energy as disclosed herein.

[00230] A similar effect is observed even if only one conducting surface, above or below, the dielectric disk, is used. If the dielectric disk is thin, the fact that the electric field is essentially zero at the surface, and continuous and smooth close to it, means that the electric field is very low everywhere close to the surface (i.e. within the dielectric disk). A single surface implementation for deflecting resonator fields away from lossy objects may be preferred for applications where one is not allowed to cover both sides of the lossy material or object (e.g. an LCD screen). Note that even a very thin surface of conducting material, on the order of a few skin-depths, may be sufficient (the skin depth in pure copper at 6.78 MHz is ~20 μ m, and at 250 kHz is ~100 μ m) to significantly improve the $Q_{(p)}$ of a resonator in the presence of lossy materials.

[00231] Lossy extraneous materials and objects may be parts of an apparatus, in which a high-*Q* resonator is to be integrated. The dissipation of energy in these lossy materials and objects may be reduced by a number of techniques including:

- by positioning the lossy materials and objects away from the resonator, or, in special positions and orientations relative to the resonator.
- by using a high conductivity material or structure to partly or entirely cover lossy materials and objects in the vicinity of a resonator
- by placing a closed surface (such as a sheet or a mesh) of high-conductivity material around a lossy object to completely cover the lossy object and shape the resonator fields such that they avoid the lossy object.
- by placing a surface (such as a sheet or a mesh) of a high-conductivity material around only a portion of a lossy object, such as along the top, the bottom, along the side, and the like, of an object or material.
- by placing even a single surface (such as a sheet or a mesh) of high-conductivity material above or below or on one side of a lossy object to reduce the strength of the fields at the location of the lossy object.

[00232] Fig. 19 shows a capacitively-loaded loop inductor forming a magnetic resonator 102 and a disk-shaped surface of high-conductivity material 1802 that completely surrounds a lossy object 1804 placed inside the loop inductor. Note that some lossy objects may be components, such as electronic circuits, that may need to interact with, communicate with, or be connected to the outside environment and thus cannot be completely electromagnetically isolated. Partially covering a lossy material with high conductivity materials may still reduce extraneous losses while enabling the lossy material or object to function properly.

[00233] Fig. 20 shows a capacitively-loaded loop inductor that is used as the resonator 102 and a surface of high-conductivity material 1802, surrounding only a portion of a lossy object 1804, that is placed inside the inductor loop.

[00234] Extraneous losses may be reduced, but may not be completely eliminated, by placing a single surface of high-conductivity material above, below, on the side, and the like, of a lossy object or material. An example is shown in Fig. 21, where a capacitively-loaded loop inductor is used as the resonator 102 and a surface of high-conductivity material 1802 is placed inside the inductor loop under a lossy object 1804 to reduce the strength of the fields at the location of the lossy object. It may be preferable to cover only one side of a material or object because of considerations of cost, weight, assembly complications, air flow, visual access, physical access, and the like.

[00235] A single surface of high-conductivity material may be used to avoid objects that cannot or should not be covered from both sides (e.g. LCD or plasma screens). Such lossy objects may be avoided using optically transparent conductors. High-conductivity optically opaque materials may instead be placed on only a portion of the lossy object, instead of, or in addition to, optically transparent conductors. The adequacy of single-sided vs. multi-sided covering implementations, and the design trade-offs inherent therein may depend on the details of the wireless energy transfer scenario and the properties of the lossy materials and objects.

[00236] Below we describe an example using high-conductivity surfaces to improve the Q-insensitivity, $\Theta_{(p)}$, of an integrated magnetic resonator used in a wireless energy-transfer system. Fig. 22 shows a wireless projector 2200. The wireless projector may include a device resonator 102C, a projector 2202, a wireless network/video adapter 2204, and power conversion circuits 2208, arranged as shown. The device resonator 102C may include a three-turn conductor loop, arranged to enclose a surface, and a capacitor network 2210. The conductor loop may be designed so that the device resonator 102C has a high Q (e.g., >100) at its operating resonant frequency. Prior to integration in the completely wireless projector 2200, this device resonator

102C has a Q of approximately 477 at the designed operating resonant frequency of 6.78 MHz. Upon integration, and placing the wireless network/video adapter card 2204 in the center of the resonator loop inductor, the resonator $Q_{(integrated)}$ was decreased to approximately 347. At least some of the reduction from Q to $Q_{(integrated)}$ was attributed to losses in the perturbing wireless network/video adapter card. As described above, electromagnetic fields associated with the magnetic resonator 102C may induce currents in and on the wireless network/video adapter card 2204, which may be dissipated in resistive losses in the lossy materials that compose the card. We observed that $Q_{(integrated)}$ of the resonator may be impacted differently depending on the composition, position, and orientation, of objects and materials placed in its vicinity.

[00237] In a completely wireless projector example, covering the network/video adapter card with a thin copper pocket (a folded sheet of copper that covered the top and the bottom of the wireless network/video adapter card, but not the communication antenna) improved the $Q_{(integrated)}$ of the magnetic resonator to a $Q_{(integrated + copper\ pocket)}$ of approximately 444. In other words, most of the reduction in $Q_{(integrated)}$ due to the perturbation caused by the extraneous network/video adapter card could be eliminated using a copper pocket to deflect the resonator fields away from the lossy materials.

[00238] In another completely wireless projector example, covering the network/video adapter card with a single copper sheet placed beneath the card provided a $Q_{(integrated + copper\ sheet)}$ approximately equal to $Q_{(integrated + copper\ pocket)}$. In that example, the high perturbed Q of the system could be maintained with a single high-conductivity sheet used to deflect the resonator fields away from the lossy adapter card.

[00239] It may be advantageous to position or orient lossy materials or objects, which are part of an apparatus including a high-Q electromagnetic resonator, in places where the fields produced by the resonator are relatively weak, so that little or no power may be dissipated in these objects and so that the Q-insensitivity, $\Theta_{(p)}$, may be large. As was shown earlier, materials of different conductivity may respond differently to electric versus magnetic fields. Therefore, according to the conductivity of the extraneous object, the positioning technique may be specialized to one or the other field.

[00240] Fig. 23 shows the magnitude of the electric 2312 and magnetic fields 2314 along a line that contains the diameter of the circular loop inductor and the electric 2318 and magnetic fields 2320 along the axis of the loop inductor for a capacitively-loaded circular loop inductor of wire of radius 30 cm, resonant at 10 MHz. It can be seen that the amplitude of the resonant near-fields reach their maxima close to the wire and decay away from the loop, 2312 , 2314 . In the plane of the loop inductor 2318, 2320, the fields reach a local minimum at the center of the loop. Therefore, given the finite size of the apparatus, it may be that the fields are weakest at the extrema of the apparatus or it may be that the field magnitudes have local minima somewhere within the apparatus. This argument holds for any other type of electromagnetic resonator 102 and any type of apparatus. Examples are shown in Figs. 24a and 24b, where a capacitively-loaded inductor loop forms a magnetic resonator 102 and an extraneous lossy object 1804 is positioned where the electromagnetic fields have minimum magnitude.

[00241] In a demonstration example, a magnetic resonator was formed using a three-turn conductor loop, arranged to enclose a square surface (with rounded corners), and a capacitor network. The Q of the resonator was approximately 619 at the designed operating resonant frequency of 6.78 MHz. The perturbed Q of this resonator depended on the placement of the perturbing object, in this case a pocket projector, relative to the resonator. When the perturbing projector was located inside the inductor loop and at its center or on top of the inductor wire turns, $Q_{(projector)}$ was approximately 96, lower than when the perturbing projector was placed outside of the resonator, in which case $Q_{(projector)}$ was approximately 513. These measurements support the analysis that shows the fields inside the inductor loop may be larger than those outside it, so lossy objects placed inside such a loop inductor may yield lower perturbed *Q's* for the system than when the lossy object is placed outside the loop inductor. Depending on the resonator designs and the material composition and orientation of the lossy object, the arrangement shown in Fig. 24b may yield a higher Q-insensitivity, $\Theta_{(projector)}$, than the arrangement shown in Fig. 24a.

[00242] High-Q resonators may be integrated inside an apparatus. Extraneous materials and objects of high dielectric permittivity, magnetic permeability, or electric conductivity may be part of the apparatus into which a high-Q resonator is to be integrated. For these extraneous materials and objects in the vicinity of a high-Q electromagnetic resonator, depending on their size, position and orientation relative to the resonator, the resonator field-profile may be distorted and deviate significantly from the original unperturbed field-profile of the resonator. Such a distortion of the unperturbed fields of the resonator may significantly decrease the Q to a lower $Q_{(p)}$, even if the extraneous objects and materials are lossless.

[00243] It may be advantageous to position high-conductivity objects, which are part of an apparatus including a high-Q electromagnetic resonator, at orientations such that the surfaces of these objects are, as much as possible, perpendicular to the electric field lines produced by the unperturbed resonator and parallel to the magnetic field lines produced by the unperturbed resonator, thus distorting the resonant field profiles by the smallest amount possible. Other common objects that may be positioned perpendicular to the plane of a magnetic resonator loop include screens (LCD, plasma, etc), batteries, cases, connectors, radiative antennas, and the like. The Q-insensitivity, $\Theta_{(p)}$, of the resonator may be much larger than if the objects were positioned at a different orientation with respect to the resonator fields.

[00244] Lossy extraneous materials and objects, which are not part of the integrated apparatus including a high-Q resonator, may be located or brought in the vicinity of the resonator, for example, during the use of the apparatus. It may be advantageous in certain circumstances to use high conductivity materials to tailor the resonator fields so that they avoid the regions where lossy extraneous objects may be located or introduced to reduce power dissipation in these materials and objects and to increase Q-insensitivity, $\Theta_{(p)}$. An example is shown in Fig. 25, where a capacitively-loaded loop inductor and capacitor are used as the resonator 102 and a surface of high-conductivity material 1802 is placed above the inductor loop to reduce the magnitude of the fields in the region above the resonator, where lossy extraneous objects 1804 may be located or introduced.

[00245] Note that a high-conductivity surface brought in the vicinity of a resonator to reshape the fields may also lead to $Q_{(cond.\ surface)}<Q$. The reduction in the perturbed Q may be due to the dissipation of energy inside the lossy conductor or to the distortion of the unperturbed resonator field profiles associated with matching the field boundary conditions at the surface of the conductor. Therefore, while a high-conductivity surface may be used to reduce the extraneous losses due to dissipation inside an extraneous lossy object, in some cases, especially in some of those where this is achieved by significantly reshaping the electromagnetic fields, using such a high-conductivity surface so that the fields avoid the lossy object may result effectively in $Q_{(p\ +\ cond.\ surface)}<Q_{(p)}$ rather than the desired result $Q_{(p\ +\ cond.\ surface)}>Q_{(p)}$.

[00246] As described above, in the presence of loss inducing objects, the perturbed quality factor of a magnetic resonator may be improved if the electromagnetic fields associated with the magnetic resonator are reshaped to avoid the loss inducing objects. Another way to reshape the unperturbed resonator fields is to use high permeability materials to completely or partially enclose or cover the loss inducing objects, thereby reducing the interaction of the magnetic field with the loss inducing objects.

[00247] Magnetic field shielding has been described previously, for example in *Electrodynamics 3rd Ed.*, Jackson, pp. 201-203. There, a spherical shell of magnetically permeable material was shown to shield its interior from external magnetic fields. For example, if a shell of inner radius a, outer radius b, and relative permeability μ_r, is placed in an initially uniform magnetic field H_0, then the field inside the shell will have a constant magnitude, $9\mu_r H_0 \big/ \left[(2\mu_r+1)(\mu_r+2)-2(a/b)^3(\mu_r-1)^2\right]$, which tends to $9H_0 \big/ 2\mu_r\left(1-(a/b)^3\right)$ if μ_r>>1. This result shows that an incident magnetic field (but not necessarily an incident electric field) may be greatly attenuated inside the shell, even if the shell is quite thin, provided the magnetic permeability is high enough. It may be advantageous in certain circumstances to use high permeability materials to partly or entirely cover lossy materials and objects so that they are avoided by the resonator magnetic fields and so that little or no power is dissipated in these materials and objects. In such an approach, the Q-insensitivity, $\Theta_{(p)}$, may be larger than if the materials and objects were not covered, possibly larger than 1.

[00248] It may be desirable to keep both the electric and magnetic fields away from loss inducing objects. As described above, one way to shape the fields in such a manner is to use high-conductivity surfaces to either completely or partially enclose or cover the loss inducing objects. A layer of magnetically permeable material, also referred to as magnetic material, (any material or meta-material having a non-trivial magnetic permeability), may be placed on or around the high-conductivity surfaces. The additional layer of magnetic material may present a lower reluctance path (compared to free space) for the deflected magnetic field to follow and may partially shield the electric conductor underneath it from the incident magnetic flux. This arrangement may reduce the losses due to induced currents in the high-conductivity surface. Under some circumstances the lower reluctance path presented by the magnetic material may improve the perturbed Q of the structure.

[00249] Fig. 26a shows an axially symmetric FEM simulation of a thin conducting 2604 (copper) disk (20 cm in diameter, 2 cm in height) exposed to an initially uniform, externally applied magnetic field (gray flux lines) along the z-axis. The axis of symmetry is at *r*=0. The magnetic streamlines shown originate at $z=-\infty$, where they are spaced from *r*=3 cm to *r*=10 cm in intervals of 1 cm. The axes scales are in meters. Imagine, for example, that this conducing cylinder encloses loss-inducing objects within an area circumscribed by a magnetic resonator in a wireless energy transfer system such as shown in Fig. 19.

[00250] This high-conductivity enclosure may increase the perturbing *Q* of the lossy objects and therefore the overall perturbed *Q* of the system, but the perturbed *Q* may still be less than the unperturbed *Q* because of induced losses in the conducting surface and changes to the profile of the electromagnetic fields. Decreases in the perturbed *Q* associated with the high-conductivity enclosure may be at least partially recovered by including a layer of magnetic material along the outer surface or surfaces of the high-conductivity enclosure. Fig. 26b shows an axially symmetric FEM simulation of the thin conducting 2604A (copper) disk (20 cm in diameter, 2 cm in height) from Fig. 26a, but with an additional layer of magnetic material placed directly on the outer surface of the high-conductivity enclosure. Note that the presence of the magnetic material may provide a lower reluctance path for the magnetic field, thereby at least partially shielding the underlying conductor and reducing losses due to induced eddy currents in the conductor.

[00251] Fig. 27 depicts a variation (in axi-symmetric view) to the system shown in Fig. 26 where not all of the lossy material 2708 may be covered by a high-conductivity surface 2706. In certain circumstances it may be useful to cover only one side of a material or object, such as due to considerations of cost, weight, assembly complications, air flow, visual access, physical access, and the like. In the exemplary arrangement shown in Fig. 27, only one surface of the lossy material 2708 is covered and the resonator inductor loop is placed on the opposite side of the high-conductivity surface.

[00252] Mathematical models were used to simulate a high-conductivity enclosure made of copper and shaped like a 20 cm diameter by 2 cm high cylindrical disk placed within an area circumscribed by a magnetic resonator whose inductive element was a single-turn wire loop with loop radius *r*=11 cm and wire radius *a*= 1 mm. Simulations for an applied 6.78 MHz electromagnetic field suggest that the perturbing quality factor of this high-conductivity enclosure, $\delta Q_{(enclosure)}$, is 1,870. When the high-conductivity enclosure was modified to include a 0.25 cm-thick layer of magnetic material with real relative permeability, $\mu_r' = 40$, and imaginary relative permeability, $\mu_r'' = 10^{-2}$, simulations suggest the perturbing quality factor is increased to $\delta Q_{(enclosure+magnetic\ material)}$=5,060.

[00253] The improvement in performance due to the addition of thin layers of magnetic material 2702 may be even more dramatic if the high-conductivity enclosure fills a larger portion of the area circumscribed by the resonator's loop inductor 2704. In the example above, if the radius of the inductor loop 2704 is reduced so that it is only 3 mm away from the surface of the high-conductivity enclosure, the perturbing quality factor may be improved from 670 (conducting enclosure only) to 2,730 (conducting enclosure with a thin layer of magnetic material) by the addition of a thin layer of magnetic material 2702around the outside of the enclosure.

[00254] The resonator structure may be designed to have highly confined electric fields, using shielding, or distributed capacitors, for example, which may yield high, even when the resonator is very close to materials that would typically induce loss.

[00255] **Coupled Electromagnetic Resonators**

[00256] The efficiency of energy transfer between two resonators may be determined by the strong-coupling figure-of-merit, $U = \kappa/\sqrt{\Gamma_s \Gamma_d} = \left(2\kappa/\sqrt{\omega_s \omega_d}\right)\sqrt{Q_s Q_d}$. In magnetic resonator implementations the coupling factor between the two resonators may be related to the inductance of the inductive elements in each of the resonators, L_1 and L_2, and the mutual inductance, M, between them by $\kappa_{12} = \omega M / 2\sqrt{L_1 L_2}$. Note that this expression assumes there is negligible coupling through electric-dipole coupling. For capacitively-loaded inductor loop resonators where the inductor loops are formed by circular conducting loops with N turns, separated by a distance D, and oriented as shown in Fig. 1(b), the mutual inductance is $M = \pi / 4 \cdot \mu_o N_1 N_2 (x_1 x_2)^2 / D^3$ where x_1 , N_1 and x_2 , N_2 are the characteristic size and number of turns of the conductor loop of the first and second resonators respectively. Note that this is a quasi-static result, and so assumes that the resonator's size is much smaller than the wavelength and the resonators' distance is much smaller than the wavelength, but also that their distance is at least a few times their size. For these circular resonators operated in the quasi-static limit and at mid-range distances, as described above, $k = 2\kappa / \sqrt{\omega_1 \omega_2} \sim \left(\sqrt{x_1 x_2}/D\right)^3$. Strong coupling (a large

U) between resonators at mid-range distances may be established when the quality factors of the resonators are large enough to compensate for the small k at mid-range distances

[00257] For electromagnetic resonators, if the two resonators include conducting parts, the coupling mechanism may be that currents are induced on one resonator due to electric and magnetic fields generated from the other. The coupling factor may be proportional to the flux of the magnetic field produced from the high-Q inductive element in one resonator crossing a closed area of the high-Q inductive element of the second resonator.

[00258] **<u>Coupled Electromagnetic Resonators with Reduced Interactions</u>**

[00259] As described earlier, a high-conductivity material surface may be used to shape resonator fields such that they avoid lossy objects, p, in the vicinity of a resonator, thereby reducing the overall extraneous losses and maintaining a high Q-insensitivity $\Theta_{(p + cond.\ surface)}$ of the resonator. However, such a surface may also lead to a perturbed coupling factor, $k_{(p + \text{cond. surface})}$, between resonators that is smaller than the perturbed coupling factor, $k_{(p)}$ and depends on the size, position, and orientation of the high-conductivity material relative to the resonators. For example, if high-conductivity materials are placed in the plane and within the area circumscribed by the inductive element of at least one of the magnetic resonators in a wireless energy transfer system, some of the magnetic flux through the area of the resonator, mediating the coupling, may be blocked and k may be reduced.

[00260] Consider again the example of Fig. 19. In the absence of the high-conductivity disk enclosure, a certain amount of the external magnetic flux may cross the circumscribed area of the loop. In the presence of the high-conductivity disk enclosure, some of this magnetic flux may be deflected or blocked and may no longer cross the area of the loop, thus leading to a smaller perturbed coupling factor $k_{12(p + cond.\ surfaces)}$. However, because the deflected magnetic-field lines may follow the edges of the high-conductivity surfaces closely, the reduction in the flux through the loop circumscribing the disk may be less than the ratio of the areas of the face of the disk to the area of the loop.

[00261] One may use high-conductivity material structures, either alone, or combined with magnetic materials to optimize perturbed quality factors, perturbed coupling factors, or perturbed efficiencies.

[00262] Consider the example of Fig. 21. Let the lossy object have a size equal to the size of the capacitively-loaded inductor loop resonator, thus filling its area A 2102. A high-conductivity surface 1802 may be placed under the lossy object 1804. Let this be resonator 1 in a system of two coupled resonators 1 and 2, and let us consider how $U_{12(object + cond.\ surface)}$ scales compared to U_{12} as the area A_s 2104 of the conducting surface increases. Without the conducting surface 1802 below the lossy object 1804, the k-insensitivity, $\beta_{12(object)}$, may be approximately one, but the Q-insensitivity, $\Theta_{1(object)}$, may be small, so the U-insensitivity $\Xi_{12(object)}$ may be small.

[00263] Where the high-conductivity surface below the lossy object covers the entire area of the inductor loop resonator (A_s=A), $k_{12(object + cond.\ surface)}$ may approach zero, because little flux is allowed to cross the inductor loop, so $U_{12(object + cond.\ surface)}$ may approach zero. For intermediate sizes of the high-conductivity surface, the suppression of extrinsic losses and the associated Q-insensitivity, $\Theta_{1(object + cond.\ surface)}$, may be large enough compared to $\Theta_{1(object)}$, while the reduction in coupling may not be significant and the associated k-insensitivity, $\beta_{12(object + cond.\ surface)}$, may be not much smaller than $\beta_{12(object)}$, so that the overall $U_{12(object + cond.\ surface)}$ may be increased compared to $U_{12(object)}$. The optimal degree of avoiding of extraneous lossy objects via high-conductivity surfaces in a system of wireless energy transfer may depend on the details of the system configuration and the application.

[00264] We describe using high-conductivity materials to either completely or partially enclose or cover loss inducing objects in the vicinity of high-Q resonators as one potential method to achieve high perturbed Q's for a system. However, using a good conductor alone to cover the objects may reduce the coupling of the resonators as described above, thereby reducing the efficiency of wireless power transfer. As the area of the conducting surface approaches the area of the magnetic resonator, for example, the perturbed coupling factor, $k_{(p)}$, may approach zero, making the use of the conducting surface incompatible with efficient wireless power transfer.

[00265] One approach to addressing the aforementioned problem is to place a layer of magnetic material around the high-conductivity materials because the additional layer of permeable material may present a lower reluctance path (compared to free space) for the deflected magnetic field to follow and may partially shield the electric conductor underneath it from incident magnetic flux. Under some circumstances the lower reluctance path presented by the magnetic material may improve the electromagnetic coupling of the resonator to other resonators. Decreases in the perturbed coupling factor associated with using conducting materials to tailor resonator fields so that they avoid lossy objects in and around high-Q magnetic resonators may be at least partially recovered by including a layer of magnetic material along the outer surface or surfaces of the conducting materials. The magnetic materials may increase the perturbed coupling factor relative to its initial unperturbed value.

[00266] Note that the simulation results in Fig. 26 show that an incident magnetic field may be deflected less by a layered magnetic material and conducting structure than by a conducting structure alone. If a magnetic resonator loop with a radius only slightly larger than that of the disks shown in Figs. 26(a) and 26(b) circumscribed the disks, it is clear that more flux lines would be captured in the case illustrated in Fig. 26(b) than in Fig. 26(a), and therefore $k_{(disk)}$ would be larger for the case illustrated in Fig. 26(b). Therefore, including a layer of magnetic material on the conducting material may improve the overall system performance. System analyses may be performed to determine whether these materials should be partially, totally, or minimally integrated into the resonator.

[00267] As described above, Fig. 27 depicts a layered conductor 2706 and magnetic material 2702 structure that may be appropriate for use when not all of a lossy material 2708 may be covered by a conductor and/or magnetic material structure. It was shown earlier that for a copper conductor disk with a 20 cm diameter and a 2 cm height, circumscribed by a resonator with an inductor loop radius of 11 cm and a wire radius a=1 mm, the calculated perturbing Q for the copper cylinder was 1,870. If the resonator and the conducting disk shell are placed in a uniform magnetic field (aligned along the axis of symmetry of the inductor loop), we calculate that the copper conductor has an associated coupling factor insensitivity of 0.34. For comparison, we model the same arrangement but include a 0.25 cm-thick layer of magnetic material with a real relative permeability, $\mu_r' = 40$, and an imaginary relative permeability, $\mu_r'' = 10^{-2}$. Using the same model and parameters described above, we find that the coupling factor insensitivity is improved to 0.64 by the addition of the magnetic material to the surface of the conductor.

[00268] Magnetic materials may be placed within the area circumscribed by the magnetic resonator to increase the coupling in wireless energy transfer systems. Consider a solid sphere of a magnetic material with relative permeability, μ_r, placed in an initially uniform magnetic field. In this example, the lower reluctance path offered by the magnetic material may cause the magnetic field to concentrate in the volume of the sphere. We find that the magnetic flux through the area circumscribed by the equator of the sphere is enhanced by a factor of

$3\mu_r/(\mu_r + 2)$, by the addition of the magnetic material. If $\mu_r >> 1$, this enhancement factor may be close to 3.

[00269] One can also show that the dipole moment of a system comprising the magnetic sphere circumscribed by the inductive element in a magnetic resonator would have its magnetic dipole enhanced by the same factor. Thus, the magnetic sphere with high permeability practically triples the dipole magnetic coupling of the resonator. It is possible to keep most of this increase in coupling if we use a spherical shell of magnetic material with inner radius *a,* and outer radius *b,* even if this shell is on top of block or enclosure made from highly conducting materials. In this case, the enhancement in the flux through the equator is

$$\frac{3\mu_r\left(1-\left(\frac{a}{b}\right)^3\right)}{\mu_r\left(1-\left(\frac{a}{b}\right)^3\right)+2\left(1+\frac{1}{2}\left(\frac{a}{b}\right)^3\right)}.$$

For μ_r=1,000 and *(a/b)*=0.99, this enhancement factor is still 2.73, so it possible to significantly improve the coupling even with thin layers of magnetic material.

[00270] As described above, structures containing magnetic materials may be used to realize magnetic resonators. Fig. 16(a) shows a 3 dimensional model of a copper and magnetic material structure 1600 driven by a square loop of current around the choke point at its center. Fig. 16(b) shows the interaction, indicated by magnetic field streamlines , between two identical structures 1600A-B with the same properties as the one shown in Fig. 16(a). Because of symmetry, and to reduce computational complexity, only one half of the system is modeled. If we fix the relative orientation between the two objects and vary their center-to-center distance (the image shown is at a relative separation of 50 cm), we find that, at 300 kHz, the coupling efficiency varies from 87% to 55% as the separation between the structures varies from 30 cm to 60 cm. Each of the example structures shown 1600 A-B includes two 20 cm x 8 cm x 2cm parallelepipeds made of copper joined by a 4 cm x 4 cm x 2 cm block of magnetic material and entirely covered with a 2 mm layer of the same magnetic material (assumed to have μ_r=1,400+j5). Resistive losses in the driving loop are ignored. Each structure has a calculated *Q* of 815.

[00271] <u>ELECTROMAGNETIC RESONATORS AND IMPEDANCE MATCHING</u>

[00272] <u>Impedance Matching Architectures for Low-Loss Inductive Elements</u>

[00273] For purposes of the present discussion, an inductive element may be any coil or loop structure (the 'loop') of any conducting material, with or without a (gapped or ungapped) core made of magnetic material, which may also be coupled inductively or in any other contactless way to other systems. The element is inductive because its impedance, including both the impedance of the loop and the so-called 'reflected' impedances of any potentially coupled systems, has positive reactance, X, and resistance, *R*.

[00274] Consider an external circuit, such as a driving circuit or a driven load or a transmission line, to which an inductive element may be connected. The external circuit (e.g. a driving circuit) may be delivering power to the inductive element and the inductive element may be delivering power to the external circuit (e.g. a driven load). The efficiency and amount of power delivered between the inductive element and the external circuit at a desired frequency may depend on the impedance of the inductive element relative to the properties of the external circuit. Impedance-matching networks and external circuit control techniques may be used to regulate the power delivery between the external circuit and the inductive element, at a desired frequency, *f*.

[00275] The external circuit may be a driving circuit configured to form a amplifier of class A, B, C, D, DE, E, F and the like, and may deliver power at maximum efficiency (namely with minimum losses within the driving circuit) when it is driving a resonant network with specific impedance Z_o^*, where Z_o may be complex and * denotes complex conjugation. The external circuit may be a driven load configured to form a rectifier of class A, B, C, D, DE, E, F and the like, and may receive power at maximum efficiency (namely with minimum losses within the driven load) when it is driven by a resonant network with specific impedance Z_o^*, where Z_o may be complex. The external circuit may be a transmission line with characteristic impedance, Z_o, and may exchange power at maximum efficiency (namely with zero reflections) when connected to an impedance Z_o^*. We will call the characteristic impedance Z_o of an external circuit the complex conjugate of the impedance that may be connected to it for power exchange at maximum efficiency.

[00276] Typically the impedance of an inductive element, $R+jX$, may be much different from Z_o^*. For example, if the inductive element has low loss (a high X/R), its resistance, R, may be much lower than the real part of the characteristic impedance, Z_0, of the external circuit. Furthermore, an inductive element by itself may not be a resonant network. An impedance-matching network connected to an inductive element may typically create a resonant network, whose impedance may be regulated.

[00277] Therefore, an impedance-matching network may be designed to maximize the efficiency of the power delivered between the external circuit and the inductive element (including the reflected impedances of any coupled systems). The efficiency of delivered power may be maximized by matching the impedance of the combination of an impedance-matching network and an inductive element to the characteristic impedance of an external circuit (or transmission line) at the desired frequency.

[00278] An impedance-matching network may be designed to deliver a specified amount of power between the external circuit and the inductive element (including the reflected impedances of any coupled systems). The delivered power may be determined by adjusting the complex ratio of the impedance of the combination of the impedance-matching network and the inductive element to the impedance of the external circuit (or transmission line) at the desired frequency.

[00279] Impedance-matching networks connected to inductive elements may create magnetic resonators. For some applications, such as wireless power transmission using strongly-coupled magnetic resonators, a high Q may be desired for the resonators. Therefore, the inductive element may be chosen to have low losses (high X/R).

[00280] Since the matching circuit may typically include additional sources of loss inside the resonator, the components of the matching circuit may also be chosen to have low losses. Furthermore, in high-power applications and/or due to the high resonator Q, large currents may run in parts of the resonator circuit and large voltages may be present across some circuit elements within the resonator. Such currents and voltages may exceed the specified tolerances for particular circuit elements and may be too high for particular components to withstand. In some cases, it may be difficult to find or implement components, such as tunable capacitors for example, with size, cost and performance (loss and current/voltage-rating) specifications sufficient to realize high-Q and high-power resonator designs for certain applications. We disclose matching circuit designs, methods, implementations and techniques that may preserve the high Q for magnetic resonators, while reducing the component requirements for low loss and/or high current/voltage-rating.

[00281] Matching-circuit topologies may be designed that minimize the loss and current-rating requirements on some of the elements of the matching circuit. The topology of a circuit matching a low-loss inductive element to an impedance, Z_0, may be chosen so that some of its components lie outside the associated high-Q resonator by being in series with the external circuit. The requirements for low series loss or high current-ratings for these components may be reduced. Relieving the low series loss and/or high-current-rating requirement on a circuit element may be particularly useful when the element needs to be variable and/or to have a large voltage-rating and/or low parallel loss.

[00282] Matching-circuit topologies may be designed that minimize the voltage rating requirements on some of the elements of the matching circuit. The topology of a circuit matching a low-loss inductive element to an impedance, Z_0, may be chosen so that some of its components lie outside the associated high-Q resonator by being in parallel with Z_0. The requirements for low parallel loss or high voltage-rating for these components may be reduced. Relieving the low parallel loss and/or high-voltage requirement on a circuit element may be particularly useful when the element needs to be variable and/or to have a large current-rating and/or low series loss.

[00283] The topology of the circuit matching a low-loss inductive element to an external characteristic impedance, Z_0, may be chosen so that the field pattern of the associated resonant mode and thus its high Q are preserved upon coupling of the resonator to the external impedance. Otherwise inefficient coupling to the desired resonant mode may occur (potentially due to coupling to other undesired resonant modes), resulting in an effective lowering of the resonator Q.

[00284] For applications where the low-loss inductive element or the external circuit, may exhibit variations, the matching circuit may need to be adjusted dynamically to match the inductive element to the external circuit impedance, Z_0, at the desired frequency, f. Since there may typically be two tuning objectives, matching or controlling both the real and imaginary part of the impedance level, Z_0, at the desired frequency, f, there may be two variable elements in the matching circuit. For inductive elements, the matching circuit may need to include at least one variable capacitive element.

[00285] A low-loss inductive element may be matched by topologies using two variable capacitors, or two networks of variable capacitors. A variable capacitor may, for example, be a tunable butterfly-type capacitor having, e.g., a center terminal for connection to a ground or other lead of a power source or load, and at least one other terminal across which a capacitance of the tunable butterfly-type capacitor can be varied or tuned, or any other capacitor having a user-configurable, variable capacitance.

[00286] A low-loss inductive element may be matched by topologies using one, or a network of, variable capacitor(s) and one, or a network of, variable inductor(s).

[00287] A low-loss inductive element may be matched by topologies using one, or a network of, variable capacitor(s) and one, or a network of, variable mutual inductance(s), which transformer-couple the inductive element either to an external circuit or to other systems.

[00288] In some cases, it may be difficult to find or implement tunable lumped elements with size, cost and performance specifications sufficient to realize high-Q, high-power, and potentially high-speed, tunable resonator designs. The topology of the circuit matching a variable inductive element to an external circuit may be designed so that some of the variability is assigned to the external circuit by varying the frequency, amplitude, phase, waveform, duty cycle, and the like, of the drive signals applied to transistors, diodes, switches and the like, in the external circuit.

[00289] The variations in resistance, R, and inductance, L, of an inductive element at the resonant frequency may be only partially compensated or not compensated at all. Adequate system performance may thus be preserved by tolerances designed into other system components or specifications. Partial adjustments, realized using fewer tunable components or less capable tunable components, may be sufficient.

[00290] Matching-circuit architectures may be designed that achieve the desired variability of the impedance matching circuit under high-power conditions, while minimizing the voltage/current rating requirements on its tunable elements and achieving a finer (i.e. more precise, with higher resolution) overall tunability. The topology of the circuit matching a variable inductive element to an impedance, Z_0, may include appropriate combinations and placements of fixed and variable elements, so that the voltage/current requirements for the variable components may be reduced and the desired tuning range may be covered with finer tuning resolution. The voltage/current requirements may be reduced on components that are not variable.

[00291] The disclosed impedance matching architectures and techniques may be used to achieve the following:

- To maximize the power delivered to, or to minimize impedance mismatches between, the source low-loss inductive elements (and any other systems wirelessly coupled to them) from the power driving generators.
- To maximize the power delivered from, or to minimize impedance mismatches between, the device low-loss inductive elements (and any other systems wirelessly coupled to them) to the power driven loads.
- To deliver a controlled amount of power to, or to achieve a certain impedance relationship between, the source low-loss inductive elements (and any other systems wirelessly coupled to them) from the power driving generators.
- To deliver a controlled amount of power from, or to achieve a certain impedance relationship between, the device low-loss inductive elements (and any other systems wirelessly coupled to them) to the power driven loads.

[00292] TOPOLOGIES FOR PRESERVATION OF MODE PROFILE (HIGH-*Q*)

[00293] The resonator structure may be designed to be connected to the generator or the load wirelessly (indirectly) or with a hard-wired connection (directly).

[00294] Consider a general indirectly coupled matching topology such as that shown by the block diagram in Fig. 28(a). There, an inductive element 2802 , labeled as (*R,L*) and represented by the circuit symbol for an inductor, may be any of the inductive elements discussed in this disclosure or in the references provided herein, and where an impedance-matching circuit 2402 includes or consists of parts A and B. B may be the part of the matching circuit that connects the impedance 2804 , Z_0, to the rest of the circuit (the combination of A and the inductive element (A+(*R,L*)) via a wireless connection (an inductive or capacitive coupling mechanism).

[00295] The combination of A and the inductive element 2802 may form a resonator 102, which in isolation may support a high-*Q* resonator electromagnetic mode, with an associated current and charge distribution. The lack of a wired connection between the external circuit, Z_0 and B, and the resonator, A + (*R,L*), may ensure that the high-*Q* resonator electromagnetic mode and its current/charge distributions may take the form of its intrinsic (in-isolation) profile, so long as the degree of wireless coupling is not too large. That is, the electromagnetic mode, current/charge distributions, and thus the high-Q of the resonator may be automatically maintained using an indirectly coupled matching topology.

[00296] This matching topology may be referred to as indirectly coupled, or transformer-coupled, or inductively-coupled, in the case where inductive coupling is used between the external circuit and the inductor loop. This type of coupling scenario was used to couple the power supply to the source resonator and the device resonator to the light bulb in the demonstration of wireless energy transfer over mid-range distances described in the referenced *Science* article.

[00297] Next consider examples in which the inductive element may include the inductive element and any indirectly coupled systems. In this case, as disclosed above, and again because of the lack of a wired connection between the external circuit or the coupled systems and the resonator, the coupled systems may not, with good approximation for not-too-large degree of indirect coupling, affect the resonator electromagnetic mode profile and the current/charge distributions of the resonator. Therefore, an indirectly-coupled matching circuit may work equally well for any general inductive element as part of a resonator as well as for inductive elements wirelessly-coupled to other systems, as defined herein. Throughout this disclosure, the matching topologies we disclose refer to matching topologies for a general inductive element of this type, that is, where any additional systems may be indirectly coupled to the low-loss inductive element, and it is to be understood that those additional systems do not greatly affect the resonator electromagnetic mode profile and the current/charge distributions of the resonator.

[00298] Based on the argument above, in a wireless power transmission system of any number of coupled source resonators, device resonators and intermediate resonators the wireless magnetic (inductive) coupling between resonators does not affect the electromagnetic mode profile and the current/charge distributions of each one of the resonators. Therefore, when these resonators have a high (unloaded and unperturbed) Q, their (unloaded and unperturbed) Q may be preserved in the presence of the wireless coupling. (Note that the loaded Q of a resonator may be reduced in the presence of wireless coupling to another resonator, but we may be interested in preserving the unloaded Q, which relates only to loss mechanisms and not to coupling/loading mechanisms.)

[00299] Consider a matching topology such as is shown in Fig. 28(b). The capacitors shown in Fig. 28(b) may represent capacitor circuits or networks. The capacitors shown may be used to form the resonator 102 and to adjust the frequency and/or impedance of the source and device resonators. This resonator 102 may be directly coupled to an impedance, Z_0, using the ports labeled “terminal connections” 2808. Fig. 28(c) shows a generalized directly coupled matching topology, where the impedance-matching circuit 2602 includes or consists of parts A, B and C. Here, circuit elements in A, B and C may be considered part of the resonator 102 as well as part of the impedance matching 2402 (and frequency tuning) topology. B and C may be the parts of the matching circuit 2402 that connect the impedance Z_0 2804 (or the network terminals) to the rest of the circuit (A and the inductive element) via a single wire connection each. Note that B and C could be empty (short-circuits). If we disconnect or open circuit parts B and C (namely those single wire connections), then, the combination of A and the inductive element (R,L) may form the resonator.

[00300] The high-*Q* resonator electromagnetic mode may be such that the profile of the voltage distribution along the inductive element has nodes, namely positions where the voltage is zero. One node may be approximately at the center of the length of the inductive element, such as the center of the conductor used to form the inductive element, (with or without magnetic materials) and at least one other node may be within A. The voltage distribution may be approximately anti-symmetric along the inductive element with respect to its voltage node. A high *Q* may be maintained by designing the matching topology (A, B, C) and/or the terminal voltages (V1, V2) so that this high-*Q* resonator electromagnetic mode distribution may be approximately preserved on the inductive element. This high-*Q* resonator electromagnetic mode distribution may be approximately preserved on the inductive element by preserving the voltage node (approximately at the center) of the inductive element. Examples that achieve these design goals are provided herein.

[00301] A, B, and C may be arbitrary (namely not having any special symmetry), and V1 and V2 may be chosen so that the voltage across the inductive element is symmetric (voltage node at the center inductive). These results may be achieved using simple matching circuits but potentially complicated terminal voltages, because a topology-dependent common-mode signal (V1+V2)/2 may be required on both terminals.

[00302] Consider an 'axis' that connects all the voltage nodes of the resonator, where again one node is approximately at the center of the length of the inductive element and the others within A. (Note that the 'axis' is really a set of points (the voltage nodes) within the electric-circuit topology and may not necessarily correspond to a linear axis of the actual physical structure. The 'axis' may align with a physical axis in cases where the physical structure has symmetry.) Two points of the resonator are electrically symmetric with respect to the 'axis', if the impedances seen between each of the two points and a point on the 'axis', namely a voltage-node point of the resonator, are the same.

[00303] B and C may be the same (C=B), and the two terminals may be connected to any two points of the resonator (A + (R,L)) that are electrically symmetric with respect to the 'axis' defined above and driven with opposite voltages (V2=-V1) as shown in Fig. 28(d). The two electrically symmetric points of the resonator 102 may be two electrically symmetric points on the inductor loop. The two electrically symmetric points of the resonator may be two electrically symmetric points inside A. If the two electrically symmetric points, (to which each of the equal parts B and C is connected), are inside A, A may need to be designed so that these electrically-symmetric points are accessible as connection points within the circuit. This topology may be referred to as a 'balanced drive' topology. These balanced-drive examples may have the advantage that any common-mode signal that may be present on the ground line, due to perturbations at the external circuitry or the power network, for example, may be automatically rejected (and may not reach the resonator). In some balanced-drive examples, this topology may require more components than other topologies.

[00304] In other examples, C may be chosen to be a short-circuit and the corresponding terminal to be connected to ground (V=0) and to any point on the electric-symmetry (zero-voltage) 'axis' of the resonator, and B to be connected to any other point of the resonator not on the electric-symmetry 'axis', as shown in Fig. 28(e). The ground-connected point on the electric-symmetry 'axis' may be the voltage node on the inductive element, approximately at the center of its conductor length. The ground-connected point on the electric-symmetry 'axis' may be inside the circuit A. Where the ground-connected point on the electric-symmetry 'axis' is inside A, A may need to be designed to include one such point on the electrical-symmetric 'axis' that is electrically accessible, namely where connection is possible.

[00305] This topology may be referred to as an 'unbalanced drive' topology. The approximately anti-symmetric voltage distribution of the electromagnetic mode along the inductive element may be approximately preserved, even though the resonator may not be driven exactly symmetrically. The reason is that the high *Q* and the large associated R-vs.-Z_0 mismatch necessitate that a small current may run through B and ground, compared to the much larger current that may flow inside the resonator, (A+(R,L)). In this scenario, the perturbation on the resonator mode may be weak and the location of the voltage node may stay at approximately the center location of the inductive element. These unbalanced-drive examples may have the advantage that they may be achieved using simple matching circuits and that there is no restriction on the driving voltage at the V1 terminal. In some unbalanced-drive examples, additional designs may be required to reduce common-mode signals that may appear at the ground terminal.

[00306] The directly-coupled impedance-matching circuit, generally including or consisting of parts A, B and C, as shown in Fig. 28(c), may be designed so that the wires and components of the circuit do not perturb the electric and magnetic field profiles of the electromagnetic mode of the inductive element and/or the resonator and thus preserve the high resonator *Q*. The wires and metallic components of the circuit may be oriented to be perpendicular to the electric field lines of the electromagnetic mode. The wires and components of the circuit may be placed in regions where the electric and magnetic field of the electromagnetic mode are weak.

[00307] <u>TOPOLOGIES FOR ALLEVIATING LOW-SERIES-LOSS AND HIGH-CURRENT-RATING REQUIREMENTS ON ELEMENTS</u>

[00308] If the matching circuit used to match a small resistance, *R*, of a low-loss inductive element to a larger characteristic impedance, Z_0, of an external circuit may be considered lossless, then $I_{Z_o}^2 Z_o = I_R^2 R \leftrightarrow I_{Z_o} / I_R = \sqrt{R / Z_o}$ and the current flowing through the terminals is much smaller than the current flowing through the inductive element. Therefore, elements connected immediately in series with the terminals (such as in directly-coupled B, C (Fig. 28(c))) may not carry high currents. Then, even if the matching circuit has lossy elements, the resistive loss present in the elements in series with the terminals may not result in a significant reduction in the high-*Q* of the resonator. That is, resistive loss in those series elements may not significantly reduce the efficiency of power transmission from Z_0 to the inductive element or vice versa. Therefore, strict requirements for low-series-loss and/or high current-ratings may not be necessary for these components. In general, such reduced requirements may lead to a wider selection of components that may be designed into the high-*Q* and/or high-power impedance matching and resonator topologies. These reduced requirements may be especially helpful in expanding the variety of variable and/or high voltage and/or low-parallel-loss components that may be used in these high-Q and/or high-power impedance-matching circuits.

[00309] TOPOLOGIES FOR ALLEVIATING LOW-PARALLEL-LOSS AND HIGH-VOLTAGE-RATING REQUIREMENTS ON ELEMENTS

[00310] If, as above, the matching circuit used to match a small resistance, *R*, of a low-loss inductive element to a larger characteristic impedance, Z_0, of an external circuit is lossless, then using the previous analysis,

$$|V_{Z_o} / V_{load}| = |I_{Z_o} Z_o / I_R (R + jX)| \approx \sqrt{R / Z_o} \cdot Z_o / X = \sqrt{Z_o / R} / (X / R),$$

and, for a low-loss (high-*X/R*) inductive element, the voltage across the terminals may be typically much smaller than the voltage across the inductive element. Therefore, elements connected immediately in parallel to the terminals may not need to withstand high voltages. Then, even if the matching circuit has lossy elements, the resistive loss present in the elements in parallel with the terminals may not result in a significant reduction in the high-*Q* of the resonator. That is, resistive loss in those parallel elements may not significantly reduce the efficiency of power transmission from Z_0 to the inductive element or vice versa. Therefore, strict requirements for low-parallel-loss and/or high voltage-ratings may not be necessary for these components. In general, such reduced requirements may lead to a wider selection of components that may be designed into the high-*Q* and/or high-power impedance matching and resonator topologies. These reduced requirements may be especially helpful in expanding the variety of variable and/or high current and/or low-series-loss components that may be used in these high-*Q* and/or high-power impedance-matching and resonator circuits.

[00311] Note that the design principles above may reduce currents and voltages on various elements differently, as they variously suggest the use of networks in series with Z_0 (such as directly-coupled B, C) or the use of networks in parallel with Z_0. The preferred topology for a given application may depend on the availability of low-series-loss/high-current-rating or low-parallel-loss/high-voltage-rating elements.

[00312] COMBINATIONS OF FIXED AND VARIABLE ELEMENTS FOR ACHIEVING FINE TUNABILITY AND ALLEVIATING HIGH-RATING REQUIREMENTS ON VARIABLE ELEMENTS

[00313] **Circuit topologies**

[00314] Variable circuit elements with satisfactory low-loss and high-voltage or current ratings may be difficult or expensive to obtain. In this disclosure, we describe impedance-matching topologies that may incorporate combinations of fixed and variable elements, such that large voltages or currents may be assigned to fixed elements in the circuit, which may be more likely to have adequate voltage and current ratings, and alleviating the voltage and current rating requirements on the variable elements in the circuit.

[00315] Variable circuit elements may have tuning ranges larger than those required by a given impedance-matching application and, in those cases, fine tuning resolution may be difficult to obtain using only such large-range elements. In this disclosure, we describe impedance-matching topologies that incorporate combinations of both fixed and variable elements, such that finer tuning resolution may be accomplished with the same variable elements.

[00316] Therefore, topologies using combinations of both fixed and variable elements may bring two kinds of advantages simultaneously: reduced voltage across, or current through, sensitive tuning components in the circuit and finer tuning resolution. Note that the maximum achievable tuning range may be related to the maximum reduction in voltage across, or current through, the tunable components in the circuit designs.

[00317] **Element topologies**

[00318] A single variable circuit-element (as opposed to the network of elements discussed above) may be implemented by a topology using a combination of fixed and variable components, connected in series or in parallel, to achieve a reduction in the rating requirements of the variable components and a finer tuning resolution. This can be demonstrated mathematically by the fact that:

$$\text{If } x_{|total|} = x_{|fixed|} + x_{|\mathrm{var}iable|},$$
$$\text{then } \Delta x_{|total|} / x_{|total|} = \Delta x_{|\mathrm{variable}|} / (x_{|fixed|} + x_{|\mathrm{variable}|}),$$
$$\text{and } X_{\mathrm{variable}} / X_{total} = X_{\mathrm{variable}} / (X_{fixed} + X_{\mathrm{variable}}),$$

where $x_{|subscript|}$ is any element value (e.g. capacitance, inductance), X is voltage or current, and the "+ sign" denotes the appropriate (series-addition or parallel-addition) combination of elements. Note that the subscript format for $x_{|subscript|}$, is chosen to easily distinguish it from the radius of the area enclosed by a circular inductive element (e.g. x, x_1, etc.).

[00319] Furthermore, this principle may be used to implement a variable electric element of a certain type (e.g. a capacitance or inductance) by using a variable element of a different type, if the latter is combined appropriately with other fixed elements.

[00320] In conclusion, one may apply a topology optimization algorithm that decides on the required number, placement, type and values of fixed and variable elements with the required tunable range as an optimization constraint and the minimization of the currents and/or voltages on the variable elements as the optimization objective.

[00321] EXAMPLES

[00322] In the following schematics, we show different specific topology implementations for impedance matching to and resonator designs for a low-loss inductive element. In addition, we indicate for each topology: which of the principles described above are used, the equations giving the values of the variable elements that may be used to achieve the matching, and the range of the complex impedances that may be matched (using both inequalities and a Smith-chart description). For these examples, we assume that Z_0 is real, but an extension to a characteristic impedance with a non-zero imaginary part is straightforward, as it implies only a small adjustment in the required values of the components of the matching network. We will use the convention that the subscript, n, on a quantity implies normalization to (division by) Z_0.

[00323] Fig. 29 shows two examples of a transformer-coupled impedance-matching circuit, where the two tunable elements are a capacitor and the mutual inductance between two inductive elements. If we define respectively $X_2=\omega L_2$ for Fig. 29(a) and $X_2=\omega L_2-1/\omega C_2$ for Fig. 29(b), and $X \equiv \omega L$, then the required values of the tunable elements are:

$$\omega C_1 = \frac{1}{X + RX_{2n}}$$
$$\omega M = \sqrt{Z_o R(1+X_{2n}^2)}\,.$$

For the topology of Fig. 29(b), an especially straightforward design may be to choose $X_2=0$. In that case, these topologies may match the impedances satisfying the inequalities:

$$R_n > 0, \; X_n > 0,$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 29(c).

[00324] Given a well pre-chosen fixed M, one can also use the above matching topologies with a tunable C_2 instead.

[00325] Fig. 30 shows six examples (a)-(f) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and six examples (h)-(m) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 30(a),(b),(c),(h),(i),(j), a common-mode signal may be required at the two terminals to preserve the voltage node of the resonator at the center of the inductive element and thus the high *Q*. Note that these examples may be described as implementations of the general topology shown in Fig. 28(c). For the symmetric topologies of Figs. 30(d),(e),(f),(k),(l),(m), the two terminals may need to be driven anti-symmetrically (balanced drive) to preserve the voltage node of the resonator at the center of the inductive element and thus the high *Q*. Note that these examples may be described as implementations of the general topology shown in Fig. 28(d). It will be appreciated that a network of capacitors, as used herein, may in general refer to any circuit topology including one or more capacitors, including without limitation any of the circuits specifically disclosed herein using capacitors, or any other equivalent or different circuit structure(s), unless another meaning is explicitly provided or otherwise clear from the context.

[00326] Let us define respectively $Z=R+j\omega L$ for Figs. 30(a),(d),(h),(k), $Z=R+j\omega L+1/j\omega C_3$ for Figs. 30(b),(e),(i),(l), and $Z=(R+j\omega L)\|(1/j\omega C_3)$ for Figs. 30(c),(f),(j),(m), where the symbol "||" means "the parallel combination of", and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.30(a)-(f) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X - \sqrt{X^2 R_n - R^2(1-R_n)}}{X^2 + R^2},$$

$$\omega C_2 = \frac{R_n \omega C_1}{1 - X\omega C_1 - R_n},$$

and these topologies can match the impedances satisfying the inequalities:

$$R_n \le 1,\ X_n \ge \sqrt{R_n(1-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 30(g).
For Figs.30(h)-(m) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X + \sqrt{X^2 R_n - R^2(1-R_n)}}{X^2 + R^2},$$
$$\omega L_2 = -\frac{1 - X\omega C_1 - R_n}{R_n \omega C_1}.$$

[00327] Fig. 31 shows three examples (a)-(c) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and three examples (e)-(g) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 31(a),(b),(c),(e),(f),(g), the ground terminal is connected between two equal-value capacitors, $2C_1$, (namely on the axis of symmetry of the main resonator) to preserve the voltage node of the resonator at the center of the inductive element and thus the high *Q*. Note that these examples may be described as implementations of the general topology shown in Fig. 28(e).

[00328] Let us define respectively Z=R+jωL for Figs. 31(a),(e), Z=R+jωL+1/jωC_3 for Figs. 31(b),(f), and Z=(R+jωL)||(1/jωC_3) for Fig. 31(c),(g), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.31(a)-(c) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X - \frac{1}{2}\sqrt{X^2 R_n - R^2(4-R_n)}}{X^2 + R^2},$$

$$\omega C_2 = \frac{R_n \omega C_1}{1 - X\omega C_1 - \frac{R_n}{2}},$$

and these topologies can match the impedances satisfying the inequalities:

$$R_n \le 1, \quad X_n \ge \sqrt{\frac{R_n}{1-R_n}(2-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 31(d). For Figs.31(e)-(g) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X + \frac{1}{2}\sqrt{X^2 R_n - R^2(4-R_n)}}{X^2 + R^2},$$

$$\omega L_2 = -\frac{1 - X\omega C_1 - \frac{R_n}{2}}{R_n \omega C_1}.$$

[00329] Fig. 32 shows three examples (a)-(c) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and three examples (e)-(g) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 32(a),(b),(c),(e),(f),(g), the ground terminal may be connected at the center of the inductive element to preserve the voltage node of the resonator at that point and thus the high Q. Note that these example may be described as implementations of the general topology shown in Fig. 28(e).

[00330] Let us define respectively Z=R+jωL for Fig. 32(a), Z=R+jωL+1/jωC_3 for Fig. 32(b), and Z=(R+jωL)||(1/jωC_3) for Fig. 32(c), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.32(a)-(c) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X - \sqrt{\frac{X^2 R_n - 2R^2(2-R_n)}{4-R_n}}}{X^2 + R^2},$$
$$\omega C_2 = \frac{R_n \omega C_1}{1 - X\omega C_1 - \frac{R_n}{2} + \frac{R_n X \omega C_1}{2(1+k)}},$$

where k is defined by $M' = -kL'$, where L' is the inductance of each half of the inductor loop and M' is the mutual inductance between the two halves, and these topologies can match the impedances satisfying the inequalities:

$$R_n \leq 2, \quad X_n \geq \sqrt{2R_n(2-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 32(d). For Figs.32(e)-(g) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{X + \sqrt{\dfrac{X^2 R_n - 2R^2(2-R_n)}{4-R_n}}}{X^2 + R^2},$$

[00331] In the circuits of Figs. 30, 31, 32, the capacitor, C_2, or the inductor, L_2, is (or the two capacitors, $2C_2$, or the two inductors, $L_2/2$, are) in series with the terminals and may not need to have very low series-loss or withstand a large current.

[00332] Fig. 33 shows six examples (a)-(f) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and six examples (h)-(m) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 33(a),(b),(c),(h),(i),(j), a common-mode signal may be required at the two terminals to preserve the voltage node of the resonator at the center of the inductive element and thus the high Q. Note that these examples may be described as implementations of the general topology shown in Fig. 28(c), where B and C are short-circuits and A is not balanced. For the symmetric topologies of Figs. 33(d),(e),(f),(k),(l),(m), the two terminals may need to be driven anti-symmetrically (balanced drive) to preserve the voltage node of the resonator at the center of the inductive element and thus the high Q. Note that these examples may be described as implementations of the general topology shown in Fig. 28(d), where B and C are short-circuits and A is balanced.

[00333] Let us define respectively $Z=R+j\omega L$ for Figs. 33(a),(d),(h),(k), $Z=R+j\omega L+1/j\omega C_3$ for Figs. 33(b),(e),(i),(l), and $Z=(R+j\omega L)\|(1/j\omega C_3)$ for Figs. 33(c),(f),(j),(m), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.33(a)-(f) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{1}{X - Z_o\sqrt{R_n(1-R_n)}},$$
$$\omega C_2 = \frac{1}{Z_o}\sqrt{\frac{1}{R_n}-1},$$

and these topologies can match the impedances satisfying the inequalities:

$$R_n \le 1, \quad X_n \ge \sqrt{R_n(1-R_n)}$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 33(g). For Figs.35(h)-(m) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{1}{X + Z_o\sqrt{R_n(1-R_n)}},$$
$$\omega L_2 = \frac{Z_o}{\sqrt{\frac{1}{R_n}-1}}.$$

[00334] Fig. 34 shows three examples (a)-(c) of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors, and three examples (e)-(g) of directly-coupled impedance-matching circuits, where the two tunable elements are one capacitor and one inductor. For the topologies of Figs. 34(a),(b),(c),(e),(f),(g), the ground terminal is connected between two equal-value capacitors, $2C_2$, (namely on the axis of symmetry of the main resonator) to preserve the voltage node of the resonator at the center of the inductive element and thus the high *Q*. Note that these examples may be described as implementations of the general topology shown in Fig. 28(e).

[00335] Let us define respectively $Z=R+j\omega L$ for Fig. 34(a),(e), $Z=R+j\omega L+1/j\omega C_3$ for Fig. 34(b),(f), and $Z=(R+j\omega L)\|(1/j\omega C_3)$ for Fig. 34(c),(g), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, for Figs.34(a)-(c) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{1}{X - Z_o\sqrt{\frac{1-R_n}{R_n}}(2-R_n)},$$

$$\omega C_2 = \frac{1}{2Z_o}\sqrt{\frac{1}{R_n}-1},$$

and these topologies can match the impedances satisfying the inequalities:

$$R_n \le 1,\ \ X_n \ge \sqrt{\frac{R_n}{1-R_n}}(2-R_n)$$

which are shown by the area enclosed by the bold lines on the Smith chart of Fig. 34(d). For Figs.34(e)-(g) the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{1}{X + Z_o\sqrt{\frac{1-R_n}{R_n}}(2-R_n)},$$

$$\omega L_2 = \frac{2Z_o}{\sqrt{\frac{1}{R_n}-1}}.$$

[00336] Fig. 35 shows three examples of directly-coupled impedance-matching circuits, where the two tunable elements are capacitors. For the topologies of Figs. 35, the ground terminal may be connected at the center of the inductive element to preserve the voltage node of the resonator at that point and thus the high Q. Note that these examples may be described as implementations of the general topology shown in Fig. 28(e).

[00337] Let us define respectively Z=R+jωL for Fig. 35(a), Z=R+jωL+1/jωC_3 for Fig. 35(b), and Z=(R+jωL)||(1/jωC_3) for Fig. 35(c), and then $R \equiv \mathrm{Re}\{Z\}$, $X \equiv \mathrm{Im}\{Z\}$. Then, the required values of the tunable elements may be given by:

$$\omega C_1 = \frac{2}{X(1+a) - \sqrt{Z_o R(4-R_n)(1+a^2)}},$$

$$\omega C_2 = \frac{2}{X(1+a) + \sqrt{Z_o R(4-R_n)(1+a^2)}},$$

where $a = \frac{R}{2Z_o - R} \cdot \frac{k}{1+k}$ and k is defined by $M' = -kL'$, where L' is the inductance of each half of the inductive element and M' is the mutual inductance between the two halves. These topologies can match the impedances satisfying the inequalities:

$$R_n \leq 2 \;\&\; \frac{2}{\gamma} \leq R_n \leq 4,$$

$$X_n \geq \sqrt{\frac{R_n(4-R_n)(2-R_n)}{2-\gamma R_n}},$$

where

$$\gamma = \frac{1-6k+k^2}{1+2k+k^2} \leq 1$$

which are shown by the area enclosed by the bold lines on the three Smith charts shown in Fig. 35(d) for k=0, Fig. 35(e) for k=0.05, and Fig. 35(f) for k=1. Note that for 0<k<1 there are two disconnected regions of the Smith chart that this topology can match.

[00338] In the circuits of Figs. 33, 34, 35, the capacitor, C_2, or the inductor, L_2, is (or one of the two capacitors, $2C_2$, or one of the two inductors, $2L_2$, are) in parallel with the terminals and thus may not need to have a high voltage-rating. In the case of two capacitors, $2C_2$, or two inductors, $2L_2$, both may not need to have a high voltage-rating, since approximately the same current flows through them and thus they experience approximately the same voltage across them.

[00339] For the topologies of Figs. 30-35, where a capacitor, C_3, is used, the use of the capacitor, C_3, may lead to finer tuning of the frequency and the impedance. For the topologies of Figs. 30-35, the use of the fixed capacitor, C_3, in series with the inductive element may ensure that a large percentage of the high inductive-element voltage will be across this fixed capacitor, C_3, thus potentially alleviating the voltage rating requirements for the other elements of the impedance matching circuit, some of which may be variable. Whether or not such topologies are preferred depends on the availability, cost and specifications of appropriate fixed and tunable components.

[00340] In all the above examples, a pair of equal-value variable capacitors without a common terminal may be implemented using ganged-type capacitors or groups or arrays of varactors or diodes biased and controlled to tune their values as an ensemble. A pair of equal-value variable capacitors with one common terminal can be implemented using a tunable butterfly-type capacitor or any other tunable or variable capacitor or group or array of varactors or diodes biased and controlled to tune their capacitance values as an ensemble.

[00341] Another criterion which may be considered upon the choice of the impedance matching network is the response of the network to different frequencies than the desired operating frequency. The signals generated in the external circuit, to which the inductive element is coupled, may not be monochromatic at the desired frequency but periodic with the desired frequency, as for example the driving signal of a switching amplifier or the reflected signal of a switching rectifier. In some such cases, it may be desirable to suppress the amount of higher-order harmonics that enter the inductive element (for example, to reduce radiation of these harmonics from this element). Then the choice of impedance matching network may be one that sufficiently suppresses the amount of such harmonics that enters the inductive element.

[00342] The impedance matching network may be such that the impedance seen by the external circuit at frequencies higher than the fundamental harmonic is high, when the external periodic signal is a signal that can be considered to behave as a voltage-source signal (such as the driving signal of a class-D amplifier with a series resonant load), so that little current flows through the inductive element at higher frequencies. Among the topologies of Figs. 30-35, those which use an inductor, L_2, may then be preferable, as this inductor presents a high impedance at high frequencies.

[00343] The impedance matching network may be such that the impedance seen by the external circuit at frequencies higher than the fundamental harmonic is low, when the external periodic signal is a signal that can be considered to behave as a current-source signal, so that little voltage is induced across the inductive element at higher frequencies. Among the topologies of Figs. 30-35, those which use a capacitor, C_2, are then preferable, as this capacitor presents a low impedance at high frequencies.

[00344] Fig. 36 shows four examples of a variable capacitance, using networks of one variable capacitor and the rest fixed capacitors. Using these network topologies, fine tunability of the total capacitance value may be achieved. Furthermore, the topologies of Figs. 36(a),(c),(d), may be used to reduce the voltage across the variable capacitor, since most of the voltage may be assigned across the fixed capacitors.

[00345] Fig. 37 shows two examples of a variable capacitance, using networks of one variable inductor and fixed capacitors. In particular, these networks may provide implementations for a variable reactance, and, at the frequency of interest, values for the variable inductor may be used such that each network corresponds to a net negative variable reactance, which may be effectively a variable capacitance.

[00346] Tunable elements such as tunable capacitors and tunable inductors may be mechanically-tunable, electrically-tunable, thermally-tunable and the like. The tunable elements may be variable capacitors or inductors, varactors, diodes, Schottky diodes, reverse-biased PN diodes, varactor arrays, diode arrays, Schottky diode arrays and the like. The diodes may be Si diodes, GaN diodes, SiC diodes, and the like. GaN and SiC diodes may be particularly attractive for high power applications. The tunable elements may be electrically switched capacitor banks, electrically-switched mechanically-tunable capacitor banks, electrically-switched varactor-array banks, electrically-switched transformer-coupled inductor banks, and the like. The tunable elements may be combinations of the elements listed above.

[00347] As described above, the efficiency of the power transmission between coupled high-Q magnetic resonators may be impacted by how closely matched the resonators are in resonant frequency and how well their impedances are matched to the power supplies and power consumers in the system. Because a variety of external factors including the relative position of extraneous objects or other resonators in the system, or the changing of those relative positions, may alter the resonant frequency and/or input impedance of a high-Q magnetic resonator, tunable impedance networks may be required to maintain sufficient levels of power transmission in various environments or operating scenarios.

[00348] The capacitance values of the capacitors shown may be adjusted to adjust the resonant frequency and/or the impedance of the magnetic resonator. The capacitors may be adjusted electrically, mechanically, thermally, or by any other known methods. They may be adjusted manually or automatically, such as in response to a feedback signal. They may be adjusted to achieve certain power transmission efficiencies or other operating characteristics between the power supply and the power consumer.

[00349] The inductance values of the inductors and inductive elements in the resonator may be adjusted to adjust the frequency and/or impedance of the magnetic resonator. The inductance may be adjusted using coupled circuits that include adjustable components such as tunable capacitors, inductors and switches. The inductance may be adjusted using transformer coupled tuning circuits. The inductance may be adjusted by switching in and out different sections of conductor in the inductive elements and/or using ferro-magnetic tuning and/or mu-tuning, and the like.

[00350] The resonant frequency of the resonators may be adjusted to or may be allowed to change to lower or higher frequencies. The input impedance of the resonator may be adjusted to or may be allowed to change to lower or higher impedance values. The amount of power delivered by the source and/or received by the devices may be adjusted to or may be allowed to change to lower or higher levels of power. The amount of power delivered to the source and/or received by the devices from the device resonator may be adjusted to or may be allowed to change to lower or higher levels of power. The resonator input impedances, resonant frequencies, and power levels may be adjusted depending on the power consumer or consumers in the system and depending on the objects or materials in the vicinity of the resonators. The resonator input impedances, frequencies, and power levels may be adjusted manually or automatically, and may be adjusted in response to feedback or control signals or algorithms.

[00351] Circuit elements may be connected directly to the resonator, that is, by physical electrical contact, for example to the ends of the conductor that forms the inductive element and/or the terminal connectors. The circuit elements may be soldered to, welded to, crimped to, glued to, pinched to, or closely position to the conductor or attached using a variety of electrical components, connectors or connection techniques. The power supplies and the power consumers may be connected to magnetic resonators directly or indirectly or inductively. Electrical signals may be supplied to, or taken from, the resonators through the terminal connections.

[00352] It is to be understood by one of ordinary skill in the art that in real implementations of the principles described herein, there may be an associated tolerance, or acceptable variation, to the values of real components (capacitors, inductors, resistors and the like) from the values calculated via the herein stated equations, to the values of real signals (voltages, currents and the like) from the values suggested by symmetry or anti-symmetry or otherwise, and to the values of real geometric locations of points (such as the point of connection of the ground terminal close to the center of the inductive element or the 'axis' points and the like) from the locations suggested by symmetry or otherwise.

[00353] **Examples**

[00354] SYSTEM BLOCK DIAGRAMS

[00355] We disclose examples of high-Q resonators for wireless power transmission systems that may wirelessly power or charge devices at mid-range distances. High-*Q* resonator wireless power transmission systems also may wirelessly power or charge devices with magnetic resonators that are different in size, shape, composition, arrangement, and the like, from any source resonators in the system.

[00356] Fig. 1(a)(b) shows high level diagrams of two exemplary two-resonator systems. These exemplary systems each have a single source resonator 102S or 104S and a single device resonator 102D or 104D. Fig. 38 shows a high level block diagram of a system with a few more features highlighted. The wirelessly powered or charged device 2310 may include or consist of a device resonator 102D, device power and control circuitry 2304, and the like, along with the device 2308 or devices, to which either DC or AC or both AC and DC power is transferred. The energy or power source for a system may include the source power and control circuitry 2302, a source resonator 102S, and the like. The device 2308 or devices that receive power from the device resonator 102D and power and control circuitry 2304 may be any kind of device 2308 or devices as described previously. The device resonator 102D and circuitry 2304 delivers power to the device/devices 2308 that may be used to recharge the battery of the device/devices, power the device/devices directly, or both when in the vicinity of the source resonator 102S.

[00357] The source and device resonators may be separated by many meters or they may be very close to each other or they may be separated by any distance in between. The source and device resonators may be offset from each other laterally or axially. The source and device resonators may be directly aligned (no lateral offset), or they may be offset by meters, or anything in between. The source and device resonators may be oriented so that the surface areas enclosed by their inductive elements are approximately parallel to each other. The source and device resonators may be oriented so that the surface areas enclosed by their inductive elements are approximately perpendicular to each other, or they may be oriented for any relative angle (0 to 360 degrees) between them.

[00358] The source and device resonators may be free standing or they may be enclosed in an enclosure, container, sleeve or housing. These various enclosures may be composed of almost any kind of material. Low loss tangent materials such as Teflon, REXOLITE, styrene, and the like may be preferable for some applications. The source and device resonators may be integrated in the power supplies and power consumers. For example, the source and device resonators may be integrated into keyboards, computer mice, displays, cell phones, etc. so that they are not visible outside these devices. The source and device resonators may be separate from the power supplies and power consumers in the system and may be connected by a standard or custom wires, cables, connectors or plugs.

[00359] The source 102S may be powered from a number of DC or AC voltage, current or power sources including a USB port of a computer. The source 102S may be powered from the electric grid, from a wall plug, from a battery, from a power supply, from an engine, from a solar cell, from a generator, from another source resonator, and the like. The source power and control circuitry 2302 may include circuits and components to isolate the source electronics from the power source, so that any reflected power or signals are not coupled out through the source input terminals. The source power and control circuits 2302 may include power factor correction circuits and may be configured to monitor power usage for monitoring accounting, billing, control, and like functionalities.

[00360] The system may be operated bi-directionally. That is, energy or power that is generated or stored in a device resonator may be fed back to a power source including the electric grid, a battery, any kind of energy storage unit, and the like. The source power and control circuits may include power factor correction circuits and may be configured to monitor power usage for monitoring accounting, billing, control, and like functionalities for bi-directional energy flow. Wireless energy transfer systems may enable or promote vehicle-to-grid (V2G) applications.

[00361] The source and the device may have tuning capabilities that allow adjustment of operating points to compensate for changing environmental conditions, perturbations, and loading conditions that can affect the operation of the source and device resonators and the efficiency of the energy exchange. The tuning capability may also be used to multiplex power delivery to multiple devices, from multiple sources, to multiple systems, to multiple repeaters or relays, and the like. The tuning capability may be manually controlled, or automatically controlled and may be performed continuously, periodically, intermittently or at scheduled times or intervals.

[00362] The device resonator and the device power and control circuitry may be integrated into any portion of the device, such as a battery compartment, or a device cover or sleeve, or on a mother board, for example, and may be integrated alongside standard rechargeable batteries or other energy storage units. The device resonator may include a device field reshaper which may shield any combination of the device resonator elements and the device power and control electronics from the electromagnetic fields used for the power transfer and which may deflect the resonator fields away from the lossy device resonator elements as well as the device power and control electronics. A magnetic material and/or high-conductivity field reshaper may be used to increase the perturbed quality factor Q of the resonator and increase the perturbed coupling factor of the source and device resonators.

[00363] The source resonator and the source power and control circuitry may be integrated into any type of furniture, structure, mat, rug, picture frame (including digital picture frames, electronic frames), plug-in modules, electronic devices, vehicles, and the like. The source resonator may include a source field reshaper which may shield any combination of the source resonator elements and the source power and control electronics from the electromagnetic fields used for the power transfer and which may deflect the resonator fields away from the lossy source resonator elements as well as the source power and control electronics. A magnetic material and/or high-conductivity field reshaper may be used to increase the perturbed quality factor Q of the resonator and increase the perturbed coupling factor of the source and device resonators.

[00364] A block diagram of the subsystems in an example of a wirelessly powered device is shown in Fig. 39. The power and control circuitry may be designed to transform the alternating current power from the device resonator 102D and convert it to stable direct current power suitable for powering or charging a device. The power and control circuitry may be designed to transform an alternating current power at one frequency from the device resonator to alternating current power at a different frequency suitable for powering or charging a device. The power and control circuitry may include or consist of impedance matching circuitry 2402D, rectification circuitry 2404, voltage limiting circuitry (not shown), current limiting circuitry (not shown), AC-to-DC converter 2408 circuitry, DC-to-DC converter 2408 circuitry, DC-to-AC converter 2408 circuitry, AC-to-AC converter 2408 circuitry, battery charge control circuitry (not shown), and the like.

[00365] The impedance-matching 2402D network may be designed to maximize the power delivered between the device resonator 102D and the device power and control circuitry 2304 at the desired frequency. The impedance matching elements may be chosen and connected such that the high-Q of the resonators is preserved. Depending on the operating conditions, the impedance matching circuitry 2402D may be varied or tuned to control the power delivered from the source to the device, from the source to the device resonator, between the device resonator and the device power and control circuitry, and the like. The power, current and voltage signals may be monitored at any point in the device circuitry and feedback algorithms circuits, and techniques, may be used to control components to achieve desired signal levels and system operation. The feedback algorithms may be implemented using analog or digital circuit techniques and the circuits may include a microprocessor, a digital signal processor, a field programmable gate array processor and the like.

[00366] The third block of Fig. 39 shows a rectifier circuit 2404 that may rectify the AC voltage power from the device resonator into a DC voltage. In this configuration, the output of the rectifier 2404 may be the input to a voltage clamp circuit. The voltage clamp circuit (not shown) may limit the maximum voltage at the input to the DC-to-DC converter 2408D or DC-to-AC converter 2408D. In general, it may be desirable to use a DC-to-DC/AC converter with a large input voltage dynamic range so that large variations in device position and operation may be tolerated while adequate power is delivered to the device. For example, the voltage level at the output of the rectifier may fluctuate and reach high levels as the power input and load characteristics of the device change. As the device performs different tasks it may have varying power demands. The changing power demands can cause high voltages at the output of the rectifier as the load characteristics change. Likewise as the device and the device resonator are brought closer and further away from the source, the power delivered to the device resonator may vary and cause changes in the voltage levels at the output of the rectifier. A voltage clamp circuit may prevent the voltage output from the rectifier circuit from exceeding a predetermined value which is within the operating range of the DC-to-DC/AC converter. The voltage clamp circuitry may be used to extend the operating modes and ranges of a wireless energy transfer system.

[00367] The next block of the power and control circuitry of the device is the DC-to-DC converter 2408D that may produce a stable DC output voltage. The DC-to-DC converter may be a boost converter, buck converter, boost-buck converter, single ended primary inductance converter (SEPIC), or any other DC-DC topology that fits the requirements of the particular application. If the device requires AC power, a DC-to-AC converter may be substituted for the DC-to-DC converter, or the DC-to-DC converter may be followed by a DC-to-AC converter. If the device contains a rechargeable battery, the final block of the device power and control circuitry may be a battery charge control unit which may manage the charging and maintenance of the battery in battery powered devices.

[00368] The device power and control circuitry 2304 may contain a processor 2410D, such as a microcontroller, a digital signal processor, a field programmable gate array processor, a microprocessor, or any other type of processor. The processor may be used to read or detect the state or the operating point of the power and control circuitry and the device resonator. The processor may implement algorithms to interpret and adjust the operating point of the circuits, elements, components, subsystems and resonator. The processor may be used to adjust the impedance matching, the resonator, the DC to DC converters, the DC to AC converters, the battery charging unit, the rectifier, and the like of the wirelessly powered device.

[00369] The processor may have wireless or wired data communication links to other devices or sources and may transmit or receive data that can be used to adjust the operating point of the system. Any combination of power, voltage, and current signals at a single, or over a range of frequencies, may be monitored at any point in the device circuitry. These signals may be monitored using analog or digital or combined analog and digital techniques. These monitored signals may be used in feedback loops or may be reported to the user in a variety of known ways or they may be stored and retrieved at later times. These signals may be used to alert a user of system failures, to indicate performance, or to provide audio, visual, vibrational, and the like, feedback to a user of the system.

[00370] Fig. 40 shows components of source power and control circuitry 2302 of an exemplary wireless power transfer system configured to supply power to a single or multiple devices. The source power and control circuitry 2302 of the exemplary system may be powered from an AC voltage source 2502 such as a home electrical outlet, a DC voltage source such as a battery, a USB port of a computer, a solar cell, another wireless power source, and the like. The source power and control circuitry 2302 may drive the source resonator 102S with alternating current, such as with a frequency greater than 10 kHz and less than 100 MHz. The source power and control circuitry 2302 may drive the source resonator 102S with alternating current of frequency less than less than 10 GHz. The source power and control circuitry 2302 may include a DC-to-DC converter 2408S, an AC-to-DC converter 2408S, or both an AC-to-DC converter 2408S and a DC-to-DC 2408S converter, an oscillator 2508, a power amplifier 2504, an impedance matching network 2402S, and the like.

[00371] The source power and control circuitry 2302 may be powered from multiple AC-or-DC voltage sources 2502 and may contain AC-to-DC and DC-to-DC converters 2408S to provide necessary voltage levels for the circuit components as well as DC voltages for the power amplifiers that may be used to drive the source resonator. The DC voltages may be adjustable and may be used to control the output power level of the power amplifier. The source may contain power factor correction circuitry.

[00372] The oscillator 2508 output may be used as the input to a power amplifier 2504 that drives the source resonator 102S. The oscillator frequency may be tunable and the amplitude of the oscillator signal may be varied as one means to control the output power level from the power amplifier. The frequency, amplitude, phase, waveform, and duty cycle of the oscillator signal may be controlled by analog circuitry, by digital circuitry or by a combination of analog and digital circuitry. The control circuitry may include a processor 2410S, such as a microprocessor, a digital signal processor, a field programmable gate array processor, and the like.

[00373] The impedance matching blocks 2402 of the source and device resonators may be used to tune the power and control circuits and the source and device resonators. For example, tuning of these circuits may adjust for perturbation of the quality factor Q of the source or device resonators due to extraneous objects or changes in distance between the source and device in a system. Tuning of these circuits may also be used to sense the operating environment, control power flow to one or more devices, to control power to a wireless power network, to reduce power when unsafe or failure mode conditions are detected, and the like.

[00374] Any combination of power, voltage, and current signals may be monitored at any point in the source circuitry. These signals may be monitored using analog or digital or combined analog and digital techniques. These monitored signals may be used in feedback circuits or may be reported to the user in a variety of known ways or they may be stored and retrieved at later times. These signals may be used to alert a user to system failures, to alert a user to exceeded safety thresholds, to indicate performance, or to provide audio, visual, vibrational, and the like, feedback to a user of the system.

[00375] The source power and control circuitry may contain a processor. The processor may be used to read the state or the operating point of the power and control circuitry and the source resonator. The processor may implement algorithms to interpret and adjust the operating point of the circuits, elements, components, subsystems and resonator. The processor may be used to adjust the impedance matching, the resonator, the DC-to-DC converters, the AC-to-DC converters, the oscillator, the power amplifier of the source, and the like. The processor and adjustable components of the system may be used to implement frequency and/or time power delivery multiplexing schemes. The processor may have wireless or wired data communication links to devices and other sources and may transmit or receive data that can be used to adjust the operating point of the system.

[00376] Although detailed and specific designs are shown in these block diagrams, it should be clear to those skilled in the art that many different modifications and rearrangements of the components and building blocks are possible within the spirit of the exemplary system. The division of the circuitry was outlined for illustrative purposes and it should be clear to those skilled in the art that the components of each block may be further divided into smaller blocks or merged or shared. In equivalent examples the power and control circuitry may be composed of individual discrete components or larger integrated circuits. For example, the rectifier circuitry may be composed of discrete diodes, or use diodes integrated on a single chip. A multitude of other circuits and integrated devices can be substituted in the design depending on design criteria such as power or size or cost or application. The whole of the power and control circuitry or any portion of the source or device circuitry may be integrated into one chip.

[00377] The impedance matching network of the device and or source may include a capacitor or networks of capacitors, an inductor or networks of inductors, or any combination of capacitors, inductors, diodes, switches, resistors, and the like. The components of the impedance matching network may be adjustable and variable and may be controlled to affect the efficiency and operating point of the system. The impedance matching may be performed by controlling the connection point of the resonator, adjusting the permeability of a magnetic material, controlling a bias field, adjusting the frequency of excitation, and the like. The impedance matching may use or include any number or combination of varactors, varactor arrays, switched elements, capacitor banks, switched and tunable elements, reverse bias diodes, air gap capacitors, compression capacitors, BZT electrically tuned capacitors, MEMS-tunable capacitors, voltage variable dielectrics, transformer coupled tuning circuits, and the like. The variable components may be mechanically tuned, thermally tuned, electrically tuned, piezo-electrically tuned, and the like. Elements of the impedance matching may be silicon devices, gallium nitride devices, silicon carbide devices and the like. The elements may be chosen to withstand high currents, high voltages, high powers, or any combination of current, voltage and power. The elements may be chosen to be high-Q elements.

[00378] The matching and tuning calculations of the source may be performed on an external device through a USB port that powers the device. The device may be a computer a PDA or other computational platform.

[00379] A demonstration system used a source resonator, coupled to a device resonator, to wirelessly power/recharge multiple electronic consumer devices including, but not limited to, a laptop, a DVD player, a projector, a cell-phone, a display, a television, a projector, a digital picture frame, a light, a TV/DVD player, a portable music player, a circuit breaker, a hand-held tool, a personal digital assistant, an external battery charger, a mouse, a keyboard, a camera, an active load, and the like. A variety of devices may be powered simultaneously from a single device resonator. Device resonators may be operated simultaneously as source resonators. The power supplied to a device resonator may pass through additional resonators before being delivered to its intended device resonator.

[00380] *Monitoring, Feedback and Control*

[00381] So-called port parameter measurement circuitry may measure or monitor certain power, voltage, and current, signals in the system and processors or control circuits may adjust certain settings or operating parameters based on those measurements. In addition to these port parameter measurements, the magnitude and phase of voltage and current signals, and the magnitude of the power signals, throughout the system may be accessed to measure or monitor the system performance. The measured signals referred to throughout this disclosure may be any combination of the port parameter signals, as well as voltage signals, current signals, power signals, and the like. These parameters may be measured using analog or digital signals, they may be sampled and processed, and they may be digitized or converted using a number of known analog and digital processing techniques. Measured or monitored signals may be used in feedback circuits or systems to control the operation of the resonators and/or the system. In general, we refer to these monitored or measured signals as reference signals, or port parameter measurements or signals, although they are sometimes also referred to as error signals, monitor signals, feedback signals, and the like. We will refer to the signals that are used to control circuit elements such as the voltages used to drive voltage controlled capacitors as the control signals.

[00382] In some cases the circuit elements may be adjusted to achieve a specified or predetermined impedance value for the source and device resonators. In other cases the impedance may be adjusted to achieve a desired impedance value for the source and device resonators when the device resonator is connected to a power consumer or consumers. In other cases the impedance may be adjusted to mitigate changes in the resonant frequency, or impedance or power level changes owing to movement of the source and/or device resonators, or changes in the environment (such as the movement of interacting materials or objects) in the vicinity of the resonators. In other cases the impedance of the source and device resonators may be adjusted to different impedance values.

[00383] The coupled resonators may be made of different materials and may include different circuits, components and structural designs or they may be the same. The coupled resonators may include performance monitoring and measurement circuitry, signal processing and control circuitry or a combination of measurement and control circuitry. Some or all of the high-Q magnetic resonators may include tunable impedance circuits. Some or all of the high-Q magnetic resonators may include automatically controlled tunable impedance circuits.

[00384] Fig. 41 shows a magnetic resonator with port parameter measurement circuitry 3802 configured to measure certain parameters of the resonator. The port parameter measurement circuitry may measure the input impedance of the structure, or the reflected power. Port parameter measurement circuits may be included in the source and/or device resonator designs and may be used to measure two port circuit parameters such as S-parameters (scattering parameters), Z-parameters (impedance parameters), Y-parameters (admittance parameters), T-parameters (transmission parameters), H-parameters (hybrid parameters), ABCD-parameters (chain, cascade or transmission parameters), and the like. These parameters may be used to describe the electrical behavior of linear electrical networks when various types of signals are applied.

[00385] Different parameters may be used to characterize the electrical network under different operating or coupling scenarios. For example, S-parameters may be used to measure matched and unmatched loads. In addition, the magnitude and phase of voltage and current signals within the magnetic resonators and/or within the sources and devices themselves may be monitored at a variety of points to yield system performance information. This information may be presented to users of the system via a user interface such as a light, a read-out, a beep, a noise, a vibration or the like, or it may be presented as a digital signal or it may be provided to a processor in the system and used in the automatic control of the system. This information may be logged, stored, or may be used by higher level monitoring and control systems.

[00386] Fig. 42 shows a circuit diagram of a magnetic resonator where the tunable impedance network may be realized with voltage controlled capacitors 3902 or capacitor networks. Such an implementation may be adjusted, tuned or controlled by electrical circuits and/or computer processors, such as a programmable voltage source 3908, and the like. For example, the voltage controlled capacitors may be adjusted in response to data acquired by the port parameter measurement circuitry 3802 and processed by a measurement analysis and control algorithm subsystem 3904. Reference signals may be derived from the port parameter measurement circuitry or other monitoring circuitry designed to measure the degree of deviation from a desired system operating point. The measured reference signals may include voltage, current, complex-impedance, reflection coefficient, power levels and the like, at one or several points in the system and at a single frequency or at multiple frequencies.

[00387] The reference signals may be fed to measurement analysis and control algorithm subsystem modules that may generate control signals to change the values of various components in a tunable impedance matching network. The control signals may vary the resonant frequency and/or the input impedance of the magnetic resonator, or the power level supplied by the source, or the power level drawn by the device, to achieve the desired power exchange between power supplies/generators and power drains/loads.

[00388] Adjustment algorithms may be used to adjust the frequency and/or impedance of the magnetic resonators. The algorithms may take in reference signals related to the degree of deviation from a desired operating point for the system and output correction or control signals related to that deviation that control variable or tunable elements of the system to bring the system back towards the desired operating point or points. The reference signals for the magnetic resonators may be acquired while the resonators are exchanging power in a wireless power transmission system, or they may be switched out of the circuit during system operation. Corrections to the system may be applied or performed continuously, periodically, upon a threshold crossing, digitally, using analog methods, and the like.

[00389] Fig. 43 shows an end-to-end wireless power transmission system. Both the source and the device may include port measurement circuitry 3802 and a processor 2410. The box labeled "coupler/switch" 4002 indicates that the port measurement circuitry 3802 may be connected to the resonator 102 by a directional coupler or a switch, enabling the measurement, adjustment and control of the source and device resonators to take place in conjunction with, or separate from, the power transfer functionality.

[00390] The port parameter measurement and/or processing circuitry may reside with some, any, or all resonators in a system. The port parameter measurement circuitry may utilize portions of the power transmission signal or may utilize excitation signals over a range of frequencies to measure the source/device resonator response (i.e. transmission and reflection between any two ports in the system), and may contain amplitude and/or phase information. Such measurements may be achieved with a swept single frequency signal or a multi-frequency signal. The signals used to measure and monitor the resonators and the wireless power transmission system may be generated by a processor or processors and standard input/output (I/O) circuitry including digital to analog converters (DACs), analog to digital converters (ADCs), amplifiers, signal generation chips, passive components and the like. Measurements may be achieved using test equipment such as a network analyzer or using customized circuitry. The measured reference signals may be digitized by ADCs and processed using customized algorithms running on a computer, a microprocessor, a DSP chip, an ASIC, and the like. The measured reference signals may be processed in an analog control loop.

[00391] The measurement circuitry may measure any set of two port parameters such as S-parameters, Y-parameters, Z-parameters, H-parameters, G-parameters, T-parameters, ABCD-parameters, and the like. Measurement circuitry may be used to characterize current and voltage signals at various points in the drive and resonator circuitry, the impedance and/or admittance of the source and device resonators at opposite ends of the system, i.e. looking into the source resonator matching network ("port 1" in Fig. 43) towards the device and vice versa.

[00392] The device may measure relevant signals and/or port parameters, interpret the measurement data, and adjust its matching network to optimize the impedance looking into the coupled system independently of the actions of the source. The source may measure relevant port parameters, interpret the measurement data, and adjust its matching network to optimize the impedance looking into the coupled system independently of the actions of the device.

[00393] Fig. 43 shows a block diagram of a source and device in a wireless power transmission system. The system may be configured to execute a control algorithm that actively adjusts the tuning/matching networks in either of or both the source and device resonators to optimize performance in the coupled system. Port measurement circuitry 3802S may measure signals in the source and communicate those signals to a processor 2410. A processor 2410 may use the measured signals in a performance optimization or stabilization algorithm and generate control signals based on the outputs of those algorithms. Control signals may be applied to variable circuit elements in the tuning/impedance matching circuits 2402S to adjust the source's operating characteristics, such as power in the resonator and coupling to devices. Control signals may be applied to the power supply or generator to turn the supply on or off, to increase or decrease the power level, to modulate the supply signal and the like.

[00394] The power exchanged between sources and devices may depend on a variety of factors. These factors may include the effective impedance of the sources and devices, the Q's of the sources and devices, the resonant frequencies of the sources and devices, the distances between sources and devices, the interaction of materials and objects in the vicinity of sources and devices and the like. The port measurement circuitry and processing algorithms may work in concert to adjust the resonator parameters to maximize power transfer, to hold the power transfer constant, to controllably adjust the power transfer, and the like, under both dynamic and steady state operating conditions.

[00395] Some, all or none of the sources and devices in a system implementation may include port measurement circuitry 3802S and processing 2410 capabilities. Fig. 44 shows an end-to-end wireless power transmission system in which only the source 102S contains port measurement circuitry 3802 and a processor 2410S. In this case, the device resonator 102D operating characteristics may be fixed or may be adjusted by analog control circuitry and without the need for control signals generated by a processor.

[00396] Fig. 45 shows an end-to-end wireless power transmission system. Both the source and the device may include port measurement circuitry 3802 but in the system of Fig. 45, only the source contains a processor 2410S. The source and device may be in communication with each other and the adjustment of certain system parameters may be in response to control signals that have been wirelessly communicated, such as though wireless communications circuitry 4202, between the source and the device. The wireless communication channel 4204 may be separate from the wireless power transfer channel 4208, or it may be the same. That is, the resonators 102 used for power exchange may also be used to exchange information. In some cases, information may be exchanged by modulating a component a source or device circuit and sensing that change with port parameter or other monitoring equipment.

[00397] Implementations where only the source contains a processor 2410 may be beneficial for multi-device systems where the source can handle all of the tuning and adjustment "decisions" and simply communicate the control signals back to the device(s). This implementation may make the device smaller and cheaper because it may eliminate the need for, or reduce the required functionality of, a processor in the device. A portion of or an entire data set from each port measurement at each device may be sent back to the source microprocessor for analysis, and the control instructions may be sent back to the devices. These communications may be wireless communications.

[00398] Fig. 46 shows an end-to-end wireless power transmission system. In this example, only the source contains port measurement circuitry 3802 and a processor 2410S. The source and device may be in communication, such as via wireless communication circuitry 4202, with each other and the adjustment of certain system parameters may be in response to control signals that have been wirelessly communicated between the source and the device.

[00399] Fig. 47 shows coupled electromagnetic resonators 102 whose frequency and impedance may be automatically adjusted using a processor or a computer. Resonant frequency tuning and continuous impedance adjustment of the source and device resonators may be implemented with reverse biased diodes, Schottky diodes and/or varactor elements contained within the capacitor networks shown as C1, C2, and C3 in Fig. 47. The circuit topology that was built and demonstrated and is described here is exemplary and is not meant to limit the discussion of automatic system tuning and control in any way. Other circuit topologies could be utilized with the measurement and control architectures discussed in this disclosure.

[00400] Device and source resonator impedances and resonant frequencies may be measured with a network analyzer 4402A-B, or by other means described above, and implemented with a controller, such as with Lab View 4404. The measurement circuitry or equipment may output data to a computer or a processor that implements feedback algorithms and dynamically adjusts the frequencies and impedances via a programmable DC voltage source.

[00401] In one arrangement, the reverse biased diodes (Schottky, semiconductor junction, and the like) used to realize the tunable capacitance drew very little DC current and could be reverse biased by amplifiers having large series output resistances. This implementation may enable DC control signals to be applied directly to the controllable circuit elements in the resonator circuit while maintaining a very high-Q in the magnetic resonator.

[00402] C2 biasing signals may be isolated from C1 and/or C3 biasing signals with a DC blocking capacitor as shown in Fig. 47, if the required DC biasing voltages are different. The output of the biasing amplifiers may be bypassed to circuit ground to isolate RF voltages from the biasing amplifiers, and to keep non-fundamental RF voltages from being injected into the resonator. The reverse bias voltages for some of the capacitors may instead be applied through the inductive element in the resonator itself, because the inductive element acts as a short circuit at DC.

[00403] The port parameter measurement circuitry may exchange signals with a processor (including any required ADCs and DACs) as part of a feedback or control system that is used to automatically adjust the resonant frequency, input impedance, energy stored or captured by the resonator or power delivered by a source or to a device load. The processor may also send control signals to tuning or adjustment circuitry in or attached to the magnetic resonator.

[00404] When utilizing varactors or diodes as tunable capacitors, it may be beneficial to place fixed capacitors in parallel and in series with the tunable capacitors operating at high reverse bias voltages in the tuning/matching circuits. This arrangement may yield improvements in circuit and system stability and in power handling capability by optimizing the operating voltages on the tunable capacitors.

[00405] Varactors or other reverse biased diodes may be used as a voltage controlled capacitor. Arrays of varactors may be used when higher voltage compliance or different capacitance is required than that of a single varactor component. Varactors may be arranged in an N by M array connected serially and in parallel and treated as a single two terminal component with different characteristics than the individual varactors in the array. For example, an N by N array of equal varactors where components in each row are connected in parallel and components in each column are connected in series may be used as a two terminal device with the same capacitance as any single varactor in the array but with a voltage compliance that is N times that of a single varactor in the array. Depending on the variability and differences of parameters of the individual varactors in the array additional biasing circuits composed of resistors, inductors, and the like may be needed. A schematic of a four by four array of unbiased varactors 4502 that may be suitable for magnetic resonator applications is shown in Fig. 48.

[00406] Further improvements in system performance may be realized by careful selection of the fixed value capacitor(s) that are placed in parallel and/or in series with the tunable (varactor/diode/capacitor) elements. Multiple fixed capacitors that are switched in or out of the circuit may be able to compensate for changes in resonator *Q*'s, impedances, resonant frequencies, power levels, coupling strengths, and the like, that might be encountered in test, development and operational wireless power transfer systems. Switched capacitor banks and other switched element banks may be used to assure the convergence to the operating frequencies and impedance values required by the system design.

[00407] An exemplary control algorithm for isolated and coupled magnetic resonators may be described for the circuit and system elements shown in Fig. 47. One control algorithm first adjusts each of the source and device resonator loops "in isolation", that is, with the other resonators in the system "shorted out" or "removed" from the system. For practical purposes, a resonator can be "shorted out" by making it resonant at a much lower frequency such as by maximizing the value of C1 and/or C3. This step effectively reduces the coupling between the resonators, thereby effectively reducing the system to a single resonator at a particular frequency and impedance.

[00408] Tuning a magnetic resonator in isolation includes varying the tunable elements in the tuning and matching circuits until the values measured by the port parameter measurement circuitry are at their predetermined, calculated or measured relative values. The desired values for the quantities measured by the port parameter measurement circuitry may be chosen based on the desired matching impedance, frequency, strong coupling parameter, and the like. For the exemplary algorithms disclosed below, the port parameter measurement circuitry measures S-parameters over a range of frequencies. The range of frequencies used to characterize the resonators may be a compromise between the system performance information obtained and computation/measurement speed. For the algorithms described below the frequency range may be approximately +/- 20% of the operating resonant frequency.

[00409] Each isolated resonator may be tuned as follows. First, short out the resonator not being adjusted. Next minimize C1, C2, and C3, in the resonator that is being characterized and adjusted. In most cases there will be fixed circuit elements in parallel with C1, C2, and C3, so this step does not reduce the capacitance values to zero. Next, start increasing C2 until the resonator impedance is matched to the "target" real impedance at any frequency in the range of measurement frequencies described above. The initial "target" impedance may be less than the expected operating impedance for the coupled system.

[00410] C2 may be adjusted until the initial "target" impedance is realized for a frequency in the measurement range. Then C1 and/or C3 may be adjusted until the loop is resonant at the desired operating frequency.

[00411] Each resonator may be adjusted according to the above algorithm. After tuning each resonator in isolation, a second feedback algorithm may be applied to optimize the resonant frequencies and/or input impedances for wirelessly transferring power in the coupled system.

[00412] The required adjustments to C1 and/or C2 and/or C3 in each resonator in the coupled system may be determined by measuring and processing the values of the real and imaginary parts of the input impedance from either and/or both "port(s)" shown in Fig. 43. For coupled resonators, changing the input impedance of one resonator may change the input impedance of the other resonator. Control and tracking algorithms may adjust one port to a desired operating point based on measurements at that port, and then adjust the other port based on measurements at that other port. These steps may be repeated until both sides converge to the desired operating point.

[00413] S-parameters may be measured at both the source and device ports and the following series of measurements and adjustments may be made. In the description that follows,

Z_0 is an input impedance and may be the target impedance. In some cases Z_0 is 50 ohms or is near 50 ohms. Z_1 and Z_2 are intermediate impedance values that may be the same value as Z_0 or may be different than Z_0. Re{value} means the real part of a value and Im{value} means the imaginary part of a value.

[00414] An algorithm that may be used to adjust the input impedance and resonant frequency of two coupled resonators is set forth below:

1) Adjust each resonator "in isolation" as described above.

2) Adjust source C1/C3 until, at ω_o, Re{S11} = (Z_1 +/- ε_{Re}) as follows:

- If Re{S11 @ ω_o } > ($Z_1 + \varepsilon_{Re}$), decrease C1/C3. If Re{S11 @ ω_o } < (Zo - ε_{Re}), increase C1/C3.

3) Adjust source C2 until, at ω_o, Im{S11} = (+/- ε_{Im}) as follows:

- If Im{S11 @ ω_o} > ε_{Im}, decrease C2. If Im{S11 @ ω_o} < - ε_{Im} , increase C2.

4) Adjust device C1/C3 until, at ω_0, Re{S22} = (Z_2 +/- ε_{Re}) as follows:

- If Re{S22 @ ω_o} > ($Z_2 + \varepsilon_{Re}$), decrease C1/C3. If Re{S22 @ ω_o} < (Zo - ε_{Re}), increase C1/C3.

5) Adjust device C2 until, at ω_o, Im{S22} = 0 as follows:

- If Im{S22 @ ω_o} > ε_{Im}, decrease C2. If Im{S22 @ ω_o} < -ε_{Im} , increase C2.

[00415] We have achieved a working system by repeating steps 1-4 until both (Re{S11}, Im{S11}) and (Re{S22}, Im{S22}) converge to ((Z_0 +/- ε_{Re}), (+/- ε_{Im})) at ω_o, where Z_0 is the desired matching impedance and ω_o is the desired operating frequency. Here, ε_{Im} represents the maximum deviation of the imaginary part, at ω_o, from the desired value of 0, and ε_{Re} represents the maximum deviation of the real part from the desired value of Z_0. It is understood that ε_{Im} and ε_{Re} can be adjusted to increase or decrease the number of steps to convergence at the potential cost of system performance (efficiency). It is also understood that steps 1-4 can be performed in a variety of sequences and a variety of ways other than that outlined above (i.e. first adjust the source imaginary part, then the source real part; or first adjust the device real part, then the device imaginary part, etc.) The intermediate impedances Z_1 and Z_2 may be adjusted during steps 1-4 to reduce the number of steps required for convergence. The desire or target impedance value may be complex, and may vary in time or under different operating scenarios.

[00416] Steps 1-4 may be performed in any order, in any combination and any number of times. Having described the above algorithm, variations to the steps or the described implementation may be apparent to one of ordinary skill in the art. The algorithm outlined above may be implemented with any equivalent linear network port parameter measurements (i.e., Z-parameters, Y-parameters, T-parameters, H-parameters, ABCD-parameters, etc.) or other monitor signals described above, in the same way that impedance or admittance can be alternatively used to analyze a linear circuit to derive the same result.

[00417] The resonators may need to be retuned owing to changes in the "loaded" resistances, Rs and Rd, caused by changes in the mutual inductance M (coupling) between the source and device resonators. Changes in the inductances, Ls and Ld, of the inductive elements themselves may be caused by the influence of external objects, as discussed earlier, and may also require compensation. Such variations may be mitigated by the adjustment algorithm described above.

[00418] A directional coupler or a switch may be used to connect the port parameter measurement circuitry to the source resonator and tuning/adjustment circuitry. The port parameter measurement circuitry may measure properties of the magnetic resonator while it is exchanging power in a wireless power transmission system, or it may be switched out of the circuit during system operation. The port parameter measurement circuitry may measure the parameters and the processor may control certain tunable elements of the magnetic resonator at start-up, or at certain intervals, or in response to changes in certain system operating parameters.

[00419] A wireless power transmission system may include circuitry to vary or tune the impedance and/or resonant frequency of source and device resonators. Note that while tuning circuitry is shown in both the source and device resonators, the circuitry may instead be included in only the source or the device resonators, or the circuitry may be included in only some of the source and/or device resonators. Note too that while we may refer to the circuitry as "tuning" the impedance and or resonant frequency of the resonators, this tuning operation simply means that various electrical parameters such as the inductance or capacitance of the structure are being varied. In some cases, these parameters may be varied to achieve a specific predetermined value, in other cases they may be varied in response to a control algorithm or to stabilize a target performance value that is changing. In some cases, the parameters are varied as a function of temperature, of other sources or devices in the area, of the environment, at the like.

[00420] **Applications**

[00421] For each listed application, it will be understood by one of ordinary skill-in-the-art that there are a variety of ways that the resonator structures used to enable wireless power transmission may be connected or integrated with the objects that are supplying or being powered. The resonator may be physically separate from the source and device objects. The resonator may supply or remove power from an object using traditional inductive techniques or through direct electrical connection, with a wire or cable for example. The electrical connection may be from the resonator output to the AC or DC power input port on the object. The electrical connection may be from the output power port of an object to the resonator input.

[00422] FIG. 49 shows a source resonator 4904 that is physically separated from a power supply and a device resonator 4902 that is physically separated from the device 4900, in this illustration a laptop computer. Power may be supplied to the source resonator, and power may be taken from the device resonator directly, by an electrical connection. One of ordinary skill in the art will understand from the materials incorporated by reference that the shape, size, material composition, arrangement, position and orientation of the resonators above are provided by way of non-limiting example, and that a wide variation in any and all of these parameters could be supported by the disclosed technology for a variety of applications.

[00423] Continuing with the example of the laptop, and without limitation, the device resonator may be physically connected to the device it is powering or charging. For example, as shown in FIG. 50a and FIG. 50b, the device resonator 5002 may be (a) integrated into the housing of the device 5000 or (b) it may be attached by an adapter. The resonator 5002 may (FIG. 50b-d) or may not (FIG. 50a) be visible on the device. The resonator may be affixed to the device, integrated into the device, plugged into the device, and the like.

[00424] The source resonator may be physically connected to the source supplying the power to the system. As described above for the devices and device resonators, there are a variety of ways the resonators may be attached to, connected to or integrated with the power supply. One of ordinary skill in the art will understand that there are a variety of ways the resonators may be integrated in the wireless power transmission system, and that the sources and devices may utilize similar or different integration techniques.

[00425] Continuing again with the example of the laptop computer, and without limitation, the laptop computer may be powered, charged or recharged by a wireless power transmission system. A source resonator may be used to supply wireless power and a device resonator may be used to capture the wireless power. A device resonator 5002 may be integrated into the edge of the screen (display) as illustrated in FIG. 50d, and/or into the base of the laptop as illustrated in FIG. 50c. The source resonator 5002 may be integrated into the base of the laptop and the device resonator may be integrated into the edge of the screen. The resonators may also or instead be affixed to the power source and/or the laptop. The source and device resonators may also or instead be physically separated from the power supply and the laptop and may be electrically connected by a cable. The source and device resonators may also or instead be physically separated from the power supply and the laptop and may be electrically coupled using a traditional inductive technique. One of ordinary skill in the art will understand that, while the preceding examples relate to wireless power transmission to a laptop, that the methods and systems disclosed for this application may be suitably adapted for use with other electrical or electronic devices. In general, the source resonator may be external to the source and supplying power to a device resonator that in turn supplies power the device, or the source resonator may be connected to the source and supplying power to a device resonator that in turn supplies power to a portion of the device, or the source resonator may internal to the source and supplying power to a device resonator that in turn supplies power to a portion of the device, as well as any combination of these.

[00426] A system or method disclosed herein may provide power to an electrical or electronics device, such as, and not limited to, phones, cell phones, cordless phones, smart phones, PDAs, audio devices, music players, MP3 players, radios, portable radios and players, wireless headphones, wireless headsets, computers, laptop computers, wireless keyboards, wireless mouse, televisions, displays, flat screen displays, computer displays, displays embedded in furniture, digital picture frames, electronic books, (e.g. the Kindle, e-ink books, magazines, and the like), remote control units (also referred to as controllers, game controllers, commanders, clickers, and the like, and used for the remote control of a plurality of electronics devices, such as televisions, video games, displays, computers, audio visual equipment, lights, and the like), lighting devices, cooling devices, air circulation devices, purification devices, personal hearing aids, power tools, security systems, alarms, bells, flashing lights, sirens, sensors, loudspeakers, electronic locks, electronic keypads, light switches, other electrical switches, and the like. Here the term electronic lock is used to indicate a door lock which operates electronically (e.g. with electronic combo-key, magnetic card, RFID card, and the like) which is placed on a door instead of a mechanical key-lock. Such locks are often battery operated, risking the possibility that the lock might stop working when a battery dies, leaving the user locked-out. This may be avoided where the battery is either charged or completely replaced by a wireless power transmission implementation as described herein.

[00427] Here, the term light switch (or other electrical switch) is meant to indicate any switch (e.g. on a wall of a room) in one part of the room that turns on/off a device (e.g. light fixture at the center of the ceiling) in another part of the room. To install such a switch by direct connection, one would have to run a wire all the way from the device to the switch. Once such a switch is installed at a particular spot, it may be very difficult to move. Alternately, one can envision a 'wireless switch', where "wireless" means the switching (on/off) commands are communicated wirelessly, but such a switch has traditionally required a battery for operation. In general, having too many battery operated switches around a house may be impractical, because those many batteries will need to be replaced periodically. So, a wirelessly communicating switch may be more convenient, provided it is also wirelessly powered. For example, there already exist communications wireless door-bells that are battery powered, but where one still has to replace the battery in them periodically. The remote doorbell button may be made to be completely wireless, where there may be no need to ever replace the battery again. Note that here, the term 'cordless' or 'wireless' or 'communications wireless' is used to indicate that there is a cordless or wireless communications facility between the device and another electrical component, such as the base station for a cordless phone, the computer for a wireless keyboard, and the like. One skilled in the art will recognize that any electrical or electronics device may include a wireless communications facility, and that the systems and methods described herein may be used to add wireless power transmission to the device. As described herein, power to the electrical or electronics device may be delivered from an external or internal source resonator, and to the device or portion of the device. Wireless power transmission may significantly reduce the need to charge and/or replace batteries for devices that enter the near vicinity of the source resonator and thereby may reduce the downtime, cost and disposal issues often associated with batteries.

[00428] The systems and methods described herein may provide power to lights without the need for either wired power or batteries. That is, the systems and methods described herein may provide power to lights without wired connection to any power source, and provide the energy to the light non-radiatively across mid-range distances, such as across a distance of a quarter of a meter, one meter, three meters, and the like. A 'light' as used herein may refer to the light source itself, such as an incandescent light bulb, florescent light bulb lamps, Halogen lamps, gas discharge lamps, fluorescent lamps, neon lamps, high-intensity discharge lamps, sodium vapor lamps, Mercury-vapor lamps, electroluminescent lamps, light emitting diodes (LED) lamps, and the like; the light as part of a light fixture, such as a table lamp, a floor lamp, a ceiling lamp, track lighting, recessed light fixtures, and the like; light fixtures integrated with other functions, such as a light/ceiling fan fixture, and illuminated picture frame, and the like. As such, the systems and methods described herein may reduce the complexity for installing a light, such as by minimizing the installation of electrical wiring, and allowing the user to place or mount the light with minimal regard to sources of wired power. For instance, a light may be placed anywhere in the vicinity of a source resonator, where the source resonator may be mounted in a plurality of different places with respect to the location of the light, such as on the floor of the room above, (e.g. as in the case of a ceiling light and especially when the room above is the attic); on the wall of the next room, on the ceiling of the room below, (e.g. as in the case of a floor lamp); in a component within the room or in the infrastructure of the room as described herein; and the like. For example, a light/ceiling fan combination is often installed in a master bedroom, and the master bedroom often has the attic above it. In this instance a user may more easily install the light/ceiling fan combination in the master bedroom, such as by simply mounting the light/ceiling fan combination to the ceiling, and placing a source coil (plugged into the house wired AC power) in the attic above the mounted fixture. In another example, the light may be an external light, such as a flood light or security light, and the source resonator mounted inside the structure. This way of installing lighting may be particularly beneficial to users who rent their homes, because now they may be able to mount lights and such other electrical components without the need to install new electrical wiring. The control for the light may also be communicated by near-field communications as described herein, or by traditional wireless communications methods.

[00429] The systems and methods described herein may provide power from a source resonator to a device resonator that is either embedded into the device component, or outside the device component, such that the device component may be a traditional electrical component or fixture. For instance, a ceiling lamp may be designed or retrofitted with a device resonator integrated into the fixture, or the ceiling lamp may be a traditional wired fixture, and plugged into a separate electrical facility equipped with the device resonator. In an example, the electrical facility may be a wireless junction box designed to have a device resonator for receiving wireless power, say from a source resonator placed on the floor of the room above (e.g. the attic), and which contains a number of traditional outlets that are powered from the device resonator. The wireless junction box, mounted on the ceiling, may now provide power to traditional wired electrical components on the ceiling (e.g. a ceiling light, track lighting, a ceiling fan). Thus, the ceiling lamp may now be mounted to the ceiling without the need to run wires through the infrastructure of the building. This type of device resonator to traditional outlet junction box may be used in a plurality of applications, including being designed for the interior or exterior of a building, to be made portable, made for a vehicle, and the like. Wireless power may be transferred through common building materials, such as wood, wall board, insulation, glass. brick, stone, concrete, and the like. The benefits of reduced installation cost, re-configurability, and increased application flexibility may provide the user significant benefits over traditional wired installations. The device resonator for a traditional outlet junction box may include a plurality of electrical components for facilitating the transfer of power from the device resonator to the traditional outlets, such as power source electronics which convert the specific frequencies needed to implement efficient power transfer to line voltage, power capture electronics which may convert high frequency AC to usable voltage and frequencies (AC and/or DC), controls which synchronize the capture device and the power output and which ensure consistent, safe, and maximally efficient power transfer, and the like.

[00430] The systems and methods described herein may provide advantages to lights or electrical components that operate in environments that are wet, harsh, controlled, and the like, such has outside and exposed to the rain, in a pool/sauna/shower, in a maritime application, in hermetically sealed components, in an explosive-proof room, on outside signage, a harsh industrial environment in a volatile environment (e.g. from volatile vapors or airborne organics, such as in a grain silo or bakery) , and the like. For example, a light mounted under the water level of a pool is normally difficult to wire up, and is required to be water-sealed despite the need for external wires. But a pool light using the principles disclosed herein may more easily be made water sealed, as there may be no external wires needed. In another example, an explosion proof room, such as containing volatile vapors, may not only need to be hermetically sealed, but may need to have all electrical contacts (that could create a spark) sealed. Again, the principles disclosed herein may provide a convenient way to supply sealed electrical components for such applications.

[00431] The systems and methods disclosed herein may provide power to game controller applications, such as to a remote handheld game controller. These game controllers may have been traditionally powered solely by batteries, where the game controller's use and power profile caused frequent changing of the battery, battery pack, rechargeable batteries, and the like, that may not have been ideal for the consistent use to the game controller, such as during extended game play. A device resonator may be placed into the game controller, and a source resonator, connected to a power source, may be placed in the vicinity. Further, the device resonator in the game controller may provide power directly to the game controller electronics without a battery; provide power to a battery, battery pack, rechargeable battery, and the like, which then provides power to the game controller electronics; and the like. The game controller may utilize multiple battery packs, where each battery pack is equipped with a device resonator, and thus may be constantly recharging while in the vicinity of the source resonator, whether plugged into the game controller or not. The source resonator may be resident in a main game controller facility for the game, where the main game controller facility and source resonator are supplied power from AC 'house' power; resident in an extension facility form AC power, such as in a source resonator integrated into an 'extension cord'; resident in a game chair, which is at least one of plugged into the wall AC, plugged into the main game controller facility, powered by a battery pack in the game chair; and the like. The source resonator may be placed and implemented in any of the configurations described herein.

[00432] The systems and methods disclosed herein may integrate device resonators into battery packs, such as battery packs that are interchangeable with other battery packs. For instance, some portable devices may use up electrical energy at a high rate such that a user may need to have multiple interchangeable battery packs on hand for use, or the user may operate the device out of range of a source resonator and need additional battery packs to continue operation, such as for power tools, portable lights, remote control vehicles, and the like. The use of the principles disclosed herein may not only provide a way for device resonator enabled battery packs to be recharged while in use and in range, but also for the recharging of battery packs not currently in use and placed in range of a source resonator. In this way, battery packs may always be ready to use when a user runs down the charge of a battery pack being used. For example, a user may be working with a wireless power tool, where the current requirements may be greater than can be realized through direct powering from a source resonator. In this case, despite the fact that the systems and methods described herein may be providing charging power to the in-use battery pack while in range, the battery pack may still run down, as the power usage may have exceeded the recharge rate. Further, the user may simply be moving in and out of range, or be completely out of range while using the device. However, the user may have placed additional battery packs in the vicinity of the source resonator, which have been recharged while not in use, and are now charged sufficiently for use. In another example, the user may be working with the power tool away from the vicinity of the source resonator, but leave the supplemental battery packs to charge in the vicinity of the source resonator, such as in a room with a portable source resonator or extension cord source resonator, in the user's vehicle, in user's tool box, and the like. In this way, the user may not have to worry about taking the time to, and/or remembering to plug in their battery packs for future use. The user may only have to change out the used battery pack for the charged battery pack and place the used one in the vicinity of the source resonator for recharging. Device resonators may be built into enclosures with known battery form factors and footprints and may replace traditional chemical batteries in known devices and applications. For example, device resonators may be built into enclosures with mechanical dimensions equivalent to AA batteries, AAA batteries, D batteries, 9V batteries, laptop batteries, cell phone batteries, and the like. The enclosures may include a smaller "button battery" in addition to the device resonator to store charge and provide extended operation, either in terms of time or distance. Other energy storage devices in addition to or instead of button batteries may be integrated with the device resonators and any associated power conversion circuitry. These new energy packs may provide similar voltage and current levels as provided by traditional batteries, but may be composed of device resonators, power conversion electronics, a small battery, and the like. These new energy packs may last longer than traditional batteries because they may be more easily recharged and may be recharging constantly when they are located in a wireless power zone. In addition, such energy packs may be lighter than traditional batteries, may be safer to use and store, may operate over wider temperature and humidity ranges, may be less harmful to the environment when thrown away, and the like. As described herein, these energy packs may last beyond the life of the product when used in wireless power zones as described herein.

[00433] The systems and methods described herein may be used to power visual displays, such as in the case of the laptop screen, but more generally to include the great variety and diversity of displays utilized in today's electrical and electronics components, such as in televisions, computer monitors, desktop monitors, laptop displays, digital photo frames, electronic books, mobile device displays (e.g. on phones, PDAs, games, navigation devices, DVD players), and the like. Displays that may be powered through one or more of the wireless power transmission systems described herein may also include embedded displays, such as embedded in electronic components (e.g. audio equipment, home appliances, automotive displays, entertainment devices, cash registers, remote controls), in furniture, in building infrastructure, in a vehicle, on the surface of an object (e.g. on the surface of a vehicle, building, clothing, signs, transportation), and the like. Displays may be very small with tiny resonant devices, such as in a smart card as described herein, or very large, such as in an advertisement sign. Displays powered using the principles disclosed herein may also be any one of a plurality of imaging technologies, such as liquid crystal display (LCD), thin film transistor LCD, passive LCD, cathode ray tube (CRT), plasma display, projector display (e.g. LCD, DLP, LCOS), surface-conduction electron-emitter display (SED), organic light-emitting diode (OLED), and the like. Source coil configurations may include attaching to a primary power source, such as building power, vehicle power, from a wireless extension cord as described herein, and the like; attached to component power, such as the base of an electrical component (e.g. the base of a computer, a cable box for a TV); an intermediate relay source coil; and the like. For example, hanging a digital display on the wall may be very appealing, such as in the case of a digital photo frame that receives its information signals wirelessly or through a portable memory device, but the need for an unsightly power cord may make it aesthetically unpleasant. However, with a device coil embedded in the digital photo frame, such as wrapped within the frame portion, may allow the digital photo frame to be hung with no wires at all. The source resonator may then be placed in the vicinity of the digital photo frame, such as in the next room on the other side of the wall, plugged directly into a traditional power outlet, from a wireless extension cord as described herein, from a central source resonator for the room, and the like.

[00434] The systems and methods described herein may provide wireless power transmission between different portions of an electronics facility. Continuing with the example of the laptop computer, and without limitation, the screen of the laptop computer may require power from the base of the laptop. In this instance, the electrical power has been traditionally routed via direct electrical connection from the base of the laptop to the screen over a hinged portion of the laptop between the screen and the base. When a wired connection is utilized, the wired connection may tend to wear out and break, the design functionality of the laptop computer may be limited by the required direct electrical connection, the design aesthetics of the laptop computer may be limited by the required direct electrical connection, and the like. However, a wireless connection may be made between the base and the screen. In this instance, the device resonator may be placed in the screen portion to power the display, and the base may be either powered by a second device resonator, by traditional wired connections, by a hybrid of resonator-battery- direct electrical connection, and the like. This may not only improve the reliability of the power connection due to the removal of the physical wired connection, but may also allow designers to improve the functional and/or aesthetic design of the hinge portion of the laptop in light of the absence of physical wires associated with the hinge. Again, the laptop computer has been used here to illustrate how the principles disclosed herein may improve the design of an electric or electronic device, and should not be taken as limiting in any way. For instance, many other electrical devices with separated physical portions could benefit from the systems and methods described herein, such as a refrigerator with electrical functions on the door, including an ice maker, a sensor system, a light, and the like; a robot with movable portions, separated by joints; a car's power system and a component in the car's door; and the like. The ability to provide power to a device via a device resonator from an external source resonator, or to a portion of the device via a device resonator from either external or internal source resonators, will be recognized by someone skilled in the art to be widely applicable across the range of electric and electronic devices.

[00435] The systems and methods disclosed herein may provide for a sharing of electrical power between devices, such as between charged devices and uncharged devices. For instance a charged up device or appliance may act like a source and send a predetermined amount of energy, dialed in amount of energy, requested and approved amount of energy, and the like, to a nearby device or appliance. For example, a user may have a cell phone and a digital camera that are both capable of transmitting and receiving power through embedded source and device resonators, and one of the devices, say the cell phone, is found to be low on charge. The user may then transfer charge from the digital camera to the cell phone. The source and device resonators in these devices may utilize the same physical resonator for both transmission and reception, utilize separate source and device resonators, one device may be designed to receive and transmit while the other is designed to receive only, one device may be designed to transmit only and the other to receive only, and the like.

[00436] To prevent complete draining the battery of a device it may have a setting allowing a user to specify how much of the power resource the receiving device is entitled to. It may be useful, for example, to put a limit on the amount of power available to external devices and to have the ability to shut down power transmission when battery power falls below a threshold.

[00437] The systems and methods described herein may provide wireless power transfer to a nearby electrical or electronics component in association with an electrical facility, where the source resonator is in the electrical facility and the device resonator is in the electronics component. The source resonator may also be connected to, plugged into, attached to the electrical facility, such as through a universal interface (e.g. a USB interface, PC card interface), supplemental electrical outlet, universal attachment point, and the like, of the electrical facility. For example, the source resonator may be inside the structure of a computer on a desk, or be integrated into some object, pad, and the like, that is connected to the computer, such as into one of the computer's USB interfaces. In the example of the source resonator embedded in the object, pad, and the like, and powered through a USB interface, the source resonator may then be easily added to a user's desktop without the need for being integrated into any other electronics device, thus conveniently providing a wireless energy zone around which a plurality of electric and/or electronics devices may be powered. The electrical facility may be a computer, a light fixture, a dedicated source resonator electrical facility, and the like, and the nearby components may be computer peripherals, surrounding electronics components, infrastructure devices, and the like, such as computer keyboards, computer mouse, fax machine, printer, speaker system, cell phone, audio device, intercom, music player, PDA, lights, electric pencil sharpener, fan, digital picture frame, calculator, electronic games, and the like. For example, a computer system may be the electrical facility with an integrated source resonator that utilizes a 'wireless keyboard' and 'wireless mouse', where the use of the term wireless here is meant to indicate that there is wireless communication facility between each device and the computer, and where each device must still contain a separate battery power source. As a result, batteries would need to be replaced periodically, and in a large company, may result in a substantial burden for support personnel for replacement of batteries, cost of batteries, and proper disposal of batteries. Alternatively, the systems and methods described herein may provide wireless power transmission from the main body of the computer to each of these peripheral devices, including not only power to the keyboard and mouse, but to other peripheral components such as a fax, printer, speaker system, and the like, as described herein. A source resonator integrated into the electrical facility may provide wireless power transmission to a plurality of peripheral devices, user devices, and the like, such that there is a significant reduction in the need to charge and/or replace batteries for devices in the near vicinity of the source resonator integrated electrical facility. The electrical facility may also provide tuning or auto-tuning software, algorithms, facilities, and the like, for adjusting the power transfer parameters between the electrical facility and the wirelessly powered device. For example, the electrical facility may be a computer on a user's desktop, and the source resonator may be either integrated into the computer or plugged into the computer (e.g. through a USB connection), where the computer provides a facility for providing the tuning algorithm (e.g. through a software program running on the computer).

[00438] The systems and methods disclosed herein may provide wireless power transfer to a nearby electrical or electronics component in association with a facility infrastructure component, where the source resonator is in, or mounted on, the facility infrastructure component and the device resonator is in the electronics component. For instance, the facility infrastructure component may be a piece of furniture, a fixed wall, a movable wall or partition, the ceiling, the floor, and the source resonator attached or integrated into a table or desk (e.g. just below/above the surface, on the side, integrated into a table top or table leg), a mat placed on the floor (e.g. below a desk, placed on a desk), a mat on the garage floor (e.g. to charge the car and/or devices in the car), in a parking lot/garage (e.g. on a post near where the car is parked), a television (e.g. for charging a remote control), a computer monitor (e.g. to power/charge a wireless keyboard, wireless mouse, cell phone), a chair (e.g. for powering electric blankets, medical devices, personal health monitors), a painting, office furniture, common household appliances, and the like. For example, the facility infrastructure component may be a lighting fixture in an office cubical, where the source resonator and light within the lighting fixture are both directly connected to the facility's wired electrical power. However, with the source resonator now provided in the lighting fixture, there would be no need to have any additional wired connections for those nearby electrical or electronics components that are connected to, or integrated with, a device resonator. In addition, there may be a reduced need for the replacement of batteries for devices with device resonators, as described herein.

[00439] The use of the systems and methods described herein to supply power to electrical and electronic devices from a central location, such as from a source resonator in an electrical facility, from a facility infrastructure component and the like, may minimize the electrical wiring infrastructure of the surrounding work area. For example, in an enterprise office space there are typically a great number of electrical and electronic devices that need to be powered by wired connections. With utilization of the systems and methods described herein, much of this wiring may be eliminated, saving the enterprise the cost of installation, decreasing the physical limitations associated with office walls having electrical wiring, minimizing the need for power outlets and power strips, and the like. The systems and methods described herein may save money for the enterprise through a reduction in electrical infrastructure associated with installation, re-installation (e.g., reconfiguring office space), maintenance, and the like. In another example, the principles disclosed herein may allow the wireless placement of an electrical outlet in the middle of a room. Here, the source could be placed on the ceiling of a basement below the location on the floor above where one desires to put an outlet. The device resonator could be placed on the floor of the room right above it. Installing a new lighting fixture (or any other electric device for that matter, e.g. camera, sensor, etc., in the center of the ceiling may now be substantially easier for the same reason).

[00440] In another example, the systems and methods described herein may provide power "through" walls. For instance, suppose one has an electric outlet in one room (e.g. on a wall), but one would like to have an outlet in the next room, but without the need to call an electrician, or drill through a wall, or drag a wire around the wall, or the like. Then one might put a source resonator on the wall in one room, and a device resonator outlet/pickup on the other side of the wall. This may power a flat-screen TV or stereo system or the like (e.g. one may not want to have an ugly wire climbing up the wall in the living room, but doesn't mind having a similar wire going up the wall in the next room, e.g. storage room or closet, or a room with furniture that blocks view of wires running along the wall). The systems and methods described herein may be used to transfer power from an indoor source to various electric devices outside of homes or buildings without requiring holes to be drilled through, or conduits installed in, these outside walls. In this case, devices could be wirelessly powered outside the building without the aesthetic or structural damage or risks associated with drilling holes through walls and siding. In addition, the systems and methods described herein may provide for a placement sensor to assist in placing an interior source resonator for an exterior device resonator equipped electrical component. For example, a home owner may place a security light on the outside of their home which includes a wireless device resonator, and now needs to adequately or optimally position the source resonator inside the home. A placement sensor acting between the source and device resonators may better enable that placement by indicating when placement is good, or to a degree of good, such as in a visual indication, an audio indication, a display indication, and the like. In another example, and in a similar way, the systems and methods described herein may provide for the installation of equipment on the roof of a home or building, such as radio transmitters and receivers, solar panels and the like. In the case of the solar panel, the source resonator may be associated with the panel, and power may be wirelessly transferred to a distribution panel inside the building without the need for drilling through the roof. The systems and methods described herein may allow for the mounting of electric or electrical components across the walls of vehicles (such as through the roof) without the need to drill holes, such as for automobiles, water craft, planes, trains, and the like. In this way, the vehicle's walls may be left intact without holes being drilled, thus maintaining the value of the vehicle, maintaining watertightness, eliminating the need to route wires, and the like. For example, mounting a siren or light to the roof of a police car decreases the future resale of the car, but with the systems and methods described herein, any light, horn, siren, and the like, may be attached to the roof without the need to drill a hole.

[00441] The systems and methods described herein may be used for wireless transfer of power from solar photovoltaic (PV) panels. PV panels with wireless power transfer capability may have several benefits including simpler installation, more flexible, reliable, and weatherproof design. Wireless power transfer may be used to transfer power from the PV panels to a device, house, vehicle, and the like. Solar PV panels may have a wireless source resonator allowing the PV panel to directly power a device that is enabled to receive the wireless power. For example, a solar PV panel may be mounted directly onto the roof of a vehicle, building, and the like. The energy captured by the PV panel may be wirelessly transferred directly to devices inside the vehicle or under the roof of a building. Devices that have resonators can wirelessly receive power from the PV panel. Wireless power transfer from PV panels may be used to transfer energy to a resonator that is coupled to the wired electrical system of a house, vehicle, and the like allowing traditional power distribution and powering of conventional devices without requiring any direct contact between the exterior PV panels and the internal electrical system.

[00442] With wireless power transfer significantly simpler installation of rooftop PV panels is possible because power may be transmitted wirelessly from the panel to a capture resonator in the house, eliminating all outside wiring, connectors, and conduits, and any holes through the roof or walls of the structure. Wireless power transfer used with solar cells may have a benefit in that it can reduced roof danger since it eliminates the need for electricians to work on the roof to interconnect panels, strings, and junction boxes. Installation of solar panels integrated with wireless power transfer may require less skilled labor since fewer electrical contacts need to be made. Less site specific design may be required with wireless power transfer since the technology gives the installer the ability to individually optimize and position each solar PV panel, significantly reducing the need for expensive engineering and panel layout services. There may not be need to carefully balance the solar load on every panel and no need for specialized DC wiring layout and interconnections.

[00443] For rooftop or on-wall installations of PV panels, the capture resonator may be mounted on the underside of the roof, inside the wall, or in any other easily accessible inside space within a foot or two of the solar PV panel. A diagram showing a possible general rooftop PV panel installation is shown in Figure 51. Various PV solar collectors may be mounted in top of a roof with wireless power capture coils mounted inside the building under the roof. The resonator coils in the PV panels can transfer their energy wirelessly through the roof to the wireless capture coils. The captured energy from the PV cells may be collected and coupled to the electrical system of the house to power electric and electronic devices or coupled to the power grid when more power than needed is generated. Energy is captured from the PV cells without requiring holes or wires that penetrate the roof or the walls of the building. Each PV panel may have a resonator that is coupled to a corresponding resonator on the interior of the vehicle or building. Multiple panels may utilize wireless power transfer between each other to transfer or collect power to one or a couple of designated panels that are coupled to resonators on the interior of the vehicle of house. Panels may have wireless power resonators on their sides or in their perimeter that can couple to resonators located in other like panels allowing transfer of power from panel to panel. An additional bus or connection structure may be provided that wirelessly couples the power from multiple panels on the exterior of a building or vehicle and transfers power to one or a more resonators on the interior of building or vehicle.

[00444] For example, as shown in Fig. 51, a source resonator 5102 may be coupled to a PV cell 5100 mounted on top of roof 5104 of a building. A corresponding capture resonator 5106 is placed inside the building. The solar energy captured by the PV cells can then be transferred between the source resonators 5102 outside to the device resonators 5106 inside the building without having direct holes and connections through the building.

[00445] Each solar PV panel with wireless power transfer may have its own inverter, significantly improving the economics of these solar systems by individually optimizing the power production efficiency of each panel, supporting a mix of panel sizes and types in a single installation, including single panel "pay-as-you-grow" system expansions. Reduction of installation costs may make a single panel economical for installation. Eliminating the need for panel string designs and careful positioning and orienting of multiple panels, and eliminating a single point of failure for the system.

[00446] Wireless power transfer in PV solar panels may enable more solar deployment scenarios because the weather-sealed solar PV panels eliminate the need to drill holes for wiring through sealed surfaces such as car roofs and ship decks, and eliminate the requirement that the panels be installed in fixed locations. With wireless power transfer, PV panels may be deployed temporarily, and then moved or removed, without leaving behind permanent alterations to the surrounding structures. They may be placed out in a yard on sunny days, and moved around to follow the sun, or brought inside for cleaning or storage, for example. For backyard or mobile solar PV applications, an extension cord with a wireless energy capture device may be thrown on the ground or placed near the solar unit. The capture extension cord can be completely sealed from the elements and electrically isolated, so that it may be used in any indoor or outdoor environment.

[00447] With wireless power transfer no wires or external connections may be necessary and the PV solar panels can be completely weather sealed. Significantly improved reliability and lifetime of electrical components in the solar PV power generation and transmission circuitry can be expected since the weather-sealed enclosures can protect components from UV radiation, humidity, weather, and the like. With wireless power transfer and weather-sealed enclosures it may be possible to use less expensive components since they will no longer be directly exposed to external factors and weather elements and it may reduce the cost of PV panels.

[00448] Power transfer between the PV panels and the capture resonators inside a building or a vehicle may be bidirectional. Energy may be transmitted from the house grid to the PV panels to provide power when the panels do not have enough energy to perform certain tasks such. Reverse power flow can be used to melt snow from the panels, or power motors that will position the panels in a more favorable positions with respect to the sun energy. Once the snow is melted or the panels are repositioned and the PV panels can generate their own energy the direction of power transfer can be returned to normal delivering power from the PV panels to buildings, vehicles, or devices.

[00449] PV panels with wireless power transfer may include auto-tuning on installation to ensure maximum and efficient power transfer to the wireless collector. Variations in roofing materials or variations in distances between the PV panels and the wireless power collector in different installations may affect the performance or perturb the properties of the resonators of the wireless power transfer. To reduce the installation complexity the wireless power transfer components may include a tuning capability to automatically adjust their operating point to compensate for any effects due to materials or distance. Frequency, impedance, capacitance, inductance, duty cycle, voltage levels and the like may be adjusted to ensure efficient and safe power transfer

[00450] The systems and methods described herein may be used to provide a wireless power zone on a temporary basis or in extension of traditional electrical outlets to wireless power zones, such as through the use of a wireless power extension cord. For example, a wireless power extension cord may be configured as a plug for connecting into a traditional power outlet, a long wire such as in a traditional power extension cord, and a resonant source coil on the other end (e.g. in place of, or in addition to, the traditional socket end of the extension The wireless extension cord may also be configured where there are source resonators at a plurality of locations along the wireless extension cord. This configuration may then replace any traditional extension cord where there are wireless power configured devices, such as providing wireless power to a location where there is no convenient power outlet (e.g. a location in the living room where there's no outlet), for temporary wireless power where there is no wired power infrastructure (e.g. a construction site), out into the yard where there are no outlets (e.g. for parties or for yard grooming equipment that is wirelessly powered to decrease the chances of cutting the traditional electrical cord), and the like. The wireless extension cord may also be used as a drop within a wall or structure to provide wireless power zones within the vicinity of the drop. For example, a wireless extension cord could be run within a wall of a new or renovated room to provide wireless power zones without the need for the installation of traditional electrical wiring and outlets.

[00451] The systems and methods described herein may be utilized to provide power between moving parts or rotating assemblies of a vehicle, a robot, a mechanical device, a wind turbine, or any other type of rotating device or structure with moving parts such as robot arms, construction vehicles, movable platforms and the like. Traditionally, power in such systems may have been provided by slip rings or by rotary joints for example. Using wireless power transfer as described herein, the design simplicity, reliability and longevity of these devices may be significantly improved because power can be transferred over a range of distances without any physical connections or contact points that may wear down or out with time. In particular, the preferred coaxial and parallel alignment of the source and device coils may provide wireless power transmission that is not severely modulated by the relative rotational motion of the two coils.

[00452] The systems and methods described herein may be utilized to extend power needs beyond the reach of a single source resonator by providing a series of source-device-source-device resonators. For instance, suppose an existing detached garage has no electrical power and the owner now wants to install a new power service. However, the owner may not want to run wires all over the garage, or have to break into the walls to wire electrical outlets throughout the structure. In this instance, the owner may elect to connect a source resonator to the new power service, enabling wireless power to be supplied to device resonator outlets throughout the back of the garage. The owner may then install a device-source ‘relay’ to supply wireless power to device resonator outlets in the front of the garage. That is, the power relay may now receive wireless power from the primary source resonator, and then supply available power to a second source resonator to supply power to a second set of device resonators in the front of the garage. This configuration may be repeated again and again to extend the effective range of the supplied wireless power.

[00453] Multiple resonators may be used to extend power needs around an energy blocking material. For instance, it may be desirable to integrate a source resonator into a computer or computer monitor such that the resonator may power devices placed around and especially in front of the monitor or computer such as keyboards, computer mice, telephones , and the like. Due to aesthetics, space constraints, and the like an energy source that may be used for the source resonator may only be located or connected to in the back of the monitor or computer. In many designs of computer or monitors metal components and metal containing circuits are used in the design and packaging which may limit and prevent power transfer from source resonator in the back of the monitor or computer to the front of the monitor or computer. An additional repeater resonator may be integrated into the base or pedestal of the monitor or computer that couples to the source resonator in the back of the monitor or computer and allows power transfer to the space in front of the monitor or computer. The intermediate resonator integrated into the base or pedestal of the monitor or computer does not require an additional power source, it captures power from the source resonator and transfers power to the front around the blocking or power shielding metal components of the monitor or computer.

[00454] The systems and methods described herein may be built-into, placed on, hung from, embedded into, integrated into, and the like, the structural portions of a space, such as a vehicle, office, home, room, building, outdoor structure, road infrastructure, and the like. For instance, one or more sources may be built into, placed on, hung from, embedded or integrated into a wall, a ceiling or ceiling panel, a floor, a divider, a doorway, a stairwell, a compartment, a road surface, a sidewalk, a ramp, a fence, an exterior structure, and the like. One or more sources may be built into an entity within or around a structure, for instance a bed, a desk, a chair, a rug, a mirror, a clock, a display, a television, an electronic device, a counter, a table, a piece of furniture, a piece of artwork, an enclosure, a compartment, a ceiling panel, a floor or door panel, a dashboard, a trunk, a wheel well, a post, a beam, a support or any like entity. For example, a source resonator may be integrated into the dashboard of a user's car so that any device that is equipped with or connected to a device resonator may be supplied with power from the dashboard source resonator. In this way, devices brought into or integrated into the car may be constantly charged or powered while in the car.

[00455] The systems and methods described herein may provide power through the walls of vehicles, such as boats, cars, trucks, busses, trains, planes, satellites and the like. For instance, a user may not want to drill through the wall of the vehicle in order to provide power to an electric device on the outside of the vehicle. A source resonator may be placed inside the vehicle and a device resonator may be placed outside the vehicle (e.g. on the opposite side of a window, wall or structure). In this way the user may achieve greater flexibility in optimizing the placement, positioning and attachment of the external device to the vehicle, (such as without regard to supplying or routing electrical connections to the device). In addition, with the electrical power supplied wirelessly, the external device may be sealed such that it is water tight, making it safe if the electric device is exposed to weather (e.g. rain), or even submerged under water. Similar techniques may be employed in a variety of applications, such as in charging or powering hybrid vehicles, navigation and communications equipment, construction equipment, remote controlled or robotic equipment and the like, where electrical risks exist because of exposed conductors. The systems and methods described herein may provide power through the walls of vacuum chambers or other enclosed spaces such as those used in semiconductor growth and processing, material coating systems, aquariums, hazardous materials handling systems and the like. Power may be provided to translation stages, robotic arms, rotating stages, manipulation and collection devices, cleaning devices and the like.

[00456] The systems and methods described herein may provide wireless power to a kitchen environment, such as to counter-top appliances, including mixers, coffee makers, toasters, toaster ovens, grills, griddles, electric skillets, electric pots, electric woks, waffle makers, blenders, food processors, crock pots, warming trays, induction cooktops, lights, computers, displays, and the like. This technology may improve the mobility and/or positioning flexibility of devices, reduce the number of power cords stored on and strewn across the counter-top, improve the washability of the devices, and the like. For example, an electric skillet may traditionally have separate portions, such as one that is submersible for washing and one that is not submersible because it includes an external electrical connection (e.g. a cord or a socket for a removable cord). However, with a device resonator integrated into the unit, all electrical connections may be sealed, and so the entire device may now be submersed for cleaning. In addition, the absence of an external cord may eliminate the need for an available electrical wall outlet, and there is no longer a need for a power cord to be placed across the counter or for the location of the electric griddle to be limited to the location of an available electrical wall outlet.

[00457] The systems and methods described herein may provide continuous power/charging to devices equipped with a device resonator because the device doesn't leave the proximity of a source resonator, such as fixed electrical devices, personal computers, intercom systems, security systems, household robots, lighting, remote control units, televisions, cordless phones, and the like. For example, a household robot (e.g. ROOMBA) could be powered/charged via wireless power, and thus work arbitrarily long without recharging. In this way, the power supply design for the household robot may be changed to take advantage of this continuous source of wireless power, such as to design the robot to only use power from the source resonator without the need for batteries, use power from the source resonator to recharge the robot's batteries, use the power from the source resonator to trickle charge the robot's batteries, use the power from the source resonator to charge a capacitive energy storage unit, and the like. Similar optimizations of the power supplies and power circuits may be enabled, designed, and realized, for any and all of the devices disclosed herein.

[00458] The systems and methods described herein may be able to provide wireless power to electrically heated blankets, heating pads/patches, and the like. These electrically heated devices may find a variety of indoor and outdoor uses. For example, hand and foot warmers supplied to outdoor workers such as guards, policemen, construction workers and the like might be remotely powered from a source resonator associated with or built into a nearby vehicle, building, utility pole, traffic light, portable power unit, and the like.

[00459] The systems and methods described herein may be used to power a portable information device that contains a device resonator and that may be powered up when the information device is near an information source containing a source resonator. For instance, the information device may be a card (e.g. credit card, smart card, electronic card, and the like) carried in a user's pocket, wallet, purse, vehicle, bike, and the like. The portable information device may be powered up when it is in the vicinity of an information source that then transmits information to the portable information device that may contain electronic logic, electronic processors, memory, a display, an LCD display, LEDs, RFID tags, and the like. For example, the portable information device may be a credit card with a display that "turns on" when it is near an information source, and provide the user with some information such as, “You just received a coupon for 50% off your next Coca Cola purchase”. The information device may store information such as coupon or discount information that could be used on subsequent purchases. The portable information device may be programmed by the user to contain tasks, calendar appointments, to-do lists, alarms and reminders, and the like. The information device may receive up-to-date price information and inform the user of the location and price of previously selected or identified items.

[00460] The systems and methods described herein may provide wireless power transmission to directly power or recharge the batteries in sensors, such as environmental sensors, security sensors, agriculture sensors, appliance sensors, food spoilage sensors, power sensors, and the like, which may be mounted internal to a structure, external to a structure, buried underground, installed in walls, and the like. For example, this capability may replace the need to dig out old sensors to physically replace the battery, or to bury a new sensor because the old sensor is out of power and no longer operational. These sensors may be charged up periodically through the use of a portable sensor source resonator charging unit. For instance, a truck carrying a source resonator equipped power source, say providing ~kW of power, may provide enough power to a ~mW sensor in a few minutes to extend the duration of operation of the sensor for more than a year. Sensors may also be directly powered, such as powering sensors that are in places where it is difficult to connect to them with a wire but they are still within the vicinity of a source resonator, such as devices outside of a house (security camera), on the other side of a wall, on an electric lock on a door, and the like. In another example, sensors that may need to be otherwise supplied with a wired power connection may be powered through the systems and methods described herein. For example, a ground fault interrupter breaker combines residual current and over-current protection in one device for installation into a service panel. However, the sensor traditionally has to be independently wired for power, and this may complicate the installation. However, with the systems and methods described herein the sensor may be powered with a device resonator, where a single source resonator is provided within the service panel, thus simplifying the installation and wiring configuration within the service panel. In addition, the single source resonator may power device resonators mounted on either side of the source resonator mounted within the service panel, throughout the service panel, to additional nearby service panels, and the like. The systems and methods described herein may be employed to provide wireless power to any electrical component associated with electrical panels, electrical rooms, power distribution and the like, such as in electric switchboards, distribution boards, circuit breakers, transformers, backup batteries, fire alarm control panels, and the like. Through the use of the systems and methods described herein, it may be easier to install, maintain, and modify electrical distribution and protection components and system installations.

[00461] In another example, sensors that are powered by batteries may run continuously, without the need to change the batteries, because wireless power may be supplied to periodically or continuously recharge or trickle charge the battery. In such applications, even low levels of power may adequately recharge or maintain the charge in batteries, significantly extending their lifetime and usefulness. In some cases, the battery life may be extended to be longer than the lifetime of the device it is powering, making it essentially a battery that “lasts forever”.

[00462] The systems and methods described herein may be used for charging implanted medical device batteries, such as in an artificial heart, pacemaker, heart pump, insulin pump, implanted coils for nerve or acupressure/acupuncture point stimulation, and the like. For instance, it may not be convenient or safe to have wires sticking out of a patient because the wires may be a constant source of possible infection and may generally be very unpleasant for the patient. The systems and methods described herein may also be used to charge or power medical devices in or on a patient from an external source, such as from a bed or a hospital wall or ceiling with a source resonator. Such medical devices may be easier to attach, read, use and monitor the patient. The systems and methods described herein may ease the need for attaching wires to the patient and the patient’s bed or bedside, making it more convenient for the patient to move around and get up out of bed without the risk of inadvertently disconnecting a medical device. This may, for example, be usefully employed with patients that have multiple sensors monitoring them, such as for measuring pulse, blood pressure, glucose, and the like. For medical and monitoring devices that utilize batteries, the batteries may need to be replaced quite often, perhaps multiple times a week. This may present risks associated with people forgetting to replace batteries, not noticing that the devices or monitors are not working because the batteries have died, infection associated with improper cleaning of the battery covers and compartments, and the like.

[00463] The systems and methods described herein may reduce the risk and complexity of medical device implantation procedures. Today many implantable medical devices such as ventricular assist devices, pacemakers, defibrillators and the like, require surgical implantation due to their device form factor, which is heavily influenced by the volume and shape of the long-life battery that is integrated in the device. In one aspect, there is described herein a non-invasive method of recharging the batteries so that the battery size may be dramatically reduced, and the entire device may be implanted, such as via a catheter. A catheter implantable device may include an integrated capture or device coil. A catheter implantable capture or device coil may be designed so that it may be wired internally, such as after implantation. The capture or device coil may be deployed via a catheter as a rolled up flexible coil (e.g. rolled up like two scrolls, easily unrolled internally with a simple spreader mechanism). The power source coil may be worn in a vest or article of clothing that is tailored to fit in such a way that places the source in proper position, may be placed in a chair cushion or bed cushion, may be integrated into a bed or piece of furniture, and the like.

[00464] The systems and methods described herein may enable patients to have a 'sensor vest', sensor patch, and the like, that may include at least one of a plurality of medical sensors and a device resonator that may be powered or charged when it is in the vicinity of a source resonator. Traditionally, this type of medical monitoring facility may have required batteries, thus making the vest, patch, and the like, heavy, and potentially impractical. But using the principles disclosed herein, no batteries (or a lighter rechargeable battery) may be required, thus making such a device more convenient and practical, especially in the case where such a medical device could be held in place without straps, such as by adhesive, in the absence of batteries or with substantially lighter batteries. A medical facility may be able to read the sensor data remotely with the aim of anticipating (e.g. a few minutes ahead of) a stroke, a heart-attack, or the like. When the vest is used by a person in a location remote from the medical facility, such as in their home, the vest may then be integrated with a cell-phone or communications device to call an ambulance in case of an accident or a medical event. The systems and methods described herein may be of particular value in the instance when the vest is to be used by an elderly person, where traditional non-wireless recharging practices (e.g. replacing batteries, plugging in at night, and the like) may not be followed as required. The systems and methods described herein may also be used for charging devices that are used by or that aid handicapped or disabled people who may have difficulty replacing or recharging batteries, or reliably supplying power to devices they enjoy or rely on.

[00465] The systems and methods described herein may be used for the charging and powering of artificial limbs. Artificial limbs have become very capable in terms of replacing the functionality of original limbs, such as arms, legs, hands and feet. However, an electrically powered artificial limb may require substantial power, (such as 10-20W) which may translate into a substantial battery. In that case, the amputee may be left with a choice between a light battery that doesn't last very long, and a heavy battery that lasts much longer, but is more difficult to 'carry' around. The systems and methods described herein may enable the artificial limb to be powered with a device resonator, where the source resonator is either carried by the user and attached to a part of the body that may more easily support the weight (such as on a belt around the waist, for example) or located in an external location where the user will spend an adequate amount of time to keep the device charged or powered, such as at their desk, in their car, in their bed, and the like.

[00466] The systems and methods described herein may be used for charging and powering of electrically powered exo-skeletons, such as those used in industrial and military applications, and for elderly/weak/sick people. An electrically powered exo-skeleton may provide up to a 10-to-20 times increase in "strength" to a person, enabling the person to perform physically strenuous tasks repeatedly without much fatigue. However, exo-skeletons may require more than 100W of power under certain use scenarios, so battery powered operation may be limited to 30 minutes or less. The delivery of wireless power as described herein may provide a user of an exo-skeleton with a continuous supply of power both for powering the structural movements of the exo-skeleton and for powering various monitors and sensors distributed throughout the structure. For instance, an exo-skeleton with an embedded device resonator(s) may be supplied with power from a local source resonator. For an industrial exo-skeleton, the source resonator may be placed in the walls of the facility. For a military exo-skeleton, the source resonator may be carried by an armored vehicle. For an exo-skeleton employed to assist a caretaker of the elderly, the source resonator(s) may be installed or placed in or the room(s) of a person's home.

[00467] The systems and methods described herein may be used for the powering/charging of portable medical equipment, such as oxygen systems, ventilators, defibrillators, medication pumps, monitors, and equipment in ambulances or mobile medical units, and the like. Being able to transport a patient from an accident scene to the hospital, or to move patients in their beds to other rooms or areas, and bring all the equipment that is attached with them and have it powered the whole time offers great benefits to the patients' health and eventual well-being. Certainly one can understand the risks and problems caused by medical devices that stop working because their battery dies or because they must be unplugged while a patient is transported or moved in any way. For example, an emergency medical team on the scene of an automotive accident might need to utilize portable medical equipment in the emergency care of patients in the field. Such portable medical equipment must be properly maintained so that there is sufficient battery life to power the equipment for the duration of the emergency. However, it is too often the case that the equipment is not properly maintained so that batteries are not fully charged and in some cases, necessary equipment is not available to the first responders. The systems and methods described herein may provide for wireless power to portable medical equipment (and associated sensor inputs on the patient) in such a way that the charging and maintaining of batteries and power packs is provided automatically and without human intervention. Such a system also benefits from the improved mobility of a patient unencumbered by a variety of power cords attached to the many medical monitors and devices used in their treatment.

[00468] The systems and methods described herein may be used to for the powering/charging of personal hearing aids. Personal hearing aids need to be small and light to fit into or around the ear of a person. The size and weight restrictions limit the size of batteries that can be used. Likewise, the size and weight restrictions of the device make battery replacement difficult due to the delicacy of the components. The dimensions of the devices and hygiene concerns make it difficult to integrate additional charging ports to allow recharging of the batteries. The systems and methods described herein may be integrated into the hearing aid and may reduce the size of the necessary batteries which may allow even smaller hearing aids. Using the principles disclosed herein, the batteries of the hearing aid may be recharged without requiring external connections or charging ports. Charging and device circuitry and a small rechargeable battery may be integrated into a form factor of a conventional hearing aid battery allowing retrofit into existing hearing aids. The hearing aid may be recharged while it is used and worn by a person. The energy source may be integrated into a pad or a cup allowing recharging when the hearing is placed on such a structure. The charging source may be integrated into a hearing aid dryer box allowing wireless recharging while the hearing aid is drying or being sterilized. The source and device resonator may be used to also heat the device reducing or eliminating the need for an additional heating element. Portable charging cases powered by batteries or AC adaptors may be used as storage and charging stations.

[00469] The source resonator for the medical systems described above may be in the main body of some or all of the medical equipment, with device resonators on the patient's sensors and devices; the source resonator may be in the ambulance with device resonators on the patient's sensors and the main body of some or all of the equipment; a primary source resonator may be in the ambulance for transferring wireless power to a device resonator on the medical equipment while the medical equipment is in the ambulance and a second source resonator is in the main body of the medical equipment and a second device resonator on the patient sensors when the equipment is away from the ambulance; and the like. The systems and methods described herein may significantly improve the ease with which medical personnel are able to transport patients from one location to another, where power wires and the need to replace or manually charge associated batteries may now be reduced.

[00470] The systems and methods described herein may be used for the charging of devices inside a military vehicle or facility, such as a tank, armored carrier, mobile shelter, and the like. For instance, when soldiers come back into a vehicle after "action" or a mission, they may typically start charging their electronic devices. If their electronic devices were equipped with device resonators, and there was a source resonator inside the vehicle, (e.g. integrated in the seats or on the ceiling of the vehicle), their devices would start charging immediately. In fact, the same vehicle could provide power to soldiers/robots (e.g. packbot from iRobot) standing outside or walking beside the vehicle. This capability may be useful in minimizing accidental battery-swapping with someone else (this may be a significant issue, as soldiers tend to trust only their own batteries); in enabling quicker exits from a vehicle under attack; in powering or charging laptops or other electronic devices inside a tank, as too many wires inside the tank may present a hazard in terms of reduced ability to move around fast in case of "trouble" and/or decreased visibility; and the like. The systems and methods described herein may provide a significant improvement in association with powering portable power equipment in a military environment.

[00471] The systems and methods described herein may provide wireless powering or charging capabilities to mobile vehicles such as golf carts or other types of carts, all-terrain vehicles, electric bikes, scooters, cars, mowers, bobcats and other vehicles typically used for construction and landscaping and the like. The systems and methods described herein may provide wireless powering or charging capabilities to miniature mobile vehicles, such as mini-helicopters, airborne drones, remote control planes, remote control boats, remote controlled or robotic rovers, remote controlled or robotic lawn mowers or equipment, bomb detection robots, and the like. For instance, mini-helicopter flying above a military vehicle to increase its field of view can fly for a few minutes on standard batteries. If these mini-helicopters were fitted with a device resonator, and the control vehicle had a source resonator, the mini-helicopter might be able to fly indefinitely. The systems and methods described herein may provide an effective alternative to recharging or replacing the batteries for use in miniature mobile vehicles. In addition, the systems and methods described herein may provide power/charging to even smaller devices, such as microelectromechanical systems (MEMS), nano-robots, nano devices, and the like. In addition, the systems and methods described herein may be implemented by installing a source device in a mobile vehicle or flying device to enable it to serve as an in-field or in-flight re-charger, that may position itself autonomously in proximity to a mobile vehicle that is equipped with a device resonator.

[00472] The systems and methods described herein may be used to provide power networks for temporary facilities, such as military camps, oil drilling setups, remote filming locations, and the like, where electrical power is required, such as for power generators, and where power cables are typically run around the temporary facility. There are many instances when it is necessary to set up temporary facilities that require power. The systems and methods described herein may enable a more efficient way to rapidly set up and tear down these facilities, and may reduce the number of wires that must be run throughout the faculties to supply power. For instance, when Special Forces move into an area, they may erect tents and drag many wires around the camp to provide the required electricity. Instead, the systems and methods described herein may enable an army vehicle, outfitted with a power supply and a source resonator, to park in the center of the camp, and provide all the power to nearby tents where the device resonator may be integrated into the tents, or some other piece of equipment associated with each tent or area. A series of source-device-source-device resonators may be used to extend the power to tents that are farther away. That is, the tents closest to the vehicle could then provide power to tents behind them. The systems and methods described herein may provide a significant improvement to the efficiency with which temporary installations may be set up and torn down, thus improving the mobility of the associated facility.

[00473] The systems and methods described herein may be used in vehicles, such as for replacing wires, installing new equipment, powering devices brought into the vehicle, charging the battery of a vehicle (e.g. for a traditional gas powered engine, for a hybrid car, for an electric car, and the like), powering devices mounted to the interior or exterior of the vehicle, powering devices in the vicinity of the vehicle, and the like. For example, the systems and methods described herein may be used to replace wires such as those are used to power lights, fans and sensors distributed throughout a vehicle. As an example, a typical car may have 50kg of wires associated with it, and the use of the systems and methods described herein may enable the elimination of a substantial amount of this wiring. The performance of larger and more weight sensitive vehicles such as airplanes or satellites could benefit greatly from having the number of cables that must be run throughout the vehicle reduced. The systems and methods described herein may allow the accommodation of removable or supplemental portions of a vehicle with electric and electrical devices without the need for electrical harnessing. For example, a motorcycle may have removable side boxes that act as a temporary trunk space for when the cyclist is going on a long trip. These side boxes may have exterior lights, interior lights, sensors, auto equipment, and the like, and if not for being equipped with the systems and methods described herein might require electrical connections and harnessing.

[00474] An in-vehicle wireless power transmission system may charge or power one or more mobile devices used in a car: mobile phone handset, Bluetooth headset, blue tooth hands free speaker phone, GPS, MP3 player, wireless audio transceiver for streaming MP3 audio through car stereo via FM, Bluetooth, and the like. The in vehicle wireless power source may utilize source resonators that are arranged in any of several possible configurations including charging pad on dash, charging pad otherwise mounted on floor, or between seat and center console, charging "cup" or receptacle that fits in cup holder or on dash, and the like.

[00475] The wireless power transmission source may utilize a rechargeable battery system such that said supply battery gets charged whenever the vehicle power is on such that when the vehicle is turned off the wireless supply can draw power from the supply battery and can continue to wirelessly charge or power mobile devices that are still in the car.

[00476] The plug-in electric cars, hybrid cars, and the like, of the future need to be charged, and the user may need to plug in to an electrical supply when they get home or to a charging station. Based on a single over-night recharging, the user may be able to drive up to 50 miles the next day. Therefore, in the instance of a hybrid car, if a person drives less than 50 miles on most days, they will be driving mostly on electricity. However, it would be beneficial if they didn't have to remember to plug in the car at night. That is, it would be nice to simply drive into a garage, and have the car take care of its own charging. To this end, a source resonator may be built into the garage floor and/or garage side-wall, and the device resonator may be built into the bottom (or side) of the car. Even a few kW transfer may be sufficient to recharge the car over-night. The in-vehicle device resonator may measure magnetic field properties to provide feedback to assist in vehicle (or any similar device) alignment to a stationary resonating source. The vehicle may use this positional feedback to automatically position itself to achieve optimum alignment, thus optimum power transmission efficiency. Another method may be to use the positional feedback to help the human operator to properly position the vehicle or device, such as by making LED's light up, providing noises, and the like when it is well positioned. In such cases where the amount of power being transmitted could present a safety hazard to a person or animal that intrudes into the active field volume, the source or receiver device may be equipped with an active light curtain or some other external device capable of sensing intrusion into the active field volume, and capable of shutting off the source device and alert a human operator. In addition, the source device may be equipped with self-sensing capability such that it may detect that its expected power transmission rate has been interrupted by an intruding element, and in such case shut off the source device and alert a human operator. Physical or mechanical structures such as hinged doors or inflatable bladder shields may be incorporated as a physical barrier to prevent unwanted intrusions. Sensors such as optical, magnetic, capacitive, inductive, and the like may also be used to detect foreign structures or interference between the source and device resonators. The shape of the source resonator may be shaped such to prevent water or debris accumulation. The source resonator may be placed in a cone shaped enclosure or may have an enclosure with an angled top to allow water and debris to roll off. The source of the system may use battery power of the vehicle or its own battery power to transmit its presence to the source to initiate power transmission.

[00477] The source resonator may be mounted on an embedded or hanging post, on a wall, on a stand, and the like for coupling to a device resonator mounted on the bumper, hood, body panel, and the like, of an electric vehicle. The source resonator may be enclosed or embedded into a flexible enclosure such as a pillow, a pad, a bellows, a spring loaded enclosure and the like so that the electric vehicle may make contact with the structure containing the source coil without damaging the car in any way. The structure containing the source may prevent objects from getting between the source and device resonators. Because the wireless power transfer may be relatively insensitive to misalignments between the source and device coils, a variety of flexible source structures and parking procedures may be appropriate for this application.

[00478] The systems and methods described herein may be used to trickle charge batteries of electric, hybrid or combustion engine vehicles. Vehicles may require small amounts of power to maintain or replenish battery power. The power may be transferred wirelessly from a source to a device resonator that may be incorporated into the front grill, roof, bottom, or other parts of the vehicle. The device resonator may be designed to fit into a shape of a logo on the front of a vehicle or around the grill as not to obstruct air flow through the radiator. The device or source resonator may have additional modes of operation that allow the resonator to be used as a heating element which can be used to melt of snow or ice from the vehicle.

[00479] An electric vehicle or hybrid vehicle may require multiple device resonators, such as to increase the ease with which the vehicle may come in proximity with a source resonator for charging (i.e. the greater the number and varied position of device resonators are, the greater the chances that the vehicle can pull in and interface with a diversity of charging stations), to increase the amount of power that can be delivered in a period of time (e.g. additional device resonators may be required to keep the local heating due to charging currents to acceptable levels), to aid in automatic parking/docking the vehicle with the charging station, and the like. For example, the vehicle may have multiple resonators (or a single resonator) with a feedback system that provides guidance to either the driver or an automated parking/docking facility in the parking of the vehicle for optimized charging conditions (i.e., the optimum positioning of the vehicle's device resonator to the charging station's source resonator may provide greater power transfer efficiency). An automated parking/docking facility may allow for the automatic parking of the vehicle based on how well the vehicle is coupled.

[00480] The power transmission system may be used to power devices and peripherals of a vehicle. Power to peripherals may be provided while a vehicle is charging, or while not charging, or power may be delivered to conventional vehicles that do not need charging. For example, power may be transferred wirelessly to conventional non-electric cars to power air conditioning, refrigeration units, heaters, lights, and the like while parked to avoid running the engine which may be important to avoid exhaust build up in garage parking lots or loading docks. Power may for example be wirelessly transferred to a bus while it is parked to allow powering of lights, peripherals, passenger devices, and the like avoiding the use of onboard engines or power sources. Power may be wirelessly transferred to an airplane while parked on the tarmac or in a hanger to power instrumentation, climate control, de-icing equipment, and the like without having to use onboard engines or power sources.

[00481] Wireless power transmission on vehicles may be used to enable the concept of Vehicle to Grid (V2G). Vehicle to grid is based on utilizing electric vehicles and plug-in hybrid electric vehicles (PHEV) as distributed energy storage devices, charged at night when the electric grid is underutilized, and available to discharge back into the grid during episodes of peak demand that occur during the day. The wireless power transmission system on a vehicle and the respective infrastructure may be implemented in such a way as to enable bidirectional energy flow—so that energy can flow back into the grid from the vehicle—without requiring a plug in connection. Vast fleets of vehicles, parked at factories, offices, parking lots, can be viewed as "peaking power capacity" by the smart grid. Wireless power transmission on vehicles can make such a V2G vision a reality. By simplifying the process of connecting a vehicle to the grid, (i.e. by simply parking it in a wireless charging enabled parking spot), it becomes much more likely that a certain number of vehicles will be "dispatchable" when the grid needs to tap their power. Without wireless charging, electric and PHEV owners will likely charge their vehicles at home, and park them at work in conventional parking spots. Who will want to plug their vehicle in at work, if they do not need charging? With wireless charging systems capable of handling 3 kW, 100,000 vehicles can provide 300 Megawatts back to the grid—using energy generated the night before by cost effective base load generating capacity. It is the streamlined ergonomics of the cordless self charging PHEV and electric vehicles that make it a viable V2G energy source.

[00482] The systems and methods described herein may be used to power sensors on the vehicle, such as sensors in tires to measure air-pressure, or to run peripheral devices in the vehicle, such as cell phones, GPS devices, navigation devices, game players, audio or video players, DVD players, wireless routers, communications equipment, anti-theft devices, radar devices, and the like. For example, source resonators described herein may be built into the main compartment of the car in order to supply power to a variety of devices located both inside and outside of the main compartment of the car. Where the vehicle is a motorcycle or the like, devices described herein may be integrated into the body of the motorcycle, such as under the seat, and device resonators may be provided in a user's helmet, such as for communications, entertainment, signaling, and the like, or device resonators may be provided in the user's jacket, such as for displaying signals to other drivers for safety, and the like.

[00483] The systems and methods described herein may be used in conjunction with transportation infrastructure, such as roads, trains, planes, shipping, and the like. For example, source resonators may be built into roads, parking lots, rail-lines, and the like. Source resonators may be built into traffic lights, signs, and the like. For example, with source resonators embedded into a road, and device resonators built into vehicles, the vehicles may be provided power as they drive along the road or as they are parked in lots or on the side of the road. The systems and methods described herein may provide an effective way for electrical systems in vehicles to be powered and/or charged while the vehicle traverses a road network, or a portion of a road network. In this way, the systems and methods described herein may contribute to the powering/charging of autonomous vehicles, automatic guided vehicles, and the like. The systems and methods described herein may provide power to vehicles in places where they typically idle or stop, such as in the vicinity of traffic lights or signs, on highway ramps, or in parking lots.

[00484] The systems and methods described herein may be used in an industrial environment, such as inside a factory for powering machinery, powering/charging robots, powering and/or charging wireless sensors on robot arms, powering/charging tools and the like. For example, using the systems and methods described herein to supply power to devices on the arms of robots may help eliminate direct wire connections across the joints of the robot arm. In this way, the wearing out of such direct wire connections may be reduced, and the reliability of the robot increased. In this case, the device resonator may be out on the arm of the robot, and the source resonator may be at the base of the robot, in a central location near the robot, integrated into the industrial facility in which the robot is providing service, and the like. The use of the systems and methods described herein may help eliminate wiring otherwise associated with power distribution within the industrial facility, and thus benefit the overall reliability of the facility.

[00485] The systems and methods described herein may be used for underground applications, such as drilling, mining, digging, and the like. For example, electrical components and sensors associated with drilling or excavation may utilize the systems and methods described herein to eliminate cabling associated with a digging mechanism, a drilling bit, and the like, thus eliminating or minimizing cabling near the excavation point. In another example, the systems and methods described herein may be used to provide power to excavation equipment in a mining application where the power requirements for the equipment may be high and the distances large, but where there are no people to be subjected to the associated required fields. For instance, the excavation area may have device resonator powered digging equipment that has high power requirements and may be digging relatively far from the source resonator. As a result the source resonator may need to provide high field intensities to satisfy these requirements, but personnel are far enough away to be outside these high intensity fields. This high power, no personnel, scenario may be applicable to a plurality of industrial applications.

[00486] The systems and methods described herein may also use the near-field non-radiative resonant scheme for information transfer rather than, or in addition to, power transfer. For instance, information being transferred by near-field non-radiative resonance techniques may not be susceptible to eavesdropping and so may provide an increased level of security compared to traditional wireless communication schemes. In addition, information being transferred by near-field non-radiative resonance techniques may not interfere with the EM radiative spectrum and so may not be a source of EM interference, thereby allowing communications in an extended frequency range and well within the limits set by any regulatory bodies. Communication services may be provided between remote, inaccessible or hard-to-reach places such as between remote sensors, between sections of a device or vehicle, in tunnels, caves and wells (e.g. oil wells, other drill sites) and between underwater or underground devices, and the like. Communications services may be provided in places where magnetic fields experience less loss than electric fields.

[00487] The systems and methods described herein may enable the simultaneous transmission of power and communication signals between sources and devices in wireless power transmission systems, or it may enable the transmission of power and communication signals during different time periods or at different frequencies. The performance characteristics of the resonator may be controllably varied to preferentially support or limit the efficiency or range of either energy or information transfer. The performance characteristics of the resonators may be controlled to improve the security by reducing the range of information transfer, for example. The performance characteristics of the resonators may be varied continuously, periodically, or according to a predetermined, computed or automatically adjusted algorithm. For example, the power and information transfer enabled by the systems and methods described herein may be provided in a time multiplexed or frequency multiplexed manner. A source and device may signal each other by tuning, changing, varying, dithering, and the like, the resonator impedance which may affect the reflected impedance of other resonators that can be detected. The information transferred as described herein may include information regarding device identification, device power requirements, handshaking protocols, and the like.

[00488] The source and device may sense, transmit, process and utilize position and location information on any other sources and/or devices in a power network. The source and device may capture or use information such as elevation, tilt, latitude and longitude, and the like from a variety of sensors and sources that may be built into the source and device or may be part of a component the source or device connect. The positioning and orientation information may include sources such as global positioning sensors (GPS), compasses, accelerometers, pressure sensors, atmospheric barometric sensors, positioning systems which use Wi-Fi or cellular network signals, and the like. The source and device may use the position and location information to find nearby wireless power transmission sources. A source may broadcast or communicate with a central station or database identifying its location. A device may obtain the source location information from the central station or database or from the local broadcast and guide a user or an operator to the source with the aid of visual, vibrational, or auditory signals. Sources and devices may be nodes in a power network, in a communications network, in a sensor network, in a navigational network, and the like or in kind of combined functionality network.

[00489] The position and location information may also be used to optimize or coordinate power delivery. Additional information about the relative position of a source and a device may be used to optimize magnetic field direction and resonator alignment. The orientation of a device and a source which may be obtained from accelerometers and magnetic sensors, and the like, for example, may be used to identify the orientation of resonators and the most favorable direction of a magnetic field such that the magnetic flux is not blocked by the device circuitry. With such information a source with the most favorable orientation, or a combination of sources, may be used. Likewise, position and orientation information may be used to move or provide feedback to a user or operator of a device to place a device in a favorable orientation or location to maximize power transmission efficiency, minimize losses, and the like.

[00490] The source and device may include power metering and measuring circuitry and capability. The power metering may be used to track how much power was delivered to a device or how much power was transferred by a source. The power metering and power usage information may be used in fee based power delivery arrangements for billing purposes. Power metering may be also be used to enable power delivery policies to ensure power is distributed to multiple devices according to specific criteria. For example, the power metering may be used to categorize devices based on the amount of power they received and priority in power delivery may be given to those having received the least power. Power metering may be used to provide tiered delivery services such as “guaranteed power” and “best effort power” which may be billed at separate rates. Power metering may be used to institute and enforce hierarchical power delivery structures and may enable priority devices to demand and receive- more power under certain circumstances or use scenarios.

[00491] Power metering may be used to optimize power delivery efficiency and minimize absorption and radiation losses. Information related to the power received by devices may be used by a source in conjunction with information about the power output of the source to identify unfavorable operating environments or frequencies. For example, a source may compare the amount of power which was received by the devices and the amount of power which it transmitted to determine if the transmission losses may be unusually or unacceptably large. Large transmission losses may be due to an unauthorized device receiving power from the source and the source and other devices may initiate frequency hopping of the resonance frequency or other defensive measures to prevent or deter unauthorized use. Large transmission losses may be due to absorption losses for example, and the device and source may tune to alternate resonance frequencies to minimize such losses. Large transmission losses may also indicate the presence of unwanted or unknown objects or materials and the source may turn down or off its power level until the unwanted or unknown object is removed or identified, at which point the source may resume powering remote devices.

[00492] The source and device may include authentication capability. Authentication may be used to ensure that only compatible sources and devices are able to transmit and receive power. Authentication may be used to ensure that only authentic devices that are of a specific manufacturer and not clones or devices and sources from other manufacturers, or only devices that are part of a specific subscription or plan, are able to receive power from a source. Authentication may be based on cryptographic request and respond protocols or it may be based on the unique signatures of perturbations of specific devices allowing them to be used and authenticated based on properties similar to physically unclonable functions. Authentication may be performed locally between each source and device with local communication or it may be used with third person authentication methods where the source and device authenticate with communications to a central authority. Authentication protocols may use position information to alert a local source or sources of a genuine device.

[00493] The source and device may use frequency hopping techniques to prevent unauthorized use of a wireless power source. The source may continuously adjust or change the resonant frequency of power delivery. The changes in frequency may be performed in a pseudorandom or predetermined manner that is known, reproducible, or communicated to authorized device but difficult to predict. The rate of frequency hopping and the number of various frequencies used may be large and frequent enough to ensure that unauthorized use is difficult or impractical. Frequency hopping may be implemented by tuning the impedance network, tuning any of the driving circuits, using a plurality of resonators tuned or tunable to multiple resonant frequencies, and the like.

[00494] The source may have a user notification capability to show the status of the source as to whether it is coupled to a device resonator and transmitting power, if it is in standby mode, or if the source resonator is detuned or perturbed by an external object. The notification capability may include visual, auditory, and vibrational methods. The notification may be as simple as three color lights, one for each state, and optionally a speaker to provide notification in case of an error in operation. Alternatively, the notification capability may involve an interactive display that shows the status of the source and optionally provides instructions on how to fix or solve any errors or problems identified.

[00495] As another example, wireless power transfer may be used to improve the safety of electronic explosive detonators. Explosive devices are detonated with an electronic detonator, electric detonator, or shock tube detonator. The electronic detonator utilizes stored electrical energy (usually in a capacitor) to activate the igniter charge, with a low energy trigger signal transmitted conductively or by radio. The electric detonator utilizes a high energy conductive trigger signal to provide both the signal and the energy required to activate the igniter charge. A shock tube sends a controlled explosion through a hollow tube coated with explosive from the generator to the igniter charge. There are safety issues associated with the electric and electronic detonators, as there are cases of stray electromagnetic energy causing unintended activation. Wireless power transfer via sharply resonant magnetic coupling can improve the safety of such systems.

[00496] Using the wireless power transfer methods disclosed herein, one can build an electronic detonation system that has no locally stored energy, thus reducing the risk of unintended activation. A wireless power source can be placed in proximity (within a few meters) of the detonator. The detonator can be equipped with a resonant capture coil. The activation energy can be transferred when the wireless power source has been triggered. The triggering of the wireless power source can be initiated by any number of mechanisms: radio, magnetic near field radio, conductive signaling, ultrasonics, laser light. Wireless power transfer based on resonant magnetic coupling also has the benefit of being able to transfer power through materials such as rock, soil, concrete, water, and other dense materials. The use of very high Q coils as receivers and sources, having very narrow band response and sharply tuned to proprietary frequencies, further ensure that the detonator circuits cannot capture stray EMI and activate unintentionally.

[00497] The resonator of a wirelessly powered device may be external, or outside of the device, and wired to the battery of the device. The battery of the device may be modified to include appropriate rectification and control circuitry to receive the alternating currents of the device resonator. This can enable configurations with larger external coils, such as might be built into a battery door of a keyboard or mouse, or digital still camera, or even larger coils that are attached to the device but wired back to the battery/converter with ribbon cable. The battery door can be modified to provide interconnection from the external coil to the battery/converter (which will need an exposed contact that can touch the battery door contacts.

[00498] While the invention has been described in connection with certain preferred embodiments, other embodiments will be understood by one of ordinary skill in the art and are intended to fall within the scope of this disclosure, which is to be interpreted in the broadest sense allowable by law.

[00499] All documents referenced herein are hereby incorporated by reference.

ABSTRACT

[00500] In embodiments of the present invention improved capabilities are described for a method and system comprising a source resonator optionally coupled to an energy source and a second resonator located a distance from the source resonator, where the source resonator and the second resonator are coupled to provide near-field wireless energy transfer among the source resonator and the second resonator and where the field of at least one of the source resonator and the second resonator is shaped to avoid a loss-inducing object.

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PATENTS
WTCY-0026-P07

In re application of Aristeidis Karalis *et al.* : Group Art Unit: 2828

Serial No. 12/647,763 : Confirmation No. 2576

Filed: December 28, 2009 : Examiner: Not yet known

COMMUNICATION REGARDING A FILING RECEIPT CORRECTION

Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

Sir/Madam:

For the above indentified application, the applicant request a correction to an applicant name listed on the Filing Receipt mailed January 14, 2010. More specifically, the application request a change from Konrad Kulikowski to Konrad J. Kulikowski as shown on the marked Filing Receipt provided herewith. A supplemental application data sheet is provided herewith that includes the correct applicant information.

The applicant respectfully request an updated Filing Receipt which reflects the correct applicant name information.

Additionally, the applicant hereby authorizes the Patent Office to charge any deficiencies or credit any overpayments associated with this filing to Deposit Account No. 50-4262.

Please direct any inquiry regarding this matter to the below signed agent.

Respectfully Submitted,

STRATEGIC PATENTS, P.C.

/John Nortrup/

John H. Nortrup
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March 15, 2010

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APPLICATION NUMBER	FILING or 371(c) DATE	GRP ART UNIT	FIL FEE REC'D	ATTY.DOCKET.NO	TOT CLAIMS	IND CLAIMS
12/647,763	12/28/2009	2828	0.00	WTCY-0026-P07	14	2

CONFIRMATION NO. 2576

43520
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FILING RECEIPT

OC000000039637579

Date Mailed: 01/14/2010

Receipt is acknowledged of this non-provisional patent application. The application will be taken up for examination in due course. Applicant will be notified as to the results of the examination. Any correspondence concerning the application must include the following identification information: the U.S. APPLICATION NUMBER, FILING DATE, NAME OF APPLICANT, and TITLE OF INVENTION. Fees transmitted by check or draft are subject to collection. Please verify the accuracy of the data presented on this receipt. **If an error is noted on this Filing Receipt, please submit a written request for a Filing Receipt Correction. Please provide a copy of this Filing Receipt with the changes noted thereon. If you received a "Notice to File Missing Parts" for this application, please submit any corrections to this Filing Receipt with your reply to the Notice. When the USPTO processes the reply to the Notice, the USPTO will generate another Filing Receipt incorporating the requested corrections**

Applicant(s)

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Power of Attorney: None

Domestic Priority data as claimed by applicant

This application is a CIP of 12/567,716 09/25/2009
which claims benefit of 61/100,721 09/27/2008
and claims benefit of 61/108,743 10/27/2008
and claims benefit of 61/147,386 01/26/2009
and claims benefit of 61/152,086 02/12/2009
and claims benefit of 61/178,508 05/15/2009
and claims benefit of 61/182,768 06/01/2009
and claims benefit of 61/121,159 12/09/2008
and claims benefit of 61/142,977 01/07/2009
and claims benefit of 61/142,885 01/06/2009
and claims benefit of 61/142,796 01/06/2009
and claims benefit of 61/142,889 01/06/2009
and claims benefit of 61/142,880 01/06/2009
and claims benefit of 61/142,818 01/06/2009
and claims benefit of 61/142,887 01/06/2009

Application Data Sheet 37 CFR 1.76	Attorney Docket Number	WTCY-0026-P07
	Application Number	

Title of Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS

The application data sheet is part of the provisional or nonprovisional application for which it is being submitted. The following form contains the bibliographic data arranged in a format specified by the United States Patent and Trademark Office as outlined in 37 CFR 1.76.
This document may be completed electronically and submitted to the Office in electronic format using the Electronic Filing System (EFS) or the document may be printed and included in a paper filed application.

Secrecy Order 37 CFR 5.2

☐ Portions or all of the application associated with this Application Data Sheet may fall under a Secrecy Order pursuant to 37 CFR 5.2 (Paper filers only. Applications that fall under Secrecy Order may not be filed electronically.)

Applicant Information:

Applicant 1 Remove

Applicant Authority ◉ Inventor ◯ Legal Representative under 35 U.S.C. 117 ◯ Party of Interest under 35 U.S.C. 118

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Application Data Sheet 37 CFR 1.76	Attorney Docket Number	WTCY-0026-P07
	Application Number	
Title of Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS	

Citizenship under 37 CFR 1.41(b) i	US				
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	Marin		Soljacic		
Residence Information (Select One)	⦿ US Residency	◯ Non US Residency	◯ Active US Military Service		
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Citizenship under 37 CFR 1.41(b) i	HR				

Application Data Sheet 37 CFR 1.76	Attorney Docket Number	WTCY-0026-P07
	Application Number	

Title of Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS

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Applicant 7 Remove

Applicant Authority	◉ Inventor	◯ Legal Representative under 35 U.S.C. 117	◯ Party of Interest under 35 U.S.C. 118

Prefix	Given Name	Middle Name	Family Name	Suffix
	Morris	P.	Kesler	

Residence Information (Select One)	◉ US Residency	◯ Non US Residency	◯ Active US Military Service

City	Bedford	State/Province	MA	Country of Residence i	US

Citizenship under 37 CFR 1.41(b) i	US

Mailing Address of Applicant:			
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Address 2			
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Postal Code	01730	Country i	US

All Inventors Must Be Listed - Additional Inventor Information blocks may be generated within this form by selecting the **Add** button. Add

Correspondence Information:

**Enter either Customer Number or complete the Correspondence Information section below.
For further information see 37 CFR 1.33(a).**

☐ **An Address is being provided for the correspondence Information of this application.**

Customer Number	43520
Email Address	patents@strategic-patents.com

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Application Information:

Title of the Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS		
Attorney Docket Number	WTCY-0026-P07	Small Entity Status Claimed	☐
Application Type	Nonprovisional		
Subject Matter	Utility		
Suggested Class (if any)		Sub Class (if any)	
Suggested Technology Center (if any)			
Total Number of Drawing Sheets (if any)	53	Suggested Figure for Publication (if any)	

Application Data Sheet 37 CFR 1.76	Attorney Docket Number	WTCY-0026-P07
	Application Number	
Title of Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS	

Publication Information:

- [ ] Request Early Publication (Fee required at time of Request 37 CFR 1.219)
- [ ] **Request Not to Publish.** I hereby request that the attached application not be published under 35 U.S.C. 122(b) and certify that the invention disclosed in the attached application **has not and will not** be the subject of an application filed in another country, or under a multilateral international agreement, that requires publication at eighteen months after filing.

Representative Information:

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Please Select One:	(●) Customer Number	( ) US Patent Practitioner	( ) Limited Recognition (37 CFR 11.9)
Customer Number	43520		

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This section allows for the applicant to either claim benefit under 35 U.S.C. 119(e), 120, 121, or 365(c) or indicate National Stage entry from a PCT application. Providing this information in the application data sheet constitutes the specific reference required by 35 U.S.C. 119(e) or 120, and 37 CFR 1.78(a)(2) or CFR 1.78(a)(4), and need not otherwise be made part of the specification.

Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/100721	2008-09-27
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/108743	2008-10-27
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/147386	2009-01-26
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/152086	2009-02-12
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/178508	2009-05-15
Prior Application Status	Pending		Remove

Application Data Sheet 37 CFR 1.76	Attorney Docket Number	WTCY-0026-P07
	Application Number	

Title of Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS

Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/182768	2009-06-01
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/121159	2008-12-09
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/142977	2009-01-07
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/142885	2009-01-06
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/142796	2009-01-06
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/142889	2009-01-06
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/142880	2009-01-06
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/142818	2009-01-06
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/142887	2009-01-06
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/156764	2009-03-02
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/143058	2009-01-07
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/152390	2009-02-13
Prior Application Status	Pending		Remove

Application Data Sheet 37 CFR 1.76	Attorney Docket Number	WTCY-0026-P07
	Application Number	
Title of Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS	

Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/163695	2009-03-26
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/172633	2009-04-24
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/169240	2009-04-14
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
12/567716	non provisional of	61/173747	2009-04-29
Prior Application Status	Pending		Remove
Application Number	Continuity Type	Prior Application Number	Filing Date (YYYY-MM-DD)
	Continuation in part of	12/567716	2009-09-25

Additional Domestic Benefit/National Stage Data may be generated within this form by selecting the **Add** button. Add

Foreign Priority Information:

This section allows for the applicant to claim benefit of foreign priority and to identify any prior foreign application for which priority is not claimed. Providing this information in the application data sheet constitutes the claim for priority as required by 35 U.S.C. 119(b) and 37 CFR 1.55(a).

Remove

Application Number	Country i	Parent Filing Date (YYYY-MM-DD)	Priority Claimed
			(•) Yes ( ) No

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Assignee Information:

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Assignee 1 Remove

If the Assignee is an Organization check here. ☐

Prefix	**Given Name**	Middle Name	**Family Name**	Suffix

Application Data Sheet 37 CFR 1.76	Attorney Docket Number	WTCY-0026-P07
	Application Number	

Title of Invention	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS

Mailing Address Information:			
Address 1			
Address 2			
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Country i		Postal Code	
Phone Number		Fax Number	
Email Address			
Additional Assignee Data may be generated within this form by selecting the **Add** button.		Add	

Signature:

A signature of the applicant or representative is required in accordance with 37 CFR 1.33 and 10.18. Please see 37 CFR 1.4(d) for the form of the signature.

Signature	/John Nortrup/			Date (YYYY-MM-DD)	2010-03-15
First Name	John	Last Name	Nortrup	Registration Number	59063

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The information provided by you in this form will be subject to the following routine uses:

1. The information on this form will be treated confidentially to the extent allowed under the Freedom of Information Act (5 U.S.C. 552) and the Privacy Act (5 U.S.C. 552a). Records from this system of records may be disclosed to the Department of Justice to determine whether the Freedom of Information Act requires disclosure of these records.

2. A record from this system of records may be disclosed, as a routine use, in the course of presenting evidence to a court, magistrate, or administrative tribunal, including disclosures to opposing counsel in the course of settlement negotiations.

3. A record in this system of records may be disclosed, as a routine use, to a Member of Congress submitting a request involving an individual, to whom the record pertains, when the individual has requested assistance from the Member with respect to the subject matter of the record.

4. A record in this system of records may be disclosed, as a routine use, to a contractor of the Agency having need for the information in order to perform a contract. Recipients of information shall be required to comply with the requirements of the Privacy Act of 1974, as amended, pursuant to 5 U.S.C. 552a(m).

5. A record related to an International Application filed under the Patent Cooperation Treaty in this system of records may be disclosed, as a routine use, to the International Bureau of the World Intellectual Property Organization, pursuant to the Patent Cooperation Treaty.

6. A record in this system of records may be disclosed, as a routine use, to another federal agency for purposes of National Security review (35 U.S.C. 181) and for review pursuant to the Atomic Energy Act (42 U.S.C. 218(c)).

7. A record from this system of records may be disclosed, as a routine use, to the Administrator, General Services, or his/her designee, during an inspection of records conducted by GSA as part of that agency's responsibility to recommend improvements in records management practices and programs, under authority of 44 U.S.C. 2904 and 2906. Such disclosure shall be made in accordance with the GSA regulations governing inspection of records for this purpose, and any other relevant (i.e., GSA or Commerce) directive. Such disclosure shall not be used to make determinations about individuals.

8. A record from this system of records may be disclosed, as a routine use, to the public after either publication of the application pursuant to 35 U.S.C. 122(b) or issuance of a patent pursuant to 35 U.S.C. 151. Further, a record may be disclosed, subject to the limitations of 37 CFR 1.14, as a routine use, to the public if the record was filed in an application which became abandoned or in which the proceedings were terminated and which application is referenced by either a published application, an application open to public inspections or an issued patent.

9. A record from this system of records may be disclosed, as a routine use, to a Federal, State, or local law enforcement agency, if the USPTO becomes aware of a violation or potential violation of law or regulation.

Electronic Patent Application Fee Transmittal

Application Number:	12647763
Filing Date:	28-Dec-2009
Title of Invention:	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS
First Named Inventor/Applicant Name:	Aristeidis Karalis
Filer:	John H. Nortrop
Attorney Docket Number:	WTCY-0026-P07

Filed as Large Entity

Utility under 35 USC 111(a) Filing Fees

Description	Fee Code	Quantity	Amount	Sub-Total in USD($)
Basic Filing:				
Utility application filing	1011	1	330	330
Utility Search Fee	1111	1	540	540
Utility Examination Fee	1311	1	220	220
Pages:				
Utility Appl Size fee per 50 sheets >100	1081	1	270	270
Claims:				
Miscellaneous-Filing:				
Late filing fee for oath or declaration	1051	1	130	130

Description	Fee Code	Quantity	Amount	Sub-Total in USD($)
Petition:				
Patent-Appeals-and-Interference:				
Post-Allowance-and-Post-Issuance:				
Extension-of-Time:				
Miscellaneous:				
	Total in USD ($)			**1490**

Electronic Acknowledgement Receipt

EFS ID:	7212935
Application Number:	12647763
International Application Number:	
Confirmation Number:	2576
Title of Invention:	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS
First Named Inventor/Applicant Name:	Aristeidis Karalis
Customer Number:	43520
Filer:	John H. Nortrop
Filer Authorized By:	
Attorney Docket Number:	WTCY-0026-P07
Receipt Date:	15-MAR-2010
Filing Date:	28-DEC-2009
Time Stamp:	19:11:01
Application Type:	Utility under 35 USC 111(a)

Payment information:

Submitted with Payment	yes
Payment Type	Deposit Account
Payment was successfully received in RAM	$1490
RAM confirmation Number	5409
Deposit Account	505087
Authorized User	

The Director of the USPTO is hereby authorized to charge indicated fees and credit any overpayment as follows:

Charge any Additional Fees required under 37 C.F.R. Section 1.16 (National application filing, search, and examination fees)

Charge any Additional Fees required under 37 C.F.R. Section 1.17 (Patent application and reexamination processing fees)

File Listing:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
1	Applicant Response to Pre-Exam Formalities Notice	WTCY-0026-P07_031510_MissingParts.pdf	114451 198b351b55ae74ffb623e63e032fcf4037eafdcc	no	2
Warnings:					
Information:					
2	Oath or Declaration filed	WTCY-0026-P07_031510_ExecutedDecPOA.pdf	615892 1008f7bb75f69e8743a90a3beea6b55dbc59cf80	no	4
Warnings:					
Information:					
3		WTCY-0026-P07_031510_PreliminaryAmendment.pdf	91455 77e002658f21f0cd3dd92030ee9152faf164c498	yes	2

Multipart Description/PDF files in .zip description

Document Description	Start	End
Preliminary Amendment	1	1
Applicant Arguments/Remarks Made in an Amendment	2	2

Warnings:					
Information:					
4		WTCY-0026-P07_031510_SubSpecificationClean.pdf	744808 c1db0d8aae2c839a6a2161da6b8600e60533f6d5	yes	142

Multipart Description/PDF files in .zip description

Document Description	Start	End
Specification	1	141
Abstract	142	142

Warnings:					
Information:					
5		WTCY-0026-P07_031510_SubSpecificationMarked.pdf	744988 c7e278bc8aa53020249d23e659b0691eed7105be	yes	142

Multipart Description/PDF files in .zip description

Document Description	Start	End
Specification	1	141

	Abstract		142	142	
Warnings:					
Information:					
6	Request for Corrected Filing Receipt	WTCY-0026-P07_031510_Communication.pdf	94551 0072fe010dc6dc1b10f497e2cbf865c7c20233a5	no	1
Warnings:					
Information:					
7	Request for Corrected Filing Receipt	WTCY-0026-P07_031510_MarkedFR_Filed.pdf	164745 b23a91cde0ba2b702fa104dadfba7cb563554bf3	no	1
Warnings:					
Information:					
8	Application Data Sheet	WTCY-0026-P07_031510_SupplementalADS.pdf	1358522 f411937e5417e5aaae458bebf0b6480bda926e86	no	8
Warnings:					
Information:					
9	Fee Worksheet (PTO-875)	fee-info.pdf	38404 45a7fd20a36705a70eff65df6ad9e7bc7c65408e	no	2
Warnings:					
Information:					
		Total Files Size (in bytes):	3967816		

This Acknowledgement Receipt evidences receipt on the noted date by the USPTO of the indicated documents, characterized by the applicant, and including page counts, where applicable. It serves as evidence of receipt similar to a Post Card, as described in MPEP 503.

New Applications Under 35 U.S.C. 111
If a new application is being filed and the application includes the necessary components for a filing date (see 37 CFR 1.53(b)-(d) and MPEP 506), a Filing Receipt (37 CFR 1.54) will be issued in due course and the date shown on this Acknowledgement Receipt will establish the filing date of the application.

National Stage of an International Application under 35 U.S.C. 371
If a timely submission to enter the national stage of an international application is compliant with the conditions of 35 U.S.C. 371 and other applicable requirements a Form PCT/DO/EO/903 indicating acceptance of the application as a national stage submission under 35 U.S.C. 371 will be issued in addition to the Filing Receipt, in due course.

New International Application Filed with the USPTO as a Receiving Office
If a new international application is being filed and the international application includes the necessary components for an international filing date (see PCT Article 11 and MPEP 1810), a Notification of the International Application Number and of the International Filing Date (Form PCT/RO/105) will be issued in due course, subject to prescriptions concerning national security, and the date shown on this Acknowledgement Receipt will establish the international filing date of the application.

Electronic Acknowledgement Receipt

EFS ID:	7212935
Application Number:	12647763
International Application Number:	
Confirmation Number:	2576
Title of Invention:	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS
First Named Inventor/Applicant Name:	Aristeidis Karalis
Customer Number:	43520
Filer:	John H. Nortrop
Filer Authorized By:	
Attorney Docket Number:	WTCY-0026-P07
Receipt Date:	15-MAR-2010
Filing Date:	28-DEC-2009
Time Stamp:	19:11:01
Application Type:	Utility under 35 USC 111(a)

Payment information:

Submitted with Payment	yes
Payment Type	Deposit Account
Payment was successfully received in RAM	$1490
RAM confirmation Number	5409
Deposit Account	505087
Authorized User	

The Director of the USPTO is hereby authorized to charge indicated fees and credit any overpayment as follows:

Charge any Additional Fees required under 37 C.F.R. Section 1.16 (National application filing, search, and examination fees)

Charge any Additional Fees required under 37 C.F.R. Section 1.17 (Patent application and reexamination processing fees)

File Listing:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
1	Applicant Response to Pre-Exam Formalities Notice	WTCY-0026-P07_031510_MissingParts.pdf	114451 198b351b55ae74ffb623e63e032fcf4037eafdcc	no	2

Warnings:

Information:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
2	Oath or Declaration filed	WTCY-0026-P07_031510_ExecutedDecPOA.pdf	615892 1008f7bb75f69e8743a90a3beea6b55dbc59cf80	no	4

Warnings:

Information:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
3		WTCY-0026-P07_031510_PreliminaryAmendment.pdf	91455 77e002658f21f0cd3dd92030ee9152faf164c498	yes	2

Multipart Description/PDF files in .zip description

Document Description	Start	End
Preliminary Amendment	1	1
Applicant Arguments/Remarks Made in an Amendment	2	2

Warnings:

Information:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
4		WTCY-0026-P07_031510_SubSpecificationClean.pdf	744808 c1db0d8aae2c839a6a2161da6b8600e60533f6d5	yes	142

Multipart Description/PDF files in .zip description

Document Description	Start	End
Specification	1	141
Abstract	142	142

Warnings:

Information:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
5		WTCY-0026-P07_031510_SubSpecificationMarked.pdf	744988 c7e278bc8aa53020249d23e659b0691eed7105be	yes	142

Multipart Description/PDF files in .zip description

Document Description	Start	End
Specification	1	141

	Abstract		142	142	
Warnings:					
Information:					
6	Request for Corrected Filing Receipt	WTCY-0026-P07_031510_Communication.pdf	94551 0072fe010dc6dc1b10f497e2cbf865c7c20233a5	no	1
Warnings:					
Information:					
7	Request for Corrected Filing Receipt	WTCY-0026-P07_031510_MarkedFR_Filed.pdf	164745 b23a91cde0ba2b702fa104dadfba7cb563554bf3	no	1
Warnings:					
Information:					
8	Application Data Sheet	WTCY-0026-P07_031510_SupplementalADS.pdf	1358522 f411937e5417e5aaae458bebf0b6480bda926e86	no	8
Warnings:					
Information:					
9	Fee Worksheet (PTO-875)	fee-info.pdf	38404 45a7fd20a36705a70eff65df6ad9e7bc7c65408e	no	2
Warnings:					
Information:					
		Total Files Size (in bytes):	3967816		

This Acknowledgement Receipt evidences receipt on the noted date by the USPTO of the indicated documents, characterized by the applicant, and including page counts, where applicable. It serves as evidence of receipt similar to a Post Card, as described in MPEP 503.

<u>New Applications Under 35 U.S.C. 111</u>
If a new application is being filed and the application includes the necessary components for a filing date (see 37 CFR 1.53(b)-(d) and MPEP 506), a Filing Receipt (37 CFR 1.54) will be issued in due course and the date shown on this Acknowledgement Receipt will establish the filing date of the application.

<u>National Stage of an International Application under 35 U.S.C. 371</u>
If a timely submission to enter the national stage of an international application is compliant with the conditions of 35 U.S.C. 371 and other applicable requirements a Form PCT/DO/EO/903 indicating acceptance of the application as a national stage submission under 35 U.S.C. 371 will be issued in addition to the Filing Receipt, in due course.

<u>New International Application Filed with the USPTO as a Receiving Office</u>
If a new international application is being filed and the international application includes the necessary components for an international filing date (see PCT Article 11 and MPEP 1810), a Notification of the International Application Number and of the International Filing Date (Form PCT/RO/105) will be issued in due course, subject to prescriptions concerning national security, and the date shown on this Acknowledgement Receipt will establish the international filing date of the application.

UNITED STATES DEPARTMENT OF COMMERCE
United States Patent and Trademark Office
Address: COMMISSIONER FOR PATENTS
P.O. Box 1450
Alexandria, Virginia 22313-1450
www.uspto.gov

APPLICATION NUMBER	FILING or 371(c) DATE	GRP ART UNIT	FIL FEE REC'D	ATTY.DOCKET.NO	TOT CLAIMS	IND CLAIMS
12/647,763	12/28/2009	2828	1490	WTCY-0026-P07	14	2

CONFIRMATION NO. 2576

UPDATED FILING RECEIPT

43520
STRATEGIC PATENTS P.C..
C/O PORTFOLIOIP
P.O. BOX 52050
MINNEAPOLIS, MN 55402

OC000000040796704

Date Mailed: 03/25/2010

Receipt is acknowledged of this non-provisional patent application. The application will be taken up for examination in due course. Applicant will be notified as to the results of the examination. Any correspondence concerning the application must include the following identification information: the U.S. APPLICATION NUMBER, FILING DATE, NAME OF APPLICANT, and TITLE OF INVENTION. Fees transmitted by check or draft are subject to collection. Please verify the accuracy of the data presented on this receipt. **If an error is noted on this Filing Receipt, please submit a written request for a Filing Receipt Correction. Please provide a copy of this Filing Receipt with the changes noted thereon. If you received a "Notice to File Missing Parts" for this application, please submit any corrections to this Filing Receipt with your reply to the Notice. When the USPTO processes the reply to the Notice, the USPTO will generate another Filing Receipt incorporating the requested corrections**

Applicant(s)

Aristeidis Karalis, Boston, MA;
Andre B. Kurs, Chestnut Hill, MA;
Andrew J. Campanella, Waltham, MA;
Konrad J. Kulikowski, Somerville, MA;
Katherine L. Hall, Westford, MA;
Marin Soljacic, Belmont, MA;
Morris P. Kesler, Bedford, MA;

Power of Attorney: The patent practitioners associated with Customer Number 43520

Domestic Priority data as claimed by applicant

This application is a CIP of 12/567,716 09/25/2009
which claims benefit of 61/100,721 09/27/2008
and claims benefit of 61/108,743 10/27/2008
and claims benefit of 61/147,386 01/26/2009
and claims benefit of 61/152,086 02/12/2009
and claims benefit of 61/178,508 05/15/2009
and claims benefit of 61/182,768 06/01/2009
and claims benefit of 61/121,159 12/09/2008
and claims benefit of 61/142,977 01/07/2009
and claims benefit of 61/142,885 01/06/2009
and claims benefit of 61/142,796 01/06/2009
and claims benefit of 61/142,889 01/06/2009
and claims benefit of 61/142,880 01/06/2009
and claims benefit of 61/142,818 01/06/2009
and claims benefit of 61/142,887 01/06/2009

and claims benefit of 61/156,764 03/02/2009
and claims benefit of 61/143,058 01/07/2009
and claims benefit of 61/152,390 02/13/2009
and claims benefit of 61/163,695 03/26/2009
and claims benefit of 61/172,633 04/24/2009
and claims benefit of 61/169,240 04/14/2009
and claims benefit of 61/173,747 04/29/2009

Foreign Applications

Projected Publication Date: To Be Determined - pending completion of Security Review

Non-Publication Request: No

Early Publication Request: No

Title

WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS

Preliminary Class

372

PROTECTING YOUR INVENTION OUTSIDE THE UNITED STATES

Since the rights granted by a U.S. patent extend only throughout the territory of the United States and have no effect in a foreign country, an inventor who wishes patent protection in another country must apply for a patent in a specific country or in regional patent offices. Applicants may wish to consider the filing of an international application under the Patent Cooperation Treaty (PCT). An international (PCT) application generally has the same effect as a regular national patent application in each PCT-member country. The PCT process **simplifies** the filing of patent applications on the same invention in member countries, but **does not result** in a grant of "an international patent" and does not eliminate the need of applicants to file additional documents and fees in countries where patent protection is desired.

Almost every country has its own patent law, and a person desiring a patent in a particular country must make an application for patent in that country in accordance with its particular laws. Since the laws of many countries differ in various respects from the patent law of the United States, applicants are advised to seek guidance from specific foreign countries to ensure that patent rights are not lost prematurely.

Applicants also are advised that in the case of inventions made in the United States, the Director of the USPTO must issue a license before applicants can apply for a patent in a foreign country. The filing of a U.S. patent application serves as a request for a foreign filing license. The application's filing receipt contains further information and guidance as to the status of applicant's license for foreign filing.

Applicants may wish to consult the USPTO booklet, "General Information Concerning Patents" (specifically, the section entitled "Treaties and Foreign Patents") for more information on timeframes and deadlines for filing foreign patent applications. The guide is available either by contacting the USPTO Contact Center at 800-786-9199, or it can be viewed on the USPTO website at http://www.uspto.gov/web/offices/pac/doc/general/index.html.

For information on preventing theft of your intellectual property (patents, trademarks and copyrights), you may wish to consult the U.S. Government website, http://www.stopfakes.gov. Part of a Department of Commerce initiative, this website includes self-help "toolkits" giving innovators guidance on how to protect intellectual property in specific countries such as China, Korea and Mexico. For questions regarding patent enforcement issues, applicants may call the U.S. Government hotline at 1-866-999-HALT (1-866-999-4158).

LICENSE FOR FOREIGN FILING UNDER

Title 35, United States Code, Section 184

Title 37, Code of Federal Regulations, 5.11 & 5.15

GRANTED

The applicant has been granted a license under 35 U.S.C. 184, if the phrase "IF REQUIRED, FOREIGN FILING LICENSE GRANTED" followed by a date appears on this form. Such licenses are issued in all applications where the conditions for issuance of a license have been met, regardless of whether or not a license may be required as set forth in 37 CFR 5.15. The scope and limitations of this license are set forth in 37 CFR 5.15(a) unless an earlier license has been issued under 37 CFR 5.15(b). The license is subject to revocation upon written notification. The date indicated is the effective date of the license, unless an earlier license of similar scope has been granted under 37 CFR 5.13 or 5.14.

This license is to be retained by the licensee and may be used at any time on or after the effective date thereof unless it is revoked. This license is automatically transferred to any related applications(s) filed under 37 CFR 1.53(d). This license is not retroactive.

The grant of a license does not in any way lessen the responsibility of a licensee for the security of the subject matter as imposed by any Government contract or the provisions of existing laws relating to espionage and the national security or the export of technical data. Licensees should apprise themselves of current regulations especially with respect to certain countries, of other agencies, particularly the Office of Defense Trade Controls, Department of State (with respect to Arms, Munitions and Implements of War (22 CFR 121-128)); the Bureau of Industry and Security, Department of Commerce (15 CFR parts 730-774); the Office of Foreign AssetsControl, Department of Treasury (31 CFR Parts 500+) and the Department of Energy.

NOT GRANTED

No license under 35 U.S.C. 184 has been granted at this time, if the phrase "IF REQUIRED, FOREIGN FILING LICENSE GRANTED" DOES NOT appear on this form. Applicant may still petition for a license under 37 CFR 5.12, if a license is desired before the expiration of 6 months from the filing date of the application. If 6 months has lapsed from the filing date of this application and the licensee has not received any indication of a secrecy order under 35 U.S.C. 181, the licensee may foreign file the application pursuant to 37 CFR 5.15(b).

Substitute for form 1449A/PTO

INFORMATION DISCLOSURE STATEMENT BY APPLICANT

(Use as many sheets as necessary)

Complete if Known	
Application Number	12/647,763
Filing Date	Dec 28, 2009
First Named Inventor	Aristeidis Karalis
Art Unit	2828
Examiner Name	Not Yet Known

Sheet	1	of	1	Attorney Docket No: WTCY-0026-P07

US PATENT DOCUMENTS

Examiner Initial *	Cite No	Document Number	Publication Date	Name of Patentee or Applicant of Cited Document	Pages, Columns, Lines, Where Relevant Passages or RelevantFigures Appear
		20070222542A1	Sep 27, 2007	Joannopoulos, J. D., et al.	
		20090085706	Apr 2, 2009	Baarman, David W., et al.	
		20090230777	Sep 17, 2009	Baarman, David W., et al.	
		6452465B1	Sep 17, 2002	Brown, B. et al.	
		7492247	Feb 17, 2009	Schmidt, Josef et al.	

FOREIGN PATENT DOCUMENTS

Examiner Initials*	Cite No	Foreign Patent Document	Publication Date	Name of Patentee or Applicant of cited Document	Pages, Columns, Lines, Where Relevant Passages or RelevantFigures Appear	T[2]

OTHER DOCUMENTS -- NON PATENT LITERATURE DOCUMENTS

Examiner Initials*	Cite No [1]	Include name of the author (in CAPITAL LETTERS), title of the article (when appropriate), title of the item (book, magazine, journal, serial, symposium, catalog, etc.), date, page(s), volume-issue number(s), publisher, city and/or country where published.	T[2]
	NPL-1	, "International Application Serial No. PCT/US09/58499, Search Report and Written Opinion mailed 12-10-2009",	

EXAMINER **DATE CONSIDERED**

Substitute Disclosure Statement Form (PTO-1449)
* EXAMINER: Initial if reference considered, whether or not citation is in conformance with MPEP 609. Draw line through citation if not in conformance and not considered. Include copy of this form with next communication to applicant.1 Applicant's unique citation designation number (optional) 2 Applicant is to place a check mark here if English language Translation is attached

PATENT COOPERATION TREATY

From the INTERNATIONAL SEARCHING AUTHORITY

PCT

To:
ROBERT A. MAZZARESE
STRATEGIC PATENTS, P.C.
C/O INTELLEVATE
P.O. BOX 52050
MINNEAPOLIS, MN 55402

NOTIFICATION OF TRANSMITTAL OF THE INTERNATIONAL SEARCH REPORT AND THE WRITTEN OPINION OF THE INTERNATIONAL SEARCHING AUTHORITY, OR THE DECLARATION

(PCT Rule 44.1)

	Date of mailing *(day/month/year)*
Applicant's or agent's file reference WTCY0026PWO	**FOR FURTHER ACTION** See paragraphs 1 and 4 below
International application No. PCT/US 09/58499	International filing date *(day/month/year)* 25 September 2009 (25.09.2009)
Applicant WITRICITY CORPORATION	

80157,E92WO1

1. ☒ The applicant is hereby notified that the international search report and the written opinion of the International Searching Authority have been established and are transmitted herewith.

   Filing of amendments and statement under Article 19:
   The applicant is entitled, if he so wishes, to amend the claims of the international application (see Rule 46):

   When? The time limit for filing such amendments is normally two months from the date of transmittal of the international search report.

   Where? Directly to the International Bureau of WIPO, 34 chemin des Colombettes 1211 Geneva 20, Switzerland, Facsimile No.: +41 22 338 8270

   For more detailed instructions, see the notes on the accompanying sheet.

2. ☐ The applicant is hereby notified that no international search report will be established and that the declaration under Article 17(2)(a) to that effect and the written opinion of the International Searching Authority are transmitted herewith.

3. ☐ **With regard to the protest** against payment of (an) additional fee(s) under Rule 40.2, the applicant is notified that:

   ☐ the protest together with the decision thereon has been transmitted to the International Bureau together with the applicant's request to forward the texts of both the protest and the decision thereon to the designated Offices.

   ☐ no decision has been made yet on the protest; the applicant will be notified as soon as a decision is made.

4. **Reminders**

   Shortly after the expiration of **18 months** from the priority date, the international application will be published by the International Bureau. If the applicant wishes to avoid or postpone publication, a notice of withdrawal of the international application, or of the priority claim, must reach the International Bureau as provided in Rules 90*bis*.1 and 90*bis*.3, respectively, before the completion of the technical preparations for international publication.

   The applicant may submit comments on an informal basis on the written opinion of the International Searching Authority to the International Bureau. The International Bureau will send a copy of such comments to all designated Offices unless an international preliminary examination report has been or is to be established. These comments would also be made available to the public but not before the expiration of 30 months from the priority date.

   Within **19 months** from the priority date, but only in respect of some designated Offices, a demand for international preliminary examination must be filed if the applicant wishes to postpone the entry into the national phase **until 30 months** from the priority date (in some Offices even later); otherwise, the applicant must, **within 20 months** from the priority date, perform the prescribed acts for entry into the national phase before those designated Offices.

   In respect of other designated Offices, the time limit of **30 months** (or later) will apply even if no demand is filed within 19 months.

   See the Annex to Form PCT/IB/301 and, for details about the applicable time limits, Office by Office, see the *PCT Applicant's Guide*, Volume II, National Chapters and the WIPO Internet site.

Name and mailing address of the ISA/US Mail Stop PCT, Attn: ISA/US Commissioner for Patents P.O. Box 1450, Alexandria, Virginia 22313-1450 Facsimile No. 571-273-3201	Authorized officer: Lee W. Young PCT Helpdesk: 571-272-4300 PCT OSP: 571-272-7774

Form PCT/ISA/220 (January 2004)

(See notes on accompanying sheet)

CPA GLOBAL
DEC 14 2009
RECEIVED
Coded ____ Verified ____

PATENT COOPERATION TREATY

From the INTERNATIONAL SEARCHING AUTHORITY

To:
ROBERT A. MAZZARESE
STRATEGIC PATENTS, P.C.
C/O INTELLEVATE
P.O. BOX 52050
MINNEAPOLIS, MN 55402

PCT

NOTIFICATION OF TRANSMITTAL OF THE INTERNATIONAL SEARCH REPORT AND THE WRITTEN OPINION OF THE INTERNATIONAL SEARCHING AUTHORITY, OR THE DECLARATION

(PCT Rule 44.1)

	Date of mailing *(day/month/year)* 10 DEC 2009
Applicant's or agent's file reference WTCY0026PWO	**FOR FURTHER ACTION** See paragraphs 1 and 4 below
International application No. PCT/US 09/58499	International filing date *(day/month/year)* 25 September 2009 (25.09.2009)
Applicant WITRICITY CORPORATION	

1. ☒ The applicant is hereby notified that the international search report and the written opinion of the International Searching Authority have been established and are transmitted herewith.

   Filing of amendments and statement under Article 19:
   The applicant is entitled, if he so wishes, to amend the claims of the international application (see Rule 46):

   When? The time limit for filing such amendments is normally two months from the date of transmittal of the international search report.

   Where? Directly to the International Bureau of WIPO, 34 chemin des Colombettes 1211 Geneva 20, Switzerland, Facsimile No.: +41 22 338 8270

   For more detailed instructions, see the notes on the accompanying sheet.

2. ☐ The applicant is hereby notified that no international search report will be established and that the declaration under Article 17(2)(a) to that effect and the written opinion of the International Searching Authority are transmitted herewith.

3. ☐ **With regard to the protest** against payment of (an) additional fee(s) under Rule 40.2, the applicant is notified that:

   ☐ the protest together with the decision thereon has been transmitted to the International Bureau together with the applicant's request to forward the texts of both the protest and the decision thereon to the designated Offices.

   ☐ no decision has been made yet on the protest; the applicant will be notified as soon as a decision is made.

4. **Reminders**

   Shortly after the expiration of **18 months** from the priority date, the international application will be published by the International Bureau. If the applicant wishes to avoid or postpone publication, a notice of withdrawal of the international application, or of the priority claim, must reach the International Bureau as provided in Rules 90*bis*.1 and 90*bis*.3, respectively, before the completion of the technical preparations for international publication.

   The applicant may submit comments on an informal basis on the written opinion of the International Searching Authority to the International Bureau. The International Bureau will send a copy of such comments to all designated Offices unless an international preliminary examination report has been or is to be established. These comments would also be made available to the public but not before the expiration of 30 months from the priority date.

   Within **19 months** from the priority date, but only in respect of some designated Offices, a demand for international preliminary examination must be filed if the applicant wishes to postpone the entry into the national phase **until 30 months** from the priority date (in some Offices even later); otherwise, the applicant must, **within 20 months** from the priority date, perform the prescribed acts for entry into the national phase before those designated Offices.

   In respect of other designated Offices, the time limit of **30 months** (or later) will apply even if no demand is filed within 19 months.

   See the Annex to Form PCT/IB/301 and, for details about the applicable time limits, Office by Office, see the *PCT Applicant's Guide*, Volume II, National Chapters and the WIPO Internet site.

Name and mailing address of the ISA/US Mail Stop PCT, Attn: ISA/US Commissioner for Patents P.O. Box 1450, Alexandria, Virginia 22313-1450 Facsimile No. 571-273-3201	Authorized officer: Lee W. Young PCT Helpdesk: 571-272-4300 PCT OSP: 571-272-7774

Form PCT/ISA/220 (January 2004) *(See notes on accompanying sheet)*

PATENT COOPERATION TREATY

PCT

INTERNATIONAL SEARCH REPORT

(PCT Article 18 and Rules 43 and 44)

Applicant's or agent's file reference WTCY0026PWO	**FOR FURTHER ACTION**	see Form PCT/ISA/220 as well as, where applicable, item 5 below.
International application No. PCT/US 09/58499	International filing date *(day/month/year)* 25 September 2009 (25.09.2009)	(Earliest) Priority Date *(day/month/year)* 27 September 2008 (27.09.2008)
Applicant WITRICITY CORPORATION		

This international search report has been prepared by this International Searching Authority and is transmitted to the applicant according to Article 18. A copy is being transmitted to the International Bureau.

This international search report consists of a total of 2 sheets.

☐ It is also accompanied by a copy of each prior art document cited in this report.

1. **Basis of the report**
   a. With regard to the **language**, the international search was carried out on the basis of:
      ☒ the international application in the language in which it was filed.
      ☐ a translation of the international application into ______________ which is the language of a translation furnished for the purposes of international search (Rules 12.3(a) and 23.1(b)).
   b. ☐ This international search report has been established taking into account the **rectification of an obvious mistake** authorized by or notified to this Authority under Rule 91 (Rule 43.6*bis*(a)).
   c. ☐ With regard to any **nucleotide and/or amino acid sequence** disclosed in the international application, see Box No. I.
2. ☐ **Certain claims were found unsearchable** (see Box No. II).
3. ☐ **Unity of invention is lacking** (see Box No. III).
4. With regard to the **title**,
   ☒ the text is approved as submitted by the applicant.
   ☐ the text has been established by this Authority to read as follows:
5. With regard to the **abstract**,
   ☒ the text is approved as submitted by the applicant.
   ☐ the text has been established, according to Rule 38.2, by this Authority as it appears in Box No. IV. The applicant may, within one month from the date of mailing of this international search report, submit comments to this Authority.
6. With regard to the **drawings**,
   a. the figure of the **drawings** to be published with the abstract is Figure No. 1
      ☒ as suggested by the applicant.
      ☐ as selected by this Authority, because the applicant failed to suggest a figure.
      ☐ as selected by this Authority, because this figure better characterizes the invention.
   b. ☐ none of the figures is to be published with the abstract.

Form PCT/ISA/210 (first sheet) (July 2009)

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US 09/58499

A. CLASSIFICATION OF SUBJECT MATTER
IPC(8) - H03B 19/00 (2009.01)
USPC - 327/113
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
USPC: 327/113

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
USPC: 327/113, 306, 530, 555; 375/323; 307/134 (keyword limited - see terms below)

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)
PubWEST (PGPB, USPT, USOC, EPAB, JPAB); GOOGLE
Search Terms: energy, power, power generator, generator, wireless, resonator, first resonator, second resonator, third resonator, Q-factor, distance, tunable, oscillating, impedance, capacitance, load

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y	US 2007/0222542 A1 (Joannopoulos et al.) 27 September 2007 (27.09.2007), entire document, especially; abstract, para. [0004], [0005], [0013], [0014], [0019], [0023], [0025], [0029], [0033]	1 - 26
Y	US 6,452,465 B1 (Brown et al.) 17 September 2002 (17.09.2002), entire document, especially; abstract, col. 2, ln 4-6, col. 3, ln 7-12, 66-67	1 - 26

☐ Further documents are listed in the continuation of Box C. ☐

* Special categories of cited documents:
- "A" document defining the general state of the art which is not considered to be of particular relevance
- "E" earlier application or patent but published on or after the international filing date
- "L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
- "O" document referring to an oral disclosure, use, exhibition or other means
- "P" document published prior to the international filing date but later than the priority date claimed
- "T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
- "X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
- "Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
- "&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
24 November 2009 (24.11.2009)	10 DEC 2009

Name and mailing address of the ISA/US	Authorized officer:
Mail Stop PCT, Attn: ISA/US, Commissioner for Patents P.O. Box 1450, Alexandria, Virginia 22313-1450 Facsimile No. 571-273-3201	Lee W. Young PCT Helpdesk: 571-272-4300 PCT OSP: 571-272-7774

Form PCT/ISA/210 (second sheet) (July 2009)

PATENT COOPERATION TREATY

From the
INTERNATIONAL SEARCHING AUTHORITY

To: ROBERT A. MAZZARESE
STRATEGIC PATENTS, P.C.
C/O INTELLEVATE
P.O. BOX 52050
MINNEAPOLIS, MN 55402

PCT

WRITTEN OPINION OF THE
INTERNATIONAL SEARCHING AUTHORITY

(PCT Rule 43*bis*.1)

Date of mailing *(day/month/year)*	10 DEC 2009
Applicant's or agent's file reference WTCY0026PWO	**FOR FURTHER ACTION** See paragraph 2 below

International application No.	International filing date *(day/month/year)*	Priority date *(day/month/year)*
PCT/US 09/58499	25 September 2009 (25.09.2009)	27 September 2008 (27.09.2008)

International Patent Classification (IPC) or both national classification and IPC
IPC(8) - H03B 19/00 (2009.01)
USPC - 327/113

Applicant WITRICITY CORPORATION

1. This opinion contains indications relating to the following items:

- [x] Box No. I Basis of the opinion
- [ ] Box No. II Priority
- [ ] Box No. III Non-establishment of opinion with regard to novelty, inventive step and industrial applicability
- [ ] Box No. IV Lack of unity of invention
- [x] Box No. V Reasoned statement under Rule 43*bis*.1(a)(i) with regard to novelty, inventive step or industrial applicability; citations and explanations supporting such statement
- [ ] Box No. VI Certain documents cited
- [ ] Box No. VII Certain defects in the international application
- [ ] Box No. VIII Certain observations on the international application

2. **FURTHER ACTION**

If a demand for international preliminary examination is made, this opinion will be considered to be a written opinion of the International Preliminary Examining Authority ("IPEA") except that this does not apply where the applicant chooses an Authority other than this one to be the IPEA and the chosen IPEA has notified the International Bureau under Rule 66.1*bis*(b) that written opinions of this International Searching Authority will not be so considered.

If this opinion is, as provided above, considered to be a written opinion of the IPEA, the applicant is invited to submit to the IPEA a written reply together, where appropriate, with amendments, before the expiration of 3 months from the date of mailing of Form PCT/ISA/220 or before the expiration of 22 months from the priority date, whichever expires later.

For further options, see Form PCT/ISA/220.

3. For further details, see notes to Form PCT/ISA/220.

Name and mailing address of the ISA/US	Date of completion of this opinion	Authorized officer:
Mail Stop PCT, Attn: ISA/US Commissioner for Patents P.O. Box 1450, Alexandria, Virginia 22313-1450 Facsimile No. 571-273-3201	27 November 2009 (27.11.2009)	Lee W. Young PCT Helpdesk: 571-272-4300 PCT OSP: 571-272-7774

Form PCT/ISA/237 (cover sheet) (July 2009)

WRITTEN OPINION OF THE INTERNATIONAL SEARCHING AUTHORITY

International application No.
PCT/US 09/58499

Box No. I Basis of this opinion

1. With regard to the **language**, this opinion has been established on the basis of:
   - [x] the international application in the language in which it was filed.
   - [ ] a translation of the international application into ______________ which is the language of a translation furnished for the purposes of international search (Rules 12.3(a) and 23.1(b)).

2. [ ] This opinion has been established taking into account the **rectification of an obvious mistake** authorized by or notified to this Authority under Rule 91 (Rule 43*bis*.1(a))

3. With regard to any **nucleotide and/or amino acid sequence** disclosed in the international application, this opinion has been established on the basis of a sequence listing filed or furnished:

   a. (means)
      - [ ] on paper
      - [ ] in electronic form

   b. (time)
      - [ ] in the international application as filed
      - [ ] together with the international application in electronic form
      - [ ] subsequently to this Authority for the purposes of search

4. [ ] In addition, in the case that more than one version or copy of a sequence listing has been filed or furnished, the required statements that the information in the subsequent or additional copies is identical to that in the application as filed or does not go beyond the application as filed, as appropriate, were furnished.

5. Additional comments:

Form PCT/ISA/237 (Box No. I) (July 2009)

WRITTEN OPINION OF THE INTERNATIONAL SEARCHING AUTHORITY

International application No.
PCT/US 09/58499

Box No. V Reasoned statement under Rule 43*bis*.1(a)(i) with regard to novelty, inventive step or industrial applicability; citations and explanations supporting such statement

1. Statement

Novelty (N)	Claims	1 - 26	YES
	Claims	None.	NO
Inventive step (IS)	Claims	None.	YES
	Claims	1 - 26	NO
Industrial applicability (IA)	Claims	1 - 26	YES
	Claims	None.	NO

2. Citations and explanations:

Claims 1 - 26 lack an inventive step under PCT Article 33(3) as being obvious over US 2007/0222542 A1 to Joannopoulos et al. (hereinafter 'Joannopoulos'), in view of US 6,452,465 B1 to Brown et al. (hereinafter 'Brown').

Regarding claim 1, Joannopoulos teaches a system, comprising: a source resonator having a Q-factor Q1 and a characteristic size X1 (abstract, para. [0005]), coupled to a power generator (external power supply, para. [0005]), and a second resonator having a Q-factor Q2 and a characteristic size X2 (para. [0005]), coupled to a load located a distance D from the source resonator (distance between the two resonators can be larger than the characteristic size of each resonator, para. [0005]), wherein the source resonator and the second resonator are coupled to exchange energy wirelessly among the source resonator and the second resonator (abstract, para. [0004], [0013]). Joannopoulos does not teach that the square root of Q1Q2 > 100. However, Brown teaches multiple resonators that are tunable (abstract), and that filters may be fabricated with low quality factor resonators (col. 2, ln 4-6). It would have been obvious to one skilled in the art to combine the teachings of Joannopoulos with those of Brown in order to provide an inexpensive resonator structure. Neither Joannopoulos nor Brown specifically teach that the square root of Q1Q2 > 100. However, this configuration could have been determined via routine experimentation, and provided based on the specific application of the system.

Regarding claims 2 and 3, neither Joannopoulos nor Brown teaches that Q1 (or Q2 - claim 3) < 100. However, it was well known it the art at the time of the invention that resonator materials were available with low quality factors, e.g. Q1 (or Q2 - claim 3) < 100. It would have been obvious to one skilled in the art to provide Q1 (or Q2 - claim 3) < 100, in order to provide an inexpensive resonator structure.

Regarding claims 4 and 5, neither Joannopoulos nor Brown specifically teaches a third resonator having a Q-factor Q3 configured to transfer energy non-raditively with the source and second resonators, wherein the square root of Q1Q3 > 100 and the square root of Q2Q3 > 100, wherein Q3 < 100 (claim - 5). However, this configuration could have been determined via routine experimentation, and provided based on the specific application of the system.

Regarding claim 6, Joannopoulos teaches that the source resonator is coupled to the power generator with direct electrical connections (para. [0014]).

Regarding claim 7, Joannopoulos teaches that the source resonator is coupled and impedance matched to the power generator with direct electrical connections (para. [0014]).

Regarding claim 8, Brown teaches a tunable circuit wherein the source resonator is coupled to the power generator through the tunable circuit with direct electrical connections (abstract, col. 3, ln 7-12).

Regarding claim 9, Joannopoulos teaches that at least one of the direct electrical connections is configured to substantially preserve a resonant mode of the source resonator (resonant cavity, para. [0033]).

Regarding claim 10, neither Joannopoulos nor Brown specifically teaches that the source resonator has a first terminal, a second terminal, and a center terminal, and wherein an impedance between the first terminal and the center terminal and between the second terminal and the center terminal are substantially equal. However, this configuration could have been determined via routine experimentation and provided to maximize power transfer.

Regarding claim 11, Joannopoulos teaches that the source resonator includes a capacitive loaded loop having a first terminal, a second terminal, and a center terminal (para. [0019]), and wherein an impedance between the first terminal and the center terminal and between the second terminal and the center terminal are substantially equal (para. [0019]).

Regarding claim 12, neither Joannopoulos nor Brown specifically teaches that the source resonator is coupled to an impedance matching network and the impedance matching network further comprises a first terminal, a second terminal, and a center terminal, and wherein an impedance between the first terminal and the center terminal and between the second terminal and the center terminal are substantially equal. However, this configuration could have been determined via routine experimentation and provided to maximize power transfer.

---(continued in Supplemental Box)---

Supplemental Box

In case the space in any of the preceding boxes is not sufficient.
Continuation of:

V.2 Citations and explanations:

Regarding claim 13, neither Joannopoulos nor Brown teaches that the first terminal and the second terminal are directly coupled to the power generator and driven with oscillating signals that are near 180 degrees out of phase. However, this configuration was well known in the art at the time of the invention, and could have been provided based on the specific application.

Regarding claim 14, Joannopoulos teaches that the source resonator has a resonant frequency WI and the first terminal and the second terminal are directly coupled to the power generator and driven with oscillating signals that are substantially equal to the resonant frequency omega sub 1 (para. [0023], [0025]).

Regarding claim 15, neither Joannopoulos nor Brown specifically that the center terminal is connected to an electrical ground. However, this configuration was well known in the art at the time of the invention, and could have been provided in order to provide a voltage reference point.

Regarding claim 16, neither Joannopoulos nor Brown teach that the source resonator has a resonant frequency omega sub 1 and the first terminal and the second terminal are directly coupled to the power generator and driven with a frequency substantially equal to the resonant frequency omega sub 1. However, this configuration could have been determined via routine experimentation and provided to maximize power transfer.

Regarding claim 17, Brown teaches a plurality of capacitors coupled to the power generator and the load (col., 3, ln 66-67).

Regarding claim 18, Joannopoulos that the source resonator and the second resonator are each enclosed in a low loss tangent material (para. [0023]).

Regarding claim 19, Joannopoulos teaches a power conversion circuit wherein the second resonator is coupled to the power conversion circuit to deliver DC power to the load (abstract).

Regarding claim 20, Joannopoulos teaches a power conversion circuit wherein the second resonator is coupled to the power conversion circuit to deliver AC power to the load (power, abstract, para. [0029]).

Regarding claim 21, Joannopoulos teaches a power conversion circuit, wherein the second resonator is coupled to the power conversion circuit to deliver both AC and DC power to the load (power, abstract, para. [0029]).

Regarding claim 22, Joannopoulos teaches a power conversion circuit and a plurality of loads, wherein the second resonator is coupled to the power conversion circuit (abstract), and the power conversion circuit is coupled to the plurality of loads (abstract).

Regarding claim 23, Brown teaches the system, wherein the impedance matching network comprises capacitors (col., 3, ln 66-67).

Regarding claim 24, Joannopoulos teaches that the impedance matching network comprises inductors (coils, para. [0025]).

Regarding claim 25, Brown teaches that the tunable circuit comprises variable capacitors (col., 3, ln 66-67).

Regarding claim 26, Joannopoulos teaches that the tunable circuit comprises variable inductors (para. [0025]).

Claims 1 - 26 have industrial applicability as defined by PCT Article 33(4) because the subject matter can be made or used in industry.

NOTES TO FORM PCT/ISA/220

These Notes are intended to give the basic instructions concerning the filing of amendments under Article 19. The Notes are based on the requirements of the Patent Cooperation Treaty, the Regulations and the Administrative Instructions under that Treaty. In case of discrepancy between these Notes and those requirements, the latter are applicable. For more detailed information, see also the *PCT Applicant's Guide*.

In these Notes, "Article," "Rule" and "Section" refer to the provisions of the PCT, the PCT Regulations and the PCT Administrative Instructions, respectively.

INSTRUCTIONS CONCERNING AMENDMENTS UNDER ARTICLE 19

The applicant has, after having received the international search report and the written opinion of the International Searching Authority, one opportunity to amend the claims of the international application. It should however be emphasized that, since all parts of the international application (claims, description and drawings) may be amended during the international preliminary examination procedure, there is usually no need to file amendments of the claims under Article 19 except where, e.g. the applicant wants the latter to be published for the purposes of provisional protection or has another reason for amending the claims before international publication. Furthermore, it should be emphasized that provisional protection is available in some States only (see *PCT Applicant's Guide*, Annex B).

The attention of the applicant is drawn to the fact that amendments to the claims under Article 19 are not allowed where the International Searching Authority has declared, under Article 17(2), that no international search report would be established (see *PCT Applicant's Guide*, International Phase, paragraph 296).

What parts of the international application may be amended?

Under Article 19, only the claims may be amended.

During the international phase, the claims may also be amended (or further amended) under Article 34 before the International Preliminary Examining Authority. The description and drawings may only be amended under Article 34 before the International Preliminary Examining Authority.

Upon entry into the national phase, all parts of the international application may be amended under Article 28 or, where applicable, Article 41.

When? Within 2 months from the date of transmittal of the international search report or 16 months from the priority date, whichever time limit expires later. It should be noted, however, that the amendments will be considered as having been received on time if they are received by the International Bureau after the expiration of the applicable time limit but before the completion of the technical preparations for international publication (Rule 46.1).

Where not to file the amendments?

The amendments may only be filed with the International Bureau and not with the receiving Office or the International Searching Authority (Rule 46.2).

Where a demand for international preliminary examination has been/is filed, see below.

How? Either by cancelling one or more entire claims, by adding one or more new claims or by amending the text of one or more of the claims as filed.

A replacement sheet or sheets containing a complete set of claims in replacement of all the claims previously filed must be submitted.

Where a claim is cancelled, no renumbering of the other claims is required. In all cases where claims are renumbered, they must be renumbered consecutively in Arabic numerals (Section 205(a)).

The amendments must be made in the language in which the international application is to be published.

What documents must/may accompany the amendments?

Letter (Section 205(b)):

The amendments must be submitted with a letter.

The letter will not be published with the international application and the amended claims. It should not be confused with the "Statement under Article 19(1)" (see below, under "Statement under Article 19(1)").

The letter must be in English or French, at the choice of the applicant. However, if the language of the international application is English, the letter must be in English; if the language of the international application is French, the letter must be in French.

ons concerning several claims may be grouped), whether

he claim is unchanged;

he claim is cancelled;

he claim is new;

he claim replaces one or more claims as filed;

ne claim is the result of the division of a claim as filed.

oles illustrate the manner in which amendments must be explained in the accompanying

there were 48 claims and after amendment of some claims there are 51]:
31, 32, 34, 35, 37 to 48 replaced by amended claims bearing the same numbers;
36 unchanged; new claims 49 to 51 added."

there were 15 claims and after amendment of all claims there are 11]:
eplaced by amended claims 1 to 11."

there were 14 claims and the amendments consist in cancelling some claims and in adding

d 14 unchanged; claims 7 to 13 cancelled; new claims 15, 16 and 17 added." or
ancelled; new claims 15, 16 and 17 added; all other claims unchanged."

nds of amendments are made]:
hanged; claims 11 to 13, 18 and 19 cancelled; claims 14, 15 and 16 replaced by amended
7 subdivided into amended claims 15, 16 and 17; new claims 20 and 21 added."

rticle 19(1)" (Rule 46.4)

be accompanied by a statement explaining the amendments and indicating any impact that ht have on the description and the drawings (which cannot be amended under Article 19(1)).

published with the international application and the amended claims.

guage in which the international application is to be published.

xceeding 500 words if in English or if translated into English.

ed with and does not replace the letter indicating the differences between the claims as filed ust be filed on a separate sheet and must be identified as such by a heading, preferably by ment under Article 19(1)."

y disparaging comments on the international search report or the relevance of citations t. Reference to citations, relevant to a given claim, contained in the international search ly in connection with an amendment of that claim.

ternational preliminary examination has already been filed

g any amendments and any accompanying statement, under Article 19, a demand for ry examination has already been submitted, the applicant must preferably, at the time of s (and any statement) with the International Bureau, also file with the International g Authority a copy of such amendments (and of any statement) and, where required, a endments for the procedure before that Authority (see Rules 55.3(a) and 62.2, first information, see the Notes to the demand form (PCT/IPEA/401).

ional preliminary examination is made, the written opinion of the International Searching in certain cases where the International Preliminary Examining Authority did not act as Authority and where it has notified the International Bureau under Rule 66.1*bis*(b), be en opinion of the International Preliminary Examining Authority. If a demand is made, the the International Preliminary Examining Authority a reply to the written opinion together. h amendments before the expiration of 3 months from the date of mailing of Form he expiration of 22 months from the priority date, whichever expires later (Rule 43*bis*.1(c)).

slation of the international application for entry into the national phase

n is drawn to the fact that, upon entry into the national phase, a translation of the claims as 19 may have to be furnished to the designated/elected Offices, instead of, or in addition to, ims as filed.

the requirements of each designated/elected Office, see the *PCT Applicant's Guide*.

SEQUENCE LISTINGS AND TABLES RELATED THERETO IN INTERNATIONAL APPLICATIONS FILED IN THE U.S. RECEIVING OFFICE

The Administrative Instructions (AIs) under the Patent Cooperation Treaty (PCT), in force as of **July 1, 2009**, contain important changes relating to the manner of filing, and applicable fees for, sequence listings and/or tables related thereto (sequence-related tables) in international applications. The complete text may be accessed at http://www.wipo.Int/pct/en/texts/index.htm.

Effective **July 1**, 2009, Part 8 and Annex C-*bis* will no longer form part of the AIs. Part 8 was introduced in 2001 as a temporary solution to problems arising from the filing of very large sequence listings on paper and provided for a *sequence listing forming part of the international application* to be filed in electronic form on physical medium (*e.g.*, CD), together with the remainder of the application on paper. In 2002, Part 8 was expanded to include sequence-related tables and Annex C-*bis* was added to provide technical requirements. All applicants may now file complete international applications in electronic form, eliminating the need for these temporary provisions.

I. AIS PART 8 AND ANNEX C-*BIS DELETED AS OF JULY 1, 2009*

A) **Sequence-related tables cannot be filed as a separate part of the description or in text format.** They must be provided as an integral part of the international application either:
- in PDF format as part of an international application filed in electronic form via EFS-Web; or
- on paper as part of an international application filed on paper.

B) A ***sequence listing forming part of an international application*** may be provided either:
- in electronic form, as part of an international application filed in electronic form via EFS-Web, in
  - Annex C/ST.25 text format (preferred), or
  - PDF format; or
- on paper as part of an international application filed on paper.

C) **A *sequence listing not forming part of the international application* (for search under PCT Rule 13*ter*) in Annex C/ST.25 text format**
- is not required where the *sequence listing forming part of the international application* was filed in Annex C/ST.25 text format as part of an international application filed in electronic form via EFS-Web
- is required for search where the *sequence listing forming part of the international application* was filed in PDF
- is required for search on physical medium (*e.g.*, CD) where the *sequence listing forming part of the international application* was filed on paper as part of an international application filed on paper.

II. CALCULATION OF THE INTERNATIONAL FILING FEE AND FEE REDUCTION UNDER AI § 707

A) A **sequence-related table** must form an integral part of the international application and will incur FULL page fees with no upper limit.

B) **A *sequence listing forming part of an international application*** filed:
- via EFS-Web in Annex C/ST.25 text format will incur NO page fees;
- on paper or in PDF format will incur FULL page fees with no upper limit.

III. AVAILABILITY OF SEQUENCE LISTINGS SUBMITTED FOR SEARCH UNDER PCT RULE *13TER*

International Searching Authorities will be required to transmit to the International Bureau a copy of an Annex C/ST.25 text format sequence listing provided for search under PCT Rule 13*ter*. Any such sequence listing will be made available on PATENTSCOPE® (*sequence listings forming part of the international application* are already available).

IV. JULY 2009 REQUEST (PCT/RO/101)

The Request now has two options for the last sheet: one for paper filings; and one for EFS-Web filings. The July 2009 Request may be accessed at http://www.wipo.int/pct/en/forms/index.htm

Electronic Acknowledgement Receipt

EFS ID:	7303333
Application Number:	12647763
International Application Number:	
Confirmation Number:	2576
Title of Invention:	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS
First Named Inventor/Applicant Name:	Aristeidis Karalis
Customer Number:	43520
Filer:	John H. Nortrop/Elizabeth Nortrup
Filer Authorized By:	John H. Nortrop
Attorney Docket Number:	WTCY-0026-P07
Receipt Date:	29-MAR-2010
Filing Date:	28-DEC-2009
Time Stamp:	12:58:09
Application Type:	Utility under 35 USC 111(a)

Payment information:

Submitted with Payment	no

File Listing:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
1	Transmittal Letter	WTCY-0026-P07_032910_IDSCommunication.pdf	89711 8df4e7217656ae9c5e9111a2c9dfdfc66ddee837	no	2

Warnings:

Information:

2	Information Disclosure Statement (IDS) Filed (SB/08)	WTCY-0026-P07_032910_Form1449.pdf	129826 5b3b5ce5c5a22299a70a8f1850633cf9a9e9c413	no	1
Warnings:					
Information:					
This is not an USPTO supplied IDS fillable form					
3	NPL Documents	NPL-1_PCTUS0958499_ISR-121009.pdf	1734426 7b273cea5fbcd6d360a048143213c8c0fe1238f9	no	11
Warnings:					
Information:					
		Total Files Size (in bytes):	1953963		

This Acknowledgement Receipt evidences receipt on the noted date by the USPTO of the indicated documents, characterized by the applicant, and including page counts, where applicable. It serves as evidence of receipt similar to a Post Card, as described in MPEP 503.

<u>New Applications Under 35 U.S.C. 111</u>
If a new application is being filed and the application includes the necessary components for a filing date (see 37 CFR 1.53(b)-(d) and MPEP 506), a Filing Receipt (37 CFR 1.54) will be issued in due course and the date shown on this Acknowledgement Receipt will establish the filing date of the application.

<u>National Stage of an International Application under 35 U.S.C. 371</u>
If a timely submission to enter the national stage of an international application is compliant with the conditions of 35 U.S.C. 371 and other applicable requirements a Form PCT/DO/EO/903 indicating acceptance of the application as a national stage submission under 35 U.S.C. 371 will be issued in addition to the Filing Receipt, in due course.

<u>New International Application Filed with the USPTO as a Receiving Office</u>
If a new international application is being filed and the international application includes the necessary components for an international filing date (see PCT Article 11 and MPEP 1810), a Notification of the International Application Number and of the International Filing Date (Form PCT/RO/105) will be issued in due course, subject to prescriptions concerning national security, and the date shown on this Acknowledgement Receipt will establish the international filing date of the application.

ATTORNEY'S DOCKET NO. WTCY-0026-P07

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

Serial No.: 12/647,763
Filing Date: December 28, 2009
Applicant: Aristeidis Karalis et al.
Title: WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS
Group Art Unit: 2828
Examiner: Not Yet Known
Conf. No.: 2576

INFORMATION DISCLOSURE STATEMENT

Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

In compliance with the duty imposed by 37 C.F.R. § 1.56, and in accordance with 37 C.F.R. §§ 1.97 *et seq*., the referenced materials are brought to the attention of the Examiner for consideration in connection with the above-identified patent application. Applicants respectfully request that this Information Disclosure Statement be entered and the documents listed on the attached Form 1449 be considered by the Examiner and made of record. Pursuant to the provisions of MPEP 609, Applicants request that a copy of the 1449 form, initialed as being considered by the Examiner, be returned to the Applicants with the next official communication.

Pursuant to 37 C.F.R. §1.97(b), it is believed that no fee or statement is required with the Information Disclosure Statement. However, if an Office Action on the merits has been mailed, the Commissioner is hereby authorized to charge the required fees to Deposit Account No. 50-4262 in order to have this Information Disclosure Statement considered.

Pursuant to 37 C.F.R. 1.98(a)(2), Applicant believes that copies of cited U.S. Patents and Published and Non-Published Applications identifiable by USPTO Serial Number are no longer required to be provided to the Office. Notification of this change to this effect was provided in the United States Patent and Trademark Office OG Notices dated October 12, 2004 and October

19, 2004. Thus, Applicant has not included copies of any US Patents or US Patent Applications identifiable by serial number that may be cited with this submission. Should the Office require copies to be provided, Applicant respectfully requests that notice of such requirement be directed to Applicant's below-signed representative. Applicant acknowledges the requirement to submit copies of foreign patent documents and non-patent literature in accordance with 37 C.F.R. 1.98(a)(2).

The Examiner is invited to contact the Applicants' Representative at the below-listed telephone number if there are any questions regarding this communication.

Respectfully submitted on March 29, 2010,

STRATEGIC PATENTS, P.C.

/John Nortrup, Reg. No. 59,063/
John H. Nortrup
Telephone: 207-985-2126

Customer Number 43520

PTO/SB/06 (07-06)
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PATENT APPLICATION FEE DETERMINATION RECORD Substitute for Form PTO-875	Application or Docket Number 12/647,763	Filing Date 12/28/2009	☐ To be Mailed

APPLICATION AS FILED – PART I

	(Column 1)	(Column 2)	SMALL ENTITY ☐		OR	OTHER THAN SMALL ENTITY	
FOR	NUMBER FILED	NUMBER EXTRA	RATE ($)	FEE ($)		RATE ($)	FEE ($)
☒ BASIC FEE (37 CFR 1.16(a), (b), or (c))	N/A	N/A	N/A			N/A	
☒ SEARCH FEE (37 CFR 1.16(k), (i), or (m))	N/A	N/A	N/A			N/A	
☒ EXAMINATION FEE (37 CFR 1.16(o), (p), or (q))	N/A	N/A	N/A			N/A	
TOTAL CLAIMS (37 CFR 1.16(i))	14 minus 20 =	*	X $ =		OR	X $ =	
INDEPENDENT CLAIMS (37 CFR 1.16(h))	2 minus 3 =	*	X $ =			X $ =	
☐ APPLICATION SIZE FEE (37 CFR 1.16(s))	If the specification and drawings exceed 100 sheets of paper, the application size fee due is $250 ($125 for small entity) for each additional 50 sheets or fraction thereof. See 35 U.S.C. 41(a)(1)(G) and 37 CFR 1.16(s).						
☐ MULTIPLE DEPENDENT CLAIM PRESENT (37 CFR 1.16(j))							
			TOTAL			TOTAL	

* If the difference in column 1 is less than zero, enter "0" in column 2.

APPLICATION AS AMENDED – PART II

AMENDMENT	(Column 1)		(Column 2)	(Column 3)	SMALL ENTITY		OR	OTHER THAN SMALL ENTITY	
03/15/2010	CLAIMS REMAINING AFTER AMENDMENT		HIGHEST NUMBER PREVIOUSLY PAID FOR	PRESENT EXTRA	RATE ($)	ADDITIONAL FEE ($)		RATE ($)	ADDITIONAL FEE ($)
Total (37 CFR 1.16(i))	* 14	Minus	** 20	=	X $ =		OR	X $ =	
Independent (37 CFR 1.16(h))	* 2	Minus	***3	=	X $ =		OR	X $ =	
☐ Application Size Fee (37 CFR 1.16(s))									
☐ FIRST PRESENTATION OF MULTIPLE DEPENDENT CLAIM (37 CFR 1.16(j))							OR		
					TOTAL ADD'L FEE		OR	TOTAL ADD'L FEE	

AMENDMENT	(Column 1)		(Column 2)	(Column 3)					
03/29/2010	CLAIMS REMAINING AFTER AMENDMENT		HIGHEST NUMBER PREVIOUSLY PAID FOR	PRESENT EXTRA	RATE ($)	ADDITIONAL FEE ($)		RATE ($)	ADDITIONAL FEE ($)
Total (37 CFR 1.16(i))	* 14	Minus	** 20	=	X $ =		OR	X $ =	
Independent (37 CFR 1.16(h))	* 2	Minus	*** 3	=	X $ =		OR	X $ =	
☐ Application Size Fee (37 CFR 1.16(s))									
☐ FIRST PRESENTATION OF MULTIPLE DEPENDENT CLAIM (37 CFR 1.16(j))							OR		
					TOTAL ADD'L FEE		OR	TOTAL ADD'L FEE	

* If the entry in column 1 is less than the entry in column 2, write "0" in column 3.
** If the "Highest Number Previously Paid For" IN THIS SPACE is less than 20, enter "20".
*** If the "Highest Number Previously Paid For" IN THIS SPACE is less than 3, enter "3".
The "Highest Number Previously Paid For" (Total or Independent) is the highest number found in the appropriate box in column 1.

Legal Instrument Examiner:
/GLORIA NORRIS/

DEPARTMENT OF DEFENSE
ACCESS ACKNOWLEDGEMENT / SECRECY ORDER RECOMMENDATION
FOR PATENT APPLICATION

Application Serial No: DP12647763 Filing Date: Date Referred: 03/17/2010

I hereby acknowledge that the Department of Defense reviewers have inspected this application in administration of 35 USC 181 on behalf of the Agencies/Commands specified below. DoD reviewers will not divulge any information from this application for any purpose other than administration of 35 USC 181.

Defense Agency	Recommendation	Reviewer Name	Date Reviewed
Army	NC - Foreign Origin Application	Tammy Richmond	20 May 2010

Type of Recommendations:	*SNR: Secrecy Not Recommended* *SR: Secrecy Recommended* *NC: No Comment*

Instructions to Reviewers:

1. All DoD personnel reviewing this application will be listed on this form regardless of whether they are making a secrecy order recommendation.

2. This form will be forwarded to USPTO once all assigned DoD entities have provided their secrecy order recommendation.

Time for Completion of Review:

Pursuant to 35 USC 184, the subject matter of this application may be filed in a foreign country for the purpose of filing a patent application without a license anytime after the expiration of six (6) months from filing date unless the application becomes the subject of a secrecy order.

The USPTO publishes patent application at 18 months from the earliest claimed filing date. The USPTO will delay the publication of a patent application made available to a defense agency under 35 USC 181 until no earlier than 6 months from the filing date or 90 days from the date of referral to that agency. This application will be cleared for publication 6 months from the filing date or 90 days from the above Date Referred, whichever is later, unless a response is provided to the USPTO regarding the necessary recommendations as to the imposition of a secrecy order.

DoD Completion of Review: Final

Forwarded to USPTO: 05/20/2010 By: Joanne Stellato

Clear

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UNITED STATES DEPARTMENT OF COMMERCE
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www.uspto.gov

APPLICATION NUMBER	FILING OR 371(C) DATE	FIRST NAMED APPLICANT	ATTY. DOCKET NO./TITLE
12/647,763	12/28/2009	Aristeidis Karalis	WTCY-0026-P07

CONFIRMATION NO. 2576

NEW OR REVISED PPD NOTICE

OC000000044559657

43520
STRATEGIC PATENTS P.C..
C/O PORTFOLIOIP
P.O. BOX 52050
MINNEAPOLIS, MN 55402

NOTICE OF NEW OR REVISED PROJECTED PUBLICATION DATE

The above-identified application has a new or revised projected publication date. The current projected publication date for this application is 02/24/2011. If this is a new projected publication date (there was no previous projected publication date), the application has been cleared by Licensing & Review or a secrecy order has been rescinded and the application is now in the publication queue.

If this is a revised projected publication date (one that is different from a previously communicated projected publication date), the publication date has been revised due to processing delays in the USPTO or the abandonment and subsequent revival of an application. The application is anticipated to be published on a date that is more than six weeks different from the originally-projected publication date.

More detailed publication information is available through the private side of Patent Application Information Retrieval (PAIR) System. The direct link to access PAIR is currently http://pair.uspto.gov. Further assistance in electronically accessing the publication, or about PAIR, is available by calling the Patent Electronic Business Center at 1-866-217-9197.

Questions relating to this Notice should be directed to the Office of Data Management, Application Assistance Unit at (571) 272-4000, or (571) 272-4200, or 1-888-786-0101.

PART 1 - ATTORNEY/APPLICANT COPY

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Substitute for form 1449A/PTO

INFORMATION DISCLOSURE STATEMENT BY APPLICANT

(Use as many sheets as necessary)

Complete if Known

Application Number	12/647,763
Filing Date	Dec 28, 2009
First Named Inventor	Aristeidis Karalis
Art Unit	2828
Examiner Name	Not Yet Known

Sheet	1	of	13	Attorney Docket No: WTCY-0026-P07

US PATENT DOCUMENTS

Examiner Initial *	Cite No	Document Number	Publication Date	Name of Patentee or Applicant of Cited Document	Pages, Columns, Lines, Where Relevant Passages or RelevantFigures Appear
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		12/466065	Nov 19, 2009	Karalis, Aristeidis et al.	
		12/553957	Apr 29, 2010	Joannopoulos, John D., et al.	
		12/571949	Jun 17, 2010	Hamam, Rafif E., et al.	
		12/639958	Aug 19, 2010	Joannopoulos, John D.	
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		12/639966	May 20, 2010	Joannopoulos, John D.	
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		12/646442	Apr 22, 2010	Joannopoulos, John D.	
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		12/649777	Jun 3, 2010	Joannopoulos, John D.	
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		12/649904	Jul 29, 2010	Joannopoulos, John D.	
		12/649973	Apr 29, 2010	Joannopoulos, John D.	
		12/688305	May 13, 2010	Joannopoulos, John D.	
		12/688339	May 13, 2010	Karalis, Aristeidis	
		12/708850	Sep 23, 2010	Karalis, Aristeidis	
		12/713556	Sep 23, 2010	Karalis, Aristeidis	
		12/717559	Oct 7, 2010	Karalis, Aristeidis	
		2002/0032471	Mar 14, 2002	Loftin, Scott M., et al.	
		2002/0105343	Aug 8, 2002	Scheible, Guntram et al.	
		2002/0118004	Aug 29, 2002	Scheible, Guntram et al.	
		2003/0038641	Feb 27, 2003	Scheible, Guntram	
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		2003/0199778	Oct 23, 2003	Mickle, Marlin et al.	
		2003/0214255	Nov 20, 2003	Baarman, David W., et al.	
		2004/0000974	Jan 1, 2004	Odenaal, Willem G., et al.	
		2004/0113847	Jun 17, 2004	Qi,, Yihong et al.	

EXAMINER **DATE CONSIDERED**

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INFORMATION DISCLOSURE STATEMENT BY APPLICANT

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Complete if Known

Application Number	12/647,763
Filing Date	Dec 28, 2009
First Named Inventor	Aristeidis Karalis
Art Unit	2828
Examiner Name	Not Yet Known

Attorney Docket No: WTCY-0026-P07

US PATENT DOCUMENTS

Examiner Initial *	Cite No	Document Number	Publication Date	Name of Patentee or Applicant of Cited Document	Pages, Columns, Lines, Where Relevant Passages or RelevantFigures Appear
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		2004/0130916	Jul 8, 2004	Baarman, David W.	
		2004/0150934	Aug 5, 2004	Baarman, David W.	
		2004/0201361	Oct 14, 2004	Koh,, Won-Jun et al.	
		2004/0222751	Nov 11, 2004	Mollema, Scott A., et al.	
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		2004/0267501	Dec 30, 2004	Freed, Mason L., et al.	
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		2005/0116650	Jun 2, 2005	Baarman, David W.	
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		2007/0171681	Jul 26, 2007	Baarman, David W.	
		2007/0178945	Aug 2, 2007	Cook, Nigel P., et al.	
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		2008/0211320	Sep 4, 2008	Cook, Nigel P., et al.	

EXAMINER **DATE CONSIDERED**

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INFORMATION DISCLOSURE STATEMENT BY APPLICANT

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Complete if Known	
Application Number	12/647,763
Filing Date	Dec 28, 2009
First Named Inventor	Aristeidis Karalis
Art Unit	2828
Examiner Name	Not Yet Known

Attorney Docket No: WTCY-0026-P07

US PATENT DOCUMENTS

Examiner Initial *	Cite No	Document Number	Publication Date	Name of Patentee or Applicant of Cited Document	Pages, Columns, Lines, Where Relevant Passages or RelevantFigures Appear
		2008/0278264	Nov 13, 2008	Karalis, Aristeidis et al.	
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		2009/0015075	Jan 15, 2009	Cook, Nigel P., et al.	
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EXAMINER **DATE CONSIDERED**

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INFORMATION DISCLOSURE STATEMENT BY APPLICANT

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Application Number	12/647,763
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First Named Inventor	Aristeidis Karalis
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Examiner Name	Not Yet Known

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US PATENT DOCUMENTS

Examiner Initial *	Cite No	Document Number	Publication Date	Name of Patentee or Applicant of Cited Document	Pages, Columns, Lines, Where Relevant Passages or RelevantFigures Appear
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		5528113	Jun 18, 1996	Boys, John T., et al.	
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		6436299	Aug 20, 2002	Baarman, David W., et al.	
		6450946	Sep 17, 2002	Forsell, Peter	
		6452465	Sep 17, 2002	Brown, Andrew et al.	
		6515878	Feb 4, 2003	Meins, Jurgen G., et al.	
		6597076	Jul 22, 2003	Scheible, Guntram et al.	
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		6731071	May 4, 2004	Baarman, David W.	
		6749119	Jun 15, 2004	Scheible, Guntram et al.	
		6798716	Sep 28, 2004	Charych, Arthur	
		6806649	Oct 19, 2004	Mollema, Scott A., et al.	
		6812645	Nov 2, 2004	Baarman, David W.	
		6825620	Nov 30, 2004	Kuennen, Roy W., et al.	

EXAMINER **DATE CONSIDERED**

Substitute Disclosure Statement Form (PTO-1449)
* EXAMINER: Initial if reference considered, whether or not citation is in conformance with MPEP 609. Draw line through citation if not in conformance and not considered. Include copy of this form with next communication to applicant.1 Applicant's unique citation designation number (optional) 2 Applicant is to place a check mark here if English language Translation is attached

PTO/SB/08a (01-08)
Approved for use through 07/31/2012. OMB 0651-0031
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INFORMATION DISCLOSURE STATEMENT BY APPLICANT

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Application Number	12/647,763
Filing Date	Dec 28, 2009
First Named Inventor	Aristeidis Karalis
Art Unit	2828
Examiner Name	Not Yet Known

Sheet 5 of 13

Attorney Docket No: WTCY-0026-P07

US PATENT DOCUMENTS

Examiner Initial *	Cite No	Document Number	Publication Date	Name of Patentee or Applicant of Cited Document	Pages, Columns, Lines, Where Relevant Passages or RelevantFigures Appear
		6831417	Dec 14, 2004	Baarman, David W.	
		6844702	Jan 18, 2005	Giannopoulos, Demetri et al.	
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		6906495	Jun 14, 2005	Cheng, Lily K., et al.	
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		6975198	Dec 13, 2005	Baarman, David W., et al.	
		7042196	May 9, 2006	Cheng, Lily K., et al.	
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First Named Inventor	Aristeidis Karalis
Art Unit	2828
Examiner Name	Not Yet Known

Attorney Docket No: WTCY-0026-P07

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* EXAMINER: Initial if reference considered, whether or not citation is in conformance with MPEP 609. Draw line through citation if not in conformance and not considered. Include copy of this form with next communication to applicant.1 Applicant's unique citation designation number (optional) 2 Applicant is to place a check mark here if English language Translation is attached

Electronic Acknowledgement Receipt

EFS ID:	9207795
Application Number:	12647763
International Application Number:	
Confirmation Number:	2576
Title of Invention:	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS
First Named Inventor/Applicant Name:	Aristeidis Karalis
Customer Number:	43520
Filer:	John H. Nortrop/Elizabeth Nortrup
Filer Authorized By:	John H. Nortrop
Attorney Docket Number:	WTCY-0026-P07
Receipt Date:	11-JAN-2011
Filing Date:	28-DEC-2009
Time Stamp:	14:44:35
Application Type:	Utility under 35 USC 111(a)

Payment information:

Submitted with Payment	no

File Listing:

Document Number	Document Description	File Name	File Size(Bytes)/ Message Digest	Multi Part /.zip	Pages (if appl.)
1	Transmittal Letter	WTCY-0026-P07_011111_IDSCommunication.pdf	90248 a1dbe090cb271ffdc8d7138e3a95be6384a2f522	no	2

Warnings:

Information:

2	Information Disclosure Statement (IDS) Filed (SB/08)	WTCY-0026-P07_011111_Form1449.pdf	221365 8bfcc647717194ea50feec7ab94911775f0ed9b7	no	13

Warnings:

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New Applications Under 35 U.S.C. 111
If a new application is being filed and the application includes the necessary components for a filing date (see 37 CFR 1.53(b)-(d) and MPEP 506), a Filing Receipt (37 CFR 1.54) will be issued in due course and the date shown on this Acknowledgement Receipt will establish the filing date of the application.

National Stage of an International Application under 35 U.S.C. 371
If a timely submission to enter the national stage of an international application is compliant with the conditions of 35 U.S.C. 371 and other applicable requirements a Form PCT/DO/EO/903 indicating acceptance of the application as a national stage submission under 35 U.S.C. 371 will be issued in addition to the Filing Receipt, in due course.

New International Application Filed with the USPTO as a Receiving Office
If a new international application is being filed and the international application includes the necessary components for an international filing date (see PCT Article 11 and MPEP 1810), a Notification of the International Application Number and of the International Filing Date (Form PCT/RO/105) will be issued in due course, subject to prescriptions concerning national security, and the date shown on this Acknowledgement Receipt will establish the international filing date of the application.

ATTORNEY'S DOCKET NO. WTCY-0026-P07

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

Serial No.:	12/647,763
Filing Date:	December 28, 2009
Applicant:	Aristeidis Karalis et al.
Title:	WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS
Group Art Unit:	2828
Examiner:	Not Yet Known
Conf. No.:	2576

SUPPLEMENTAL INFORMATION DISCLOSURE STATEMENT

Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

In compliance with the duty imposed by 37 C.F.R. § 1.56, and in accordance with 37 C.F.R. §§ 1.97 *et seq*., the referenced materials are brought to the attention of the Examiner for consideration in connection with the above-identified patent application. Applicants respectfully request that this Supplemental Information Disclosure Statement be entered and the documents listed on the attached Form 1449 be considered by the Examiner and made of record. Pursuant to the provisions of MPEP 609, Applicants request that a copy of the 1449 form, initialed as being considered by the Examiner, be returned to the Applicants with the next official communication.

Pursuant to 37 C.F.R. §1.98(d), copies of the listed foreign and non-patent literature documents are not provided as these references were previously cited by or submitted to the U.S. Patent Office in connection with Applicants' prior U.S. application, Serial No. 12/567,716, filed on September 25, 2009, which is relied upon for an earlier filing date under 35 U.S.C. §120.

Pursuant to 37 C.F.R. §1.97(b), it is believed that no fee or statement is required with the Supplemental Information Disclosure Statement. However, if an Office Action on the merits has been mailed, the Commissioner is hereby authorized to charge the required fees to Deposit

Account No. 50-4262 in order to have this Supplemental Information Disclosure Statement considered.

Pursuant to 37 C.F.R. 1.98(a)(2), Applicant believes that copies of cited U.S. Patents and Published and Non-Published Applications identifiable by USPTO Serial Number are no longer required to be provided to the Office. Notification of this change to this effect was provided in the United States Patent and Trademark Office OG Notices dated October 12, 2004 and October 19, 2004. Thus, Applicant has not included copies of any US Patents or US Patent Applications identifiable by serial number that may be cited with this submission. Should the Office require copies to be provided, Applicant respectfully requests that notice of such requirement be directed to Applicant's below-signed representative. Applicant acknowledges the requirement to submit copies of foreign patent documents and non-patent literature in accordance with 37 C.F.R. 1.98(a)(2).

The Examiner is invited to contact the Applicants' Representative at the below-listed telephone number if there are any questions regarding this communication.

Respectfully submitted on January 11, 2011,

STRATEGIC PATENTS, P.C.

/John H. Nortrup, Reg. No. 59,063/
John H. Nortrup
Telephone: 207-985-2126

Customer Number 43520

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APPLICATION NUMBER	FILING OR 371(C) DATE	FIRST NAMED APPLICANT	ATTY. DOCKET NO./TITLE
12/647,763	12/28/2009	Aristeidis Karalis	WTCY-0026-P07

CONFIRMATION NO. 2576

PUBLICATION NOTICE

OC000000046166508

43520
STRATEGIC PATENTS P.C..
C/O CPA Global
P.O. BOX 52050
MINNEAPOLIS, MN 55402

Title:WIRELESS ENERGY TRANSFER USING FIELD SHAPING TO REDUCE LOSS

Publication No.US-2011-0043047-A1
Publication Date:02/24/2011

NOTICE OF PUBLICATION OF APPLICATION

The above-identified application will be electronically published as a patent application publication pursuant to 37 CFR 1.211, et seq. The patent application publication number and publication date are set forth above.

The publication may be accessed through the USPTO's publically available Searchable Databases via the Internet at www.uspto.gov. The direct link to access the publication is currently http://www.uspto.gov/patft/.

The publication process established by the Office does not provide for mailing a copy of the publication to applicant. A copy of the publication may be obtained from the Office upon payment of the appropriate fee set forth in 37 CFR 1.19(a)(1). Orders for copies of patent application publications are handled by the USPTO's Office of Public Records. The Office of Public Records can be reached by telephone at (703) 308-9726 or (800) 972-6382, by facsimile at (703) 305-8759, by mail addressed to the United States Patent and Trademark Office, Office of Public Records, Alexandria, VA 22313-1450 or via the Internet.

In addition, information on the status of the application, including the mailing date of Office actions and the dates of receipt of correspondence filed in the Office, may also be accessed via the Internet through the Patent Electronic Business Center at www.uspto.gov using the public side of the Patent Application Information and Retrieval (PAIR) system. The direct link to access this status information is currently http://pair.uspto.gov/. Prior to publication, such status information is confidential and may only be obtained by applicant using the private side of PAIR.

Further assistance in electronically accessing the publication, or about PAIR, is available by calling the Patent Electronic Business Center at 1-866-217-9197.

Office of Data Managment, Application Assistance Unit (571) 272-4000, or (571) 272-4200, or 1-888-786-0101

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INFORMATION DISCLOSURE STATEMENT BY APPLICANT

(Use as many sheets as necessary)

Complete if Known

Application Number	12/647,763
Filing Date	Dec 28, 2009
First Named Inventor	Aristeidis Karalis
Art Unit	2828
Examiner Name	Not Yet Known

Sheet	1	of	7

Attorney Docket No: WTCY-0026-P07

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Examiner Initial *	Cite No	Document Number	Publication Date	Name of Patentee or Applicant of Cited Document	Pages, Columns, Lines, Where Relevant Passages or RelevantFigures Appear
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		2005/0288740	Dec 29, 2005	Hassler, Jr., William L.	
		2007/0182367	Aug 9, 2007	Partovi, Afshin	
		2008/0067874	Mar 20, 2008	Tseng, Ryan	
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		2009/0108679	Apr 30, 2009	Porwal, Gunjan	
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		2009/0237194	Sep 24, 2009	Waffenschmidt, Eberhard et al.	
		2009/0251008	Oct 8, 2009	Sugaya, Shigeru	
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		2009/0271047	Oct 29, 2009	Wakamatsu, Masataka	
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		2009/0281678	Nov 12, 2009	Wakamatsu, Masataka	
		2009/0284082	Nov 19, 2009	Mohammadian, Alireza H.	
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		2009/0284218	Nov 19, 2009	Mohammadian, Alireza H., et al.	
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Application Number	12/647,763
Filing Date	Dec 28, 2009
First Named Inventor	Aristeidis Karalis
Art Unit	2828
Examiner Name	Not Yet Known

Sheet 2 of 7

Attorney Docket No: WTCY-0026-P07

US PATENT DOCUMENTS

Examiner Initial *	Cite No	Document Number	Publication Date	Name of Patentee or Applicant of Cited Document	Pages, Columns, Lines, Where Relevant Passages or RelevantFigures Appear
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INFORMATION DISCLOSURE STATEMENT BY APPLICANT

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Application Number	12/647,763
Filing Date	Dec 28, 2009
First Named Inventor	Aristeidis Karalis
Art Unit	2828
Examiner Name	Not Yet Known

Attorney Docket No: WTCY-0026-P07

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Application Number	12/647,763
Filing Date	Dec 28, 2009
First Named Inventor	Aristeidis Karalis
Art Unit	2828
Examiner Name	Not Yet Known

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US PATENT DOCUMENTS

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INFORMATION DISCLOSURE STATEMENT BY APPLICANT

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Application Number	12/647,763
Filing Date	Dec 28, 2009
First Named Inventor	Aristeidis Karalis
Art Unit	2828
Examiner Name	Not Yet Known

Sheet 5 of 7

Attorney Docket No: WTCY-0026-P07

US PATENT DOCUMENTS

Examiner Initial *	Cite No	Document Number	Publication Date	Name of Patentee or Applicant of Cited Document	Pages, Columns, Lines, Where Relevant Passages or RelevantFigures Appear
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First Named Inventor	Aristeidis Karalis
Art Unit	2828
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Sheet 6 of 7

Attorney Docket No: WTCY-0026-P07

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Examiner Initial *	Cite No	Document Number	Publication Date	Name of Patentee or Applicant of Cited Document	Pages, Columns, Lines, Where Relevant Passages or RelevantFigures Appear
		7639514	Dec 29, 2009	Baarman, David W.	
		7741734	Jun 22, 2010	Joannopoulos, John D., et al.	
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Examiner Initials*	Cite No [1]	Include name of the author (in CAPITAL LETTERS), title of the article (when appropriate), title of the item (book, magazine, journal, serial, symposium, catalog, etc.), date, page(s), volume-issue number(s), publisher, city and/or country where published.	T[2]
	NPL-104	, "Application Serial No. 12/613,686 (Atty Ref WTCY-0026-P02), Notice of Allowance mailed 01-06-11", 1/6/2011 , 10	
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Substitute Disclosure Statement Form (PTO-1449)
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OTHER DOCUMENTS -- NON PATENT LITERATURE DOCUMENTS			
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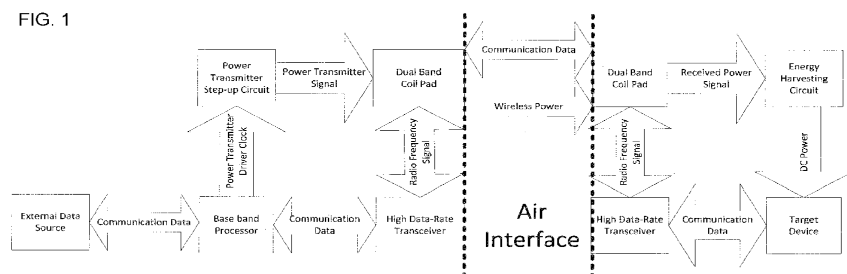
FIG. 1



(57) Abstract: In one aspect, the present disclosure describes a planar near-field wireless power charging system that is capable of charging small portable devices. Embodiments can incorporate coils for generating time-varying magnetic fields into a pad. In an exemplary embodiment, the charging system incorporates a charging pad that can act as an electrically small coil antenna for the low frequencies and long wavelengths, used for charging, and long wavelengths, and which can also be used for communication purposes by treating it as an electrically large antenna at higher frequencies, and shorter wavelengths, used for communications. In an exemplary embodiment, the system uses multiple lower powered transmitters, where each transmitter feeds a separate coil. The separate coils can be stacked so that the magnetic fields are substantially coextensive. The simultaneous driving of the multiple coils by the multiple transmitters can achieve similar power delivery as a single high powered transmitter. Multiple stacked sets of coils can be integrated into a pad such that each stacked set of coils provides vertical magnetic fields over a section of the pad. Embodiments of the subject invention can be designed to couple energy to the receiver coil of the device via magnetic fields that have a substantial vertical component. An embodiment of the present disclosure describes a receiver coil attached to a portable electronic device with a mechanical connection that allows the receiver coil to be positioned such that the vertical fields do not need to pass through a substantial portion of the device to pass through the receiver coil during charging and can allow the receiver coil to be conveniently positioned adjacent the device when not changing.

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DESCRIPTION

PLANAR NEAR-FIELD WIRELESS POWER CHARGER AND HIGH-SPEED DATA COMMUNICATION PLATFORM

CROSS-REFERENCE TO RELATED APPLICATION

The present application claims the benefit of U.S. Provisional Application Serial No. 60/970,201, filed September 5, 2007, which is hereby incorporated by reference herein in its entirety, including any figures, tables, or drawings.

BACKGROUND OF INVENTION

Portable electronic equipment, such as mobile phones, handheld computers, and personal data assistants, are normally powered by batteries. In many cases, rechargeable batteries are preferred because of environmental and economical concerns. The most common way to charge rechargeable batteries is to use a conventional charger, which normally includes an AC-DC power supply when the AC mains are used, or a DC-DC power supply when a car battery is used. Conventional chargers normally use an electric cable to connect the charger circuit to the battery located in the portable electronic equipment.

In large part due to the inconvenience of conventional chargers, wireless power solutions have been presented. Examples of wireless power solutions include electric toothbrushes and electric razors. Current wireless power solutions have limitations such as requiring a particular alignment of the device being charged to the charging unit, or other inconveniences such as the need for the device being charged to be within a certain distance of a certain section of the charging unit.

Thus, there remains a need for a system for wireless charging of portable electronic devices in an efficient manner.

BRIEF SUMMARY

In one aspect, the present disclosure describes a planar near-field wireless power charging system that is capable of charging small portable devices. Embodiments can incorporate coils for generating time-varying magnetic fields into a pad. In an exemplary embodiment, the charging system incorporates a charging pad that can act as an electrically small coil antenna for the low frequencies and long wavelengths, used for charging, and long wavelengths, and which can also be used for communication purposes by treating it as an electrically large antenna at higher frequencies, and shorter wavelengths, used for communications.

In an exemplary embodiment, the system uses multiple lower powered transmitters, where each transmitter feeds a separate coil. The separate coils can be stacked so that the magnetic fields are substantially coextensive. The simultaneous driving of the multiple coils by the multiple transmitters can achieve similar power delivery as a single high powered transmitter. Multiple stacked sets of coils can be integrated into a pad such that each stacked set of coils provides vertical magnetic fields over a section of the pad. Embodiments of the subject invention can be designed to couple energy to the receiver coil of the device via magnetic fields that have a substantial vertical component.

An embodiment of the present disclosure describes a receiver coil attached to a portable electronic device with a mechanical connection that allows the receiver coil to be positioned such that the vertical fields do not need to pass through a substantial portion of the device to pass through the receiver coil during charging and can allow the receiver coil to be conveniently positioned adjacent the device when not charging. In exemplary embodiments, the mechanical connection can be a flip or slide mechanism that allows the receiver coil to be positioned such that a vertical magnetic field does not need to pass through the portable device in order to pass through the receiver coil during charging, but allows the receiver coil to reside adjacent to the device when not charging in order to make the device with receiver coil easier to carry and store.

BRIEF DESCRIPTION OF DRAWINGS

Figure 1 shows one embodiment of a system architecture of a planar near-field wireless power charger and high-speed data communication platform.

Figure 2 shows an embodiment of a planar wireless power charger platform in accordance with the present disclosure.

Figure 3 shows a diagram of one embodiment of a multiple-transmitter power delivery system.

Figure 4 shows a system block diagram of a multiple-transmitter power delivery system.

Figures 5A-5B show embodiments of a single transmitter coil for a wireless power system.

Figure 6 shows an existing technology using a horizontal field to charge the portable device.

Figure 7 shows a cell phone incorporating an embodiment of a power charging receiver in accordance with the present disclosure.

Figures 8A-8B show a power charging receiver attached to a cell phone using a flip mechanism.

Figure 9 shows a power charging receiver attached to a cell phone using the a slide mechanism.

DETAILED DISCLOSURE

In one aspect, the present disclosure describes a planar near-field wireless power charging system that is capable of charging small portable devices. Embodiments can incorporate coils for generating time-varying magnetic fields into a pad. Embodiments of the wireless power charging system can allow powering devices at close proximity. In an embodiment, a charging efficiency of greater than 75% can be achieved. In a further embodiment, a charging efficiency of greater than 85% can be achieved. Specific embodiments can allow charging to occur at a distance of, for example, up to about 5 inches above the pad. The ability to charge the device with up to a 5 inch separation between the device and the charging unit, or portion thereof, allows versatility as to where devices can be positioned during the charging process. In a specific embodiment, a receiver coil as small as 10% of the transmitting coil can be utilized. In an exemplary embodiment, the system is capable of charging multiple devices at the same time.

In an exemplary embodiment, the charging system incorporates a charging pad that can act as an electrically small coil antenna for the low frequencies and long wavelengths, used for charging, and long wavelengths, and which can also be used for communication purposes by treating it as an electrically large antenna at higher frequencies, and shorter wavelengths, used for communications. In a specific embodiment, a power signal for charging can have a frequency less than or equal to 1MHz. in a further embodiment, the data signal for communications can have a frequency in the ultra wide band (UWB) of 3.1GHz-10.6GHz and use a carrier-less signal with pulses, or a carrier based signal with a carrier frequency in the WiFi band, e.g., at 2.4GHz.

Figure 1 shows one embodiment of a system architecture of a planar near-field wireless power charger and high-speed data communication platform. The system architecture can include a power transmitter, power receiver and energy harvesting circuit, and transmitting coil pad. In an exemplary embodiment, the subject platform is designed for short range applications using near field coupling. This embodiment can achieve high data rates using a simple transceiver architecture. In various embodiments, the base station can: transmit a power signal; transmit a power signal and transmit a data signal; transmit a power signal, transmit a data signal, and receive a data signal; or transmit a power signal and receive a data signal. The base station can break the various transmit and/or receive functions into separate transmitters and receivers, or combine such functions into one or more transceivers. By using a single coil to transmit and receive costs and space can be saved. Likewise, the receiver can: receive a power signal; receive a power signal and transmit a data signal; receive a power signal, transmit a data signal, and receive a data signal; or receive a power signal and transmit a data signal. The transmit power, transmit data, receive power, and receive data functions can be performed sequentially or simultaneously.

The wireless transfer of power and data can allow a variety of configurations. An embodiment can use a first pad and a second pad, each having a base station, and a television having a receiver, with the TV on the first pad and a DVD player having a receiver on a second pad. Power is provided to the DVD player and data can be received by the second pad from the DVD player. The received data can then be transferred from the second pad to the first pad and then from the data transmitter in the first pad to the receiver in the TV. Other embodiments can use a pad and an audio device, such as and MP3 player, where power is supplied to the audio device and music data is received by the pad from the audio device. The received audio data can then be delivered to, for example, speakers or some sort of device to further utilize the audio data. The pad can be networked to a home or office network, such that received data can be distributed as needed. The data can also include information about the charging state of the receiver device. In a specific embodiment, the data can be transmitted at a rate of 50Mb-500Mb. In another specific embodiment, the data can be transmitted at a rate up to 1Gbps.

Figure 2 shows an embodiment of a platform in accordance with the present disclosure. Exemplary applications of this system include its use as a platform for wireless power charging and high data rate communication for portable devices, such as mobile phones and mp3 players. The system can be implemented in various locations, such as homes, airports, and hotel rooms. This adds convenience as it provides a universal charging interface and can reduce, or eliminate, the need to bring multiple chargers. In addition, it can serve as a high speed data link between devices so that data can be exchanged on the same platform, instead of separate ones. This can be integrated with a smart home or smart office space. The system can have simple hardware architecture so as to achieve low cost for mass production. In one embodiment, a wireless near-field high-speed data communication capability is implemented as part of the system.

Preferably, the magnetic flux generated by the transmitting coil pad is substantially uniform over at least a major part of the planar charging surface. In this way, the impact that the precise position and orientation of the electronic device on the charging surface has on the charging efficiency is reduced. The system illustrated in Figure 2 includes a power transmitter **4** that powers a power-transmitting coil pad **6**. The power transmitter **4** is powered by power source **2**. The transmitting coil pad **6** wirelessly sends power to a power receiver and energy harvesting circuit **8**. The power receiver is connected to the target electronic device **10**. In alternative embodiments, the power receiver and energy harvesting circuit can be attached or integrated with the target device, such that the target device can be placed on or near the coil pad for charging. In an exemplary embodiment, discussed later, the power receiver is formed as a back cover of the electronic device. In an exemplary embodiment, the equipment is charged simply by placing the power receiver on, or near, the transmitting coil pad surface.

In an exemplary embodiment, the system uses multiple lower powered transmitters, where each transmitter feeds a separate coil. The separate coils can be stacked so that the magnetic fields are substantially coextensive. The simultaneous driving of the multiple coils by the multiple transmitters can achieve similar power delivery as a single high powered transmitter. In a specific embodiment, the size of each lower powered transmitter can be reduced up to 80-90% compared with a single high-powered transmitted, such that the large transformer typically used to step up the voltage to a high voltage can be eliminated from the system. Removing the large transformer can enhance safety. Figure 3 shows a diagram of one embodiment incorporating a multiple-transmitter power delivery system for a multiple coil design having seven coils. Multiple stacked sets of coils can be integrated into a pad such that each stacked set of coils provides vertical magnetic fields over a section of the pad. Figure 4 shows a system block diagram of a multiple-transmitter power delivery system that can be used with a coil design having eight coils.

Figures 5A -5B show embodiments of a single transmitter coil for a wireless power system. The coils can have a variety of shapes. This coil can be integrated with a pad, or other object, for placing devices to be charged on or near the pad. In an embodiment, the coils are positioned such that the resulting magnetic field is normal to a surface of the pad, such that receiver coils can be located on the pad's surface to receive the magnetic fields. In a specific embodiment, the transmitter coil, such as the coil shown in Figure 5, can be driven by multiple transmitters, which can each be selectively turned on or off, and tuned. In a specific embodiment incorporating a stacked set of coils, each transmitter can be tuned independently, thereby achieving a wide tuning range. Individual tuning of the coils can be performed in near real time and increase energy transfer efficiency for different charging circumstances. Power control can be achieved by individually turning on and off the individual transmitters depending on the charging load, making it more energy efficient. The multiple-transmitter design enables the transmitter system to be more flexible and adaptive to a wider range of environmental situations while consuming less power. Thus, a significant improvement in efficiency can be attained. In a specific embodiment, each transmitter for the stacked set of coils is identical, or each transmitter for the multiple transmitters driving a single coil, allowing low cost for mass production. Since the transmitters are the same, additional transmitters can be easily added to the system to boost the maximum output, thereby offering a great advantage for scalability. Having multiple transmitters also gives the system many extra degrees of freedom for achieving a higher dynamic range. In a specific embodiment, the coils in the stacked configurations are driven in place by the plurality of transmitters.

Referring to Figure 6, a device incorporating a coil design to receive power from surface magnetic fields that are substantially horizontal is shown. In accordance with current wireless power charging technology that uses horizontal magnetic fields for coupling of energy to charge portable electronic devices. The receiver coil used with the device to be charged receives magnetic field flux that is substantially parallel with the surface of the pad the device is placed on and does not extend very far above the surface of the pad. Such use of horizontal magnetic fields is inefficient because the cross sectional area of the receiver coil is limited; thus, the coupling factor is low. Embodiments of the subject invention can be designed to couple energy to the receiver coil of the device via magnetic fields that have a substantial vertical component. Examples of transmitter coil design that can be used to produce magnetic fields that have a substantial vertical component are shown in Figure 3 and Figure 5. Other designs can also be used in accordance with the invention.

Using a vertical field for coupling of energy can be more efficient. However, by mounting the receiver on the back of a portable device, the efficiency of such coupling can be reduced as the magnetic field must pass through the portable device in order to pass through the receiver coil. An embodiment of the present disclosure describes a receiver coil attached to a portable electronic device with a mechanical connection that allows the receiver coil to be positioned such that the vertical fields do not need to pass through a substantial portion of the device to pass through the receiver coil during charging and can allow the receiver coil to be conveniently positioned adjacent the device when not charging. In specific embodiments, the receiver coil can be positioned so that the magnetic fields pass through less than one half of the body of the device. In exemplary embodiments, the mechanical connection can be a flip or slide mechanism that allows the receiver coil to be positioned such that a vertical magnetic field does not need to pass through the portable device in order to pass through the receiver coil during charging, but allows the receiver coil to reside adjacent to the device when not charging in order to make the device with receiver coil easier to carry and store. The use of such a mechanical connection can enable efficient charging of the device without significant modification to the device's form factor, as shown in Figure 7.

The receiver can be designed as an add-on device or as part of a portable electronic device such as a mobile phone. In an exemplary application, a user need not replace an entire phone to take advantage of the new technology. For example, the user can replace the back panel or the battery cover of a phone or device to make the device compliant to this technology.

Figures 8A and 8B show a power charging receiver coil attached to a cell phone using a flip mechanism. In one embodiment, the flip mechanism includes a catch at one end of the device and a hinge on the other end of the device for allowing the receiver coil to flip open. The receiver coil can flip out, as shown by arrows in Figure 8B, such that the receiver coil is rotated 180 degrees and is parallel with the surface of the device from where it rotated from. In an alternative embodiment, the receiver coil can rotate 90 degrees and be used to receive horizontal magnetic fields for charging the device. Figure 9 shows a power charging receiver attached to a cell phone using a slide mechanism. In this embodiment, the receiver coil is slid out away from the phone body and placed on the charging pad so that the receiver coil is substantially parallel with the surface of the charging pad so as to allow the vertical portion of the magnetic field to pass through the receiver coil. Other mechanisms can be incorporated with embodiments of the invention to coordinate the relative position of the device body and the receiver coil.

The large cross-sectional area of the receiver coil provides for increased coupling for use with a near-field planar wireless power charger system. Positioning the receiver coil, for example, via the flip mechanism or the slide mechanism, such that the magnetic field flux does not pass through the body of the device enhances the charging efficiency. The device can be charged in the pre-flip or pre-slide position, but at a slower rate. To fully realize the benefits of the subject receiver coil design, the wireless power charger system can incorporate a vertical magnetic field for coupling of energy to the receiver coil. Such a vertical magnetic field achieves a higher efficiency coupling and, therefore, higher efficiency charging. Further, embodiments using vertical fields can allow the receiver coil to be positioned higher off of the pad having the transmitted coils than with systems using horizontal fields, while achieving the same coupling efficiency. In a specific embodiment, the receiver coil can be up to 5 inches above the pad.

All patents, patent applications, provisional applications, and publications referred to or cited herein are incorporated by reference in their entirety, including all figures and tables, to the extent they are not inconsistent with the explicit teachings of this specification.

It should be understood that the examples and embodiments described herein are for illustrative purposes only and that various modifications or changes in light thereof will be suggested to persons skilled in the art and are to be included within the spirit and purview of this application.

CLAIMS

1. A wireless power transfer system, comprising:

a base station, wherein the base station comprises:

a transmitter coil;

a power transmitter, wherein the power transmitter is capable of driving the transmitter coil such that a first magnetic field is produced; and

a data receiver,

a receiver station, wherein the receiver station comprises:

a receiver coil, wherein the receiver coil inductively couples the first magnetic field to produce a received power signal; and

a data transmitter, wherein the data transmitter drives the receiver coil to produce a transmitted data signal,

wherein the transmitter coil inductively couples the transmitted data signal to produce a received RF data signal, wherein the data receiver receives the received RF data signal and outputs a received data signal.

2. The wireless power transfer system according to claim 1, where in the receiver station further comprises an energy harvesting circuit, wherein the energy harvesting circuit receives the received power signal and outputs DC power.

3. The wireless power transfer system according to claim 1, wherein the receiver station further comprises a target device, wherein the target device provides a data input signal to the data transmitter, wherein the data input signal is used by the data transmitter to produce the data signal.

4. The wireless power transfer system according to claim 1, wherein the base station further comprises:

a second data transmitter, wherein the second data transmitter drives the transmitter coil to produce a second transmitted data signal, wherein the receiver station further comprises a second data receiver, wherein the receiver coil inductively couples the second transmitted data signal to produce a second received RF data signal.

5. The wireless power transfer system according to claim 1, wherein the power transmitter and the data receiver are integrated as a first transceiver.

6. The wireless power transfer system according to claim 4, wherein the data transmitter and the second data receiver are integrated as a second transceiver.

7. The wireless power transfer system according to claim 4, wherein the second data transmitter is integrated with the power transmitter and the data receiver as the first transceiver.

8. The wireless power transfer system according to claim 4, wherein the second data receiver is integrated with the data transmitter and the second data receiver as the second transceiver.

9. The wireless power transfer system according to claim 4, wherein the power transmitter and the second data transmitter are capable of simultaneously driving the transmitter coil, wherein the power transmitter drives the transmitter coil at a first wavelength and the second data transmitter drives the transmitter coil at a second wavelength.

10. A wireless power transfer system, comprising:

a base station, wherein the base station comprises:

a transmitter coil;

a power transmitter, wherein the power transmitter is capable of driving the transmitter coil such that a first magnetic field is produced; and

a data transmitter,

a receiver station, wherein the receiver station comprises:

a receiver coil, wherein the receiver coil inductively couples the first magnetic field to produce a received power signal; and

a data receiver, wherein the data transmitter drives the transmitter coil to produce a transmitted data signal,

wherein the reciever coil inductively couples the transmitted data signal to produce a received RF data signal, wherein the data receiver receives the received RF data signal and outputs a received data signal.

11. A method of wireless power transfer, comprising:

providing a base station, wherein the base station comprises:

a transmitter coil;

a power transmitter; and

a data receiver,

providing a receiver station, wherein the receiver station comprises:

a receiver coil; and

a data transmitter,

driving the transmitter coil with the power transmitter such that a first magnetic field is produced;

inductively coupling the first magnetic field to the receiver coil to produce a received power signal;

driving the receiver coil with the data transmitter to produce the transmitted data signal;

inductively coupling the transmitted data signal to the transmitter coil to produce a received RF data signal;

receiving the received RF data signal by the data receiver; and

outputting a received data signal by the data receiver.

12. The method according to claim 11, where in the receiver station further comprises an energy harvesting circuit, further comprising receiving the received power signal by the energy harvesting circuit and outputting DC power by the energy harvesting circuit.

13. The method according to claim 11, wherein the receiver station further comprises a target device, further comprising providing a data input signal from the target device to the data transmitter, wherein the data input signal is used by the data transmitter to produce the data signal.

14. The method according to claim 11, wherein the base station further comprises:

a second data transmitter and the receiver station further comprises

a second data receiver, the method further comprising:

driving the transmitter coil with the second data transmitter to produce a second transmitted data signal; and

inductively coupling the second transmitted data signal to the receiver coil to produce a second received RF data signal.

15. The method according to claim 11, wherein the power transmitter and the data receiver are integrated as a first transceiver.

16. The method according to claim 14, wherein the data transmitter and the second data receiver are integrated as a second transceiver.

17. The method according to claim 14, wherein the second data transmitter is integrated with the power transmitter and the data receiver as the first transceiver.

18. The method according to claim 14, wherein the second data receiver is integrated with the data transmitter and the second data receiver as the second transceiver.

19. The method according to claim 14, wherein the power transmitter and the second data transmitter are capable of simultaneously driving the transmitter coil, wherein the power transmitter drives the transmitter coil at a first wavelength and the second data transmitter drives the transmitter coil at a second wavelength.

20. A method of wireless power transfer, comprising:

providing a base station, wherein the base station comprises:

a transmitter coil;

a power transmitter; and

a data transmitter,

providing a receiver station, wherein the receiver station comprises:

a receiver coil; and

a data receiver,

driving the transmitter coil with the power transmitter such that a first magnetic field is produced;

inductively coupling the first magnetic field to the receiver coil to produce a received power signal;

driving the transmitter coil with the data transmitter to produce the transmitted data signal;

inductively coupling the transmitted data signal to the receiver coil to produce a received RF data signal;

receiving the received RF data signal by the data receiver; and

outputting a received data signal by the data receiver.

21. A wireless power transfer system, comprising:

a plurality of transmitters;

a corresponding plurality of transmitter coils, wherein each transmitter delivers power to a corresponding transmitter coil, so as to produce a corresponding plurality of magnetic fields, wherein the plurality of transmitter coils are positioned such that corresponding plurality of magnetic fields are substantially coextensive.

22. The system according to claim 21, wherein the plurality of transmitter coils are positioned such that the plurality of magnetic fields are normal to a surface.

23. The system according to claim 22, further comprising at least one receiver coil, wherein the at least one receiver coil is positionable with respect to the surface such that the at least one receiver coil is substantially parallel with the surface.

24. The system according to claim 22, wherein the plurality of transmitter coils are integrated in a pad, wherein the surface is on the pad.

25. The system according to claim 21, wherein each of the plurality of transmitters can be tuned independently.

26. The system according to claim 21, wherein the plurality of transmitters can be individually turned on and off depending on a needed charging load.

27. The system according to claim 21, wherein the plurality of transmitters drive the plurality of transmitter coils in phase.

28. An electronic device, comprising:

a body;

a receiver coil; and

a connector, wherein the connector moveably connects the receiver coil to the body such that the receiver coil can transition between at least two positions relative to the body, wherein at least one of the at least two positions the receiver coil can receive magnetic fields normal to the receiver coil without the received magnetic fields passing through a substantial portion of the body.

29. The device according to claim 28, wherein the connector allows the receiver coil to slide relative to the body to transition between the at least two positions.

30. The device according to claim 28, wherein the connector allows the receiver coil to rotate relative to the body to transition between the at least two positions.

31. The device according to claim 28, wherein in at least one other of the at least two positions the receiver coil is adjacent the body such that magnetic fields normal to the receiver coil pass through a substantial portion of the body.

32. The device according to claim 31, wherein the connector comprises a hinge locateded at an end of the body, wherein the hinge allows the receiver coil to rotate from a first position adjacent the body to a second position that is an approximately 180° rotation of the receive coil with to the hinge.

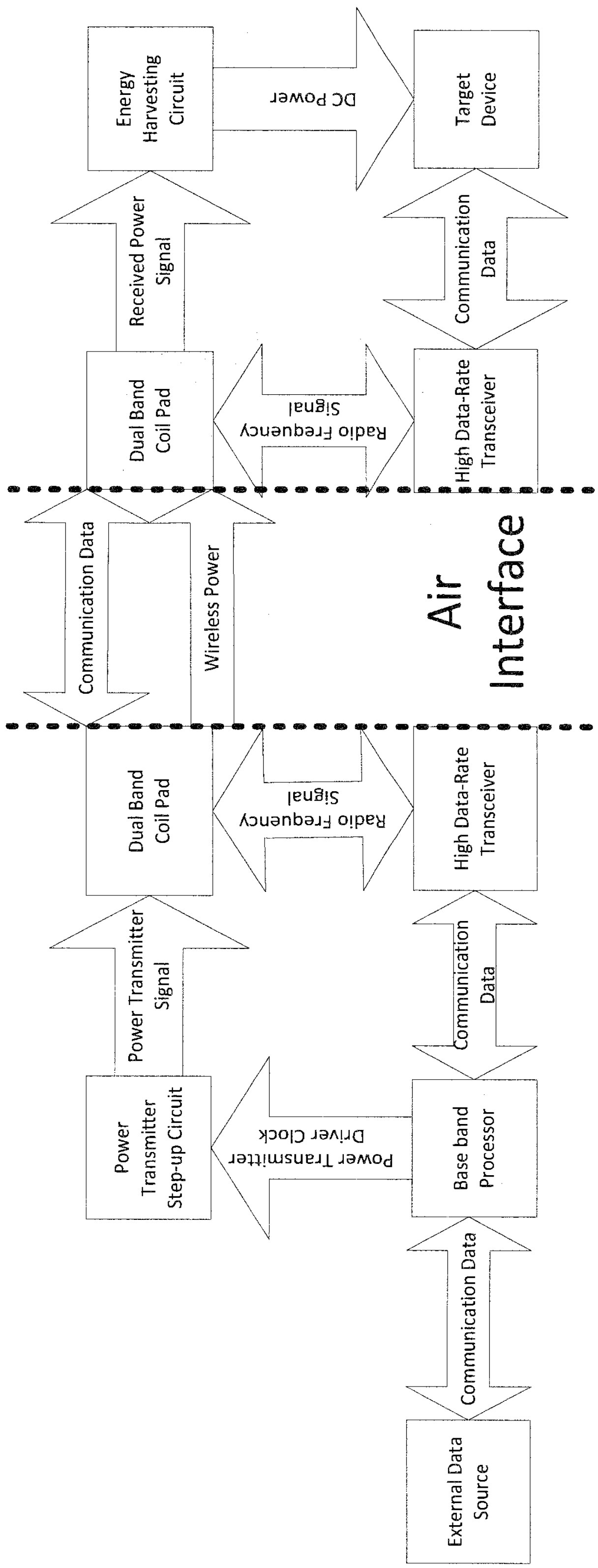
External Data Source
Communication Data
Base band Processor
Power Transmitter Driver Clock
Power Transmitter Step-up Circuit
Power Transmitter Signal
Communication Data
Dual Band Coil Pad
Radio Frequency Signal
High Data-Rate Transceiver
Communication Data
Wireless Power
Air Interface
Dual Band Coil Pad
Radio Frequency Signal
High Data-Rate Transceiver
Received Power Signal
Communication Data
Energy Harvesting Circuit
DC Power
Target Device

FIG. 1

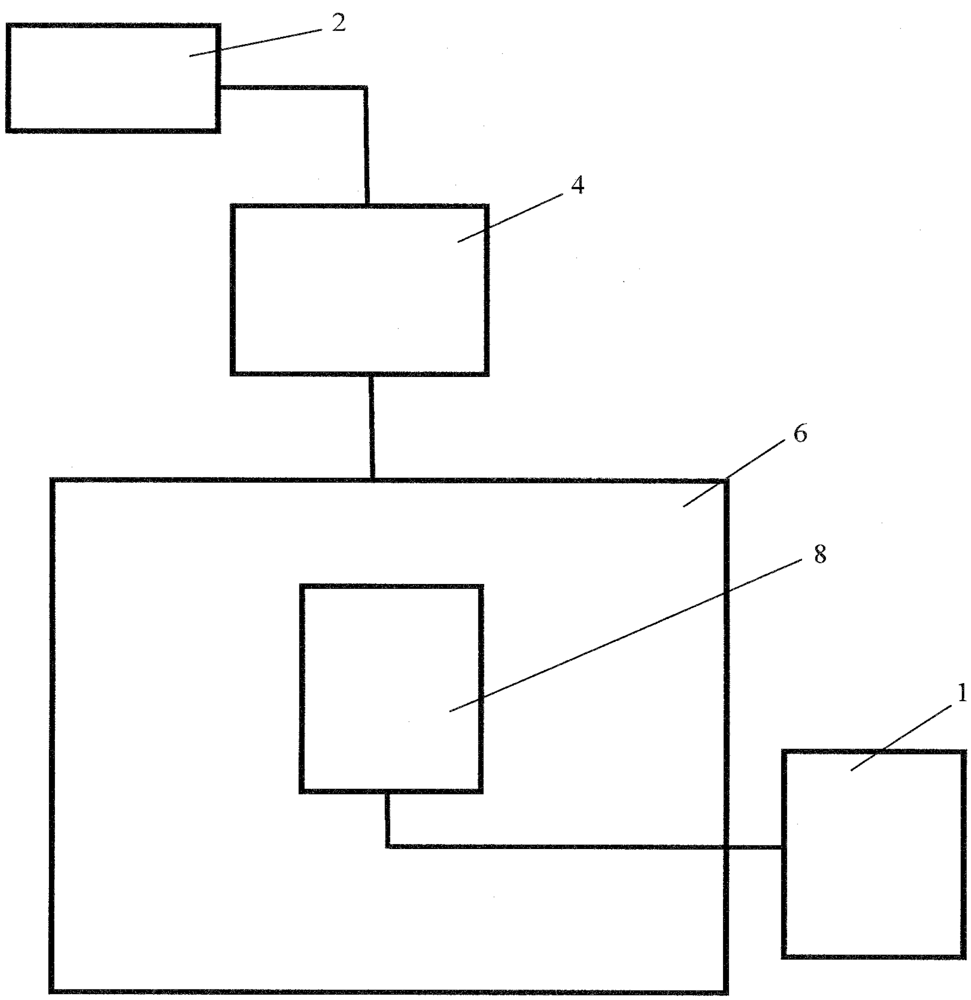

FIG. 2

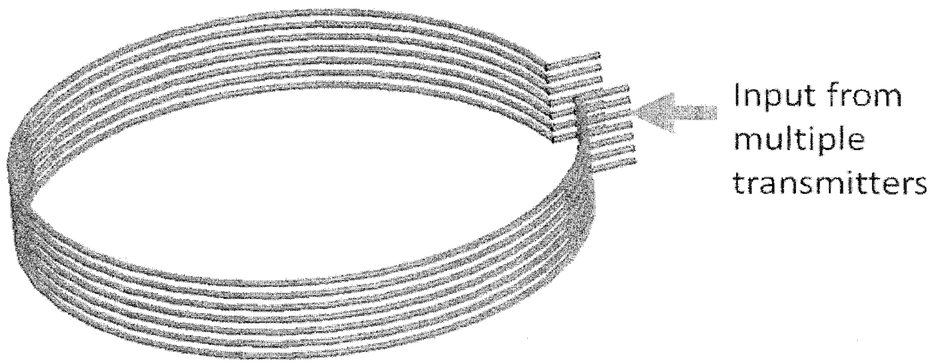
Input from
multiple
transmitters

FIG. 3

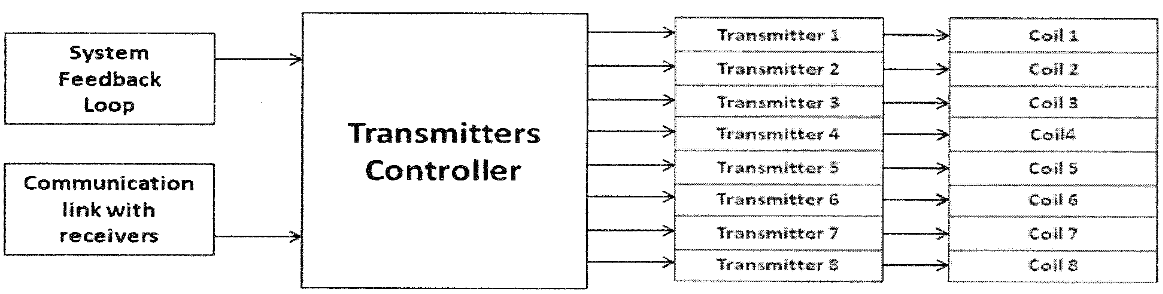
System
Feedback
Loop
Communication
link with
receivers
Transmitters
Controller
Transmitter 1
Transmitter 2
Transmitter 3
Transmitter 4
Transmitter 5
Transmitter 6
Transmitter 7
Transmitter 8
Coil 1
Coil 2
Coil 3
Coil4
Coil 5
Coil 6
Coil 7
Coil 8

FIG. 4

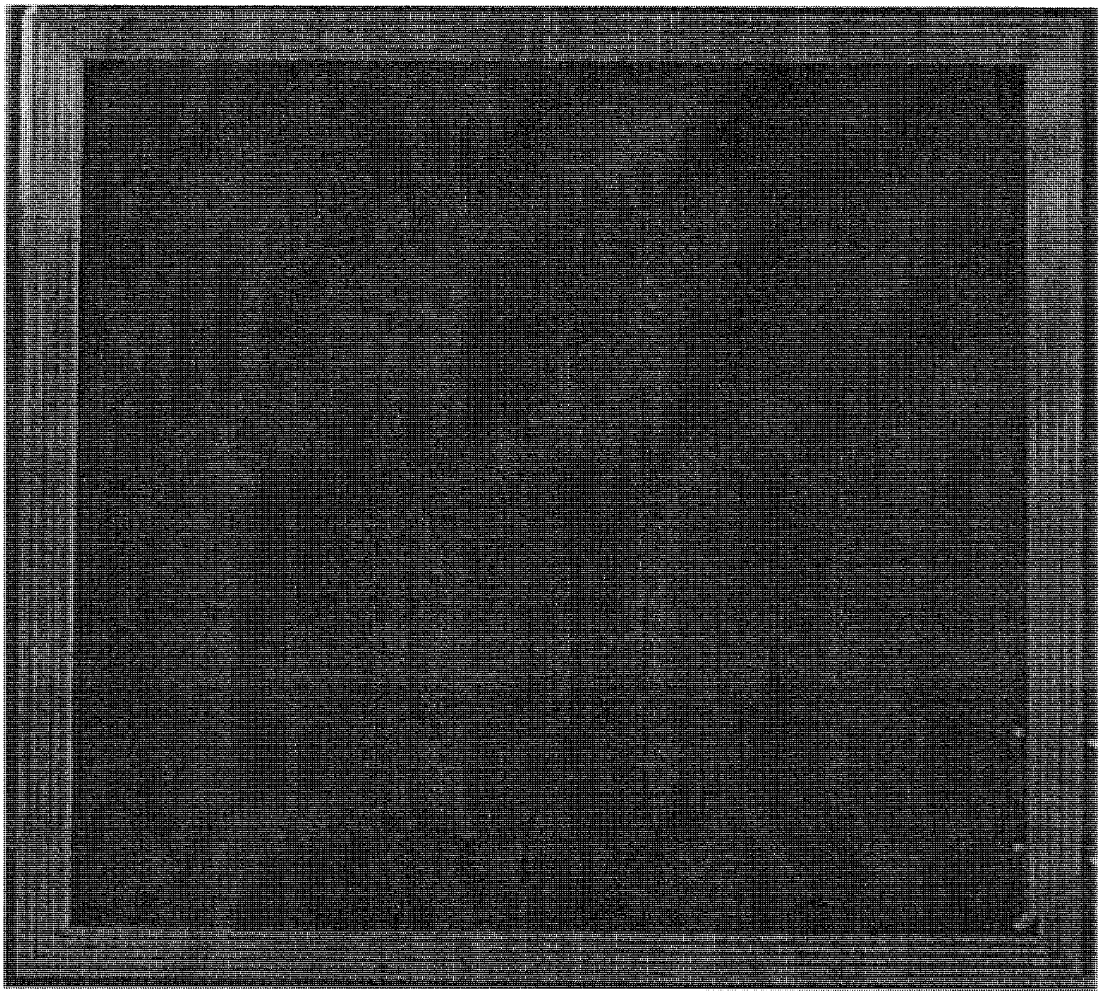

FIG. 5A

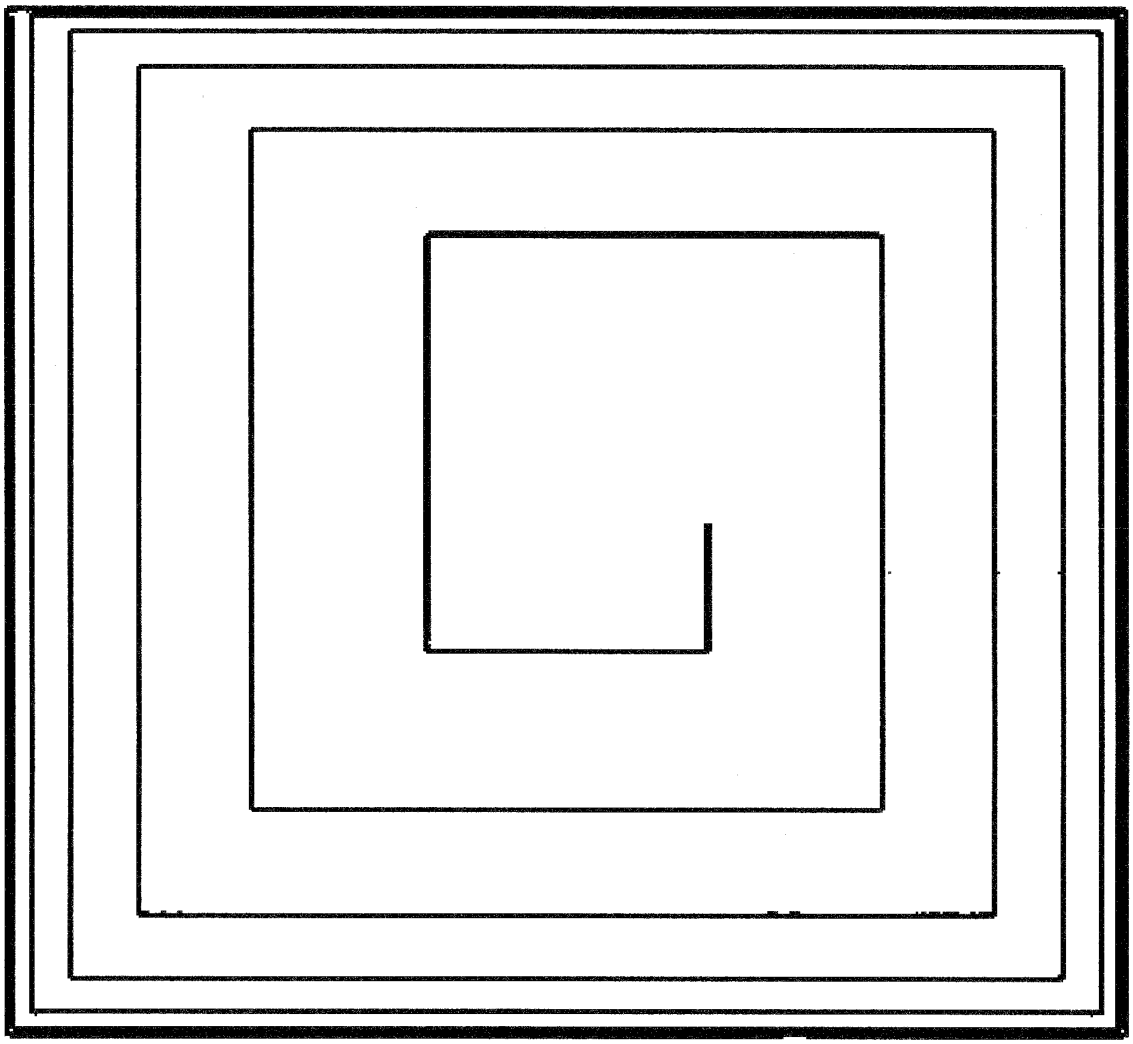

FIG. 5B

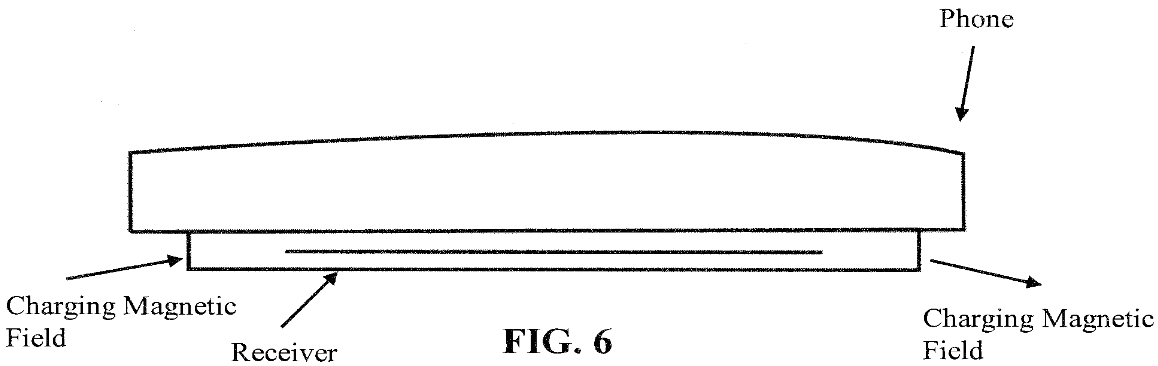

FIG. 6

(Prior Art)

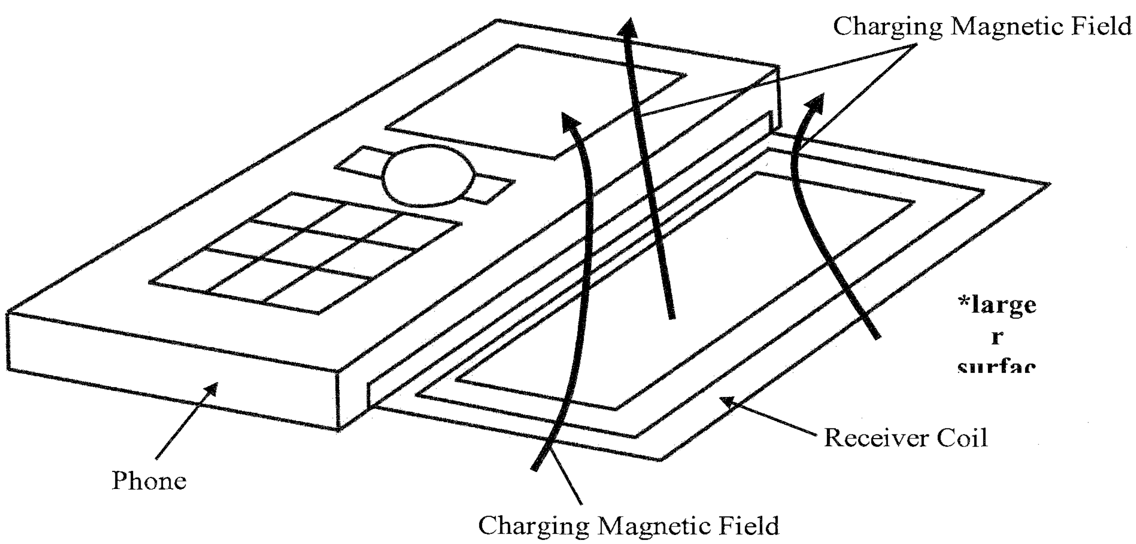

FIG. 7

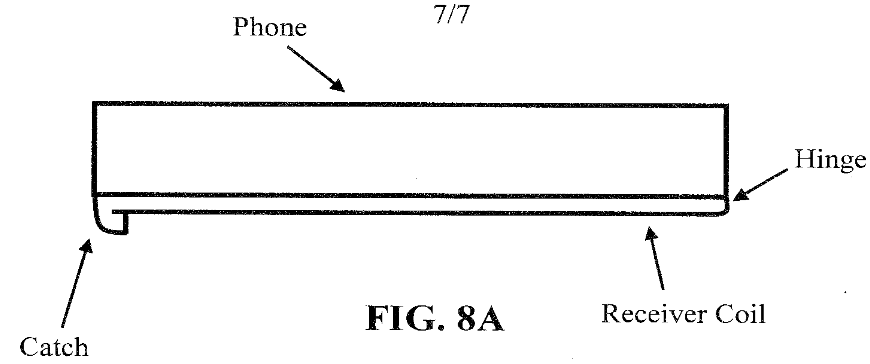

FIG. 8A

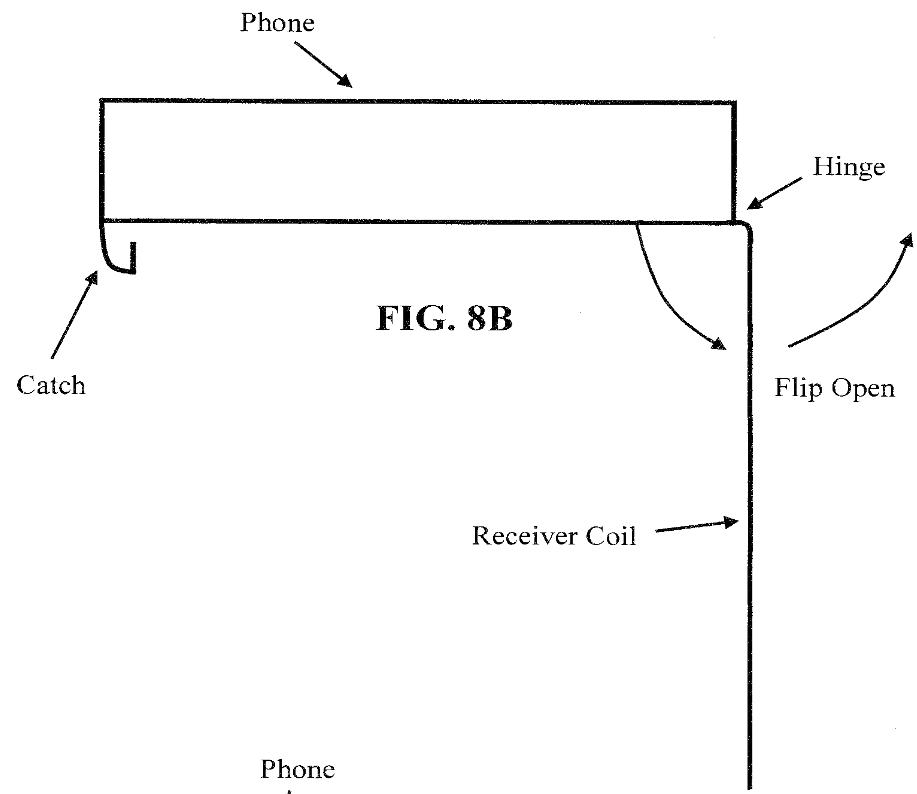

FIG. 8B

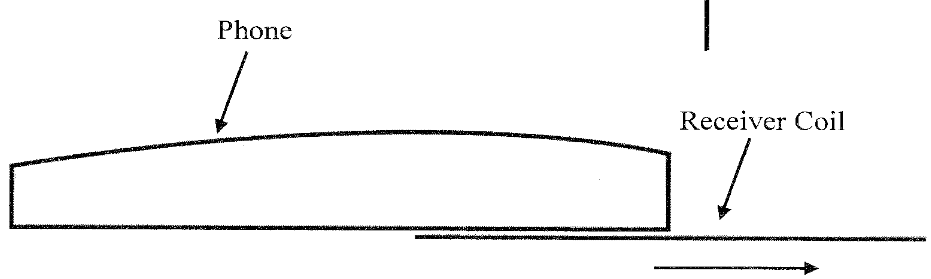

FIG. 9

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Published:
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(54) **Title:** METHOD AND APPARATUS FOR HIGH EFFICIENCY SCALABLE NEAR-FIELD WIRELESS POWER TRANSFER

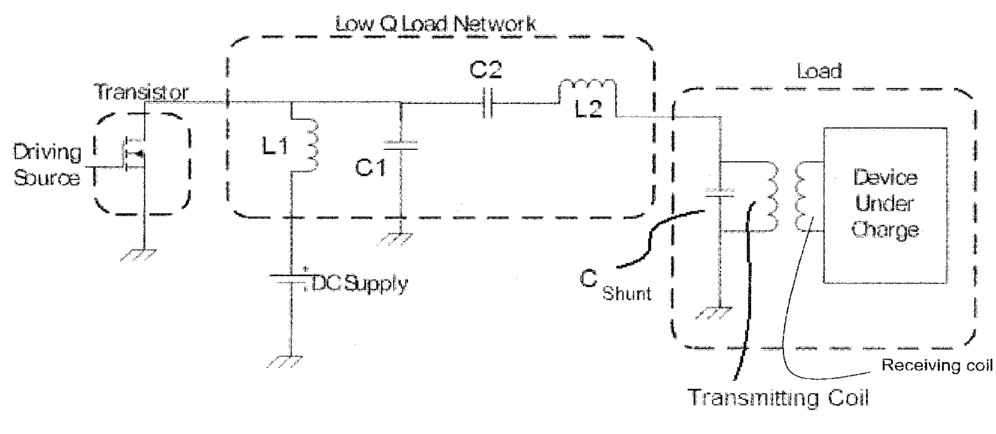


FIG. 1

(57) **Abstract:** Wireless power transfer systems are provided. A wireless power transmitter can include a transistor block providing a switching mode amplifier, and a low-Q output network including a tunable supply voltage. The transistor block can be scalable. Multiple scalable transistor blocks can be incorporated in a power transfer system using a single tunable supply voltage and transmitting coil. The tunable supply voltage can be controlled by an adaptive power supply tuning circuit to accommodate for a range of loading conditions. Embodiments of the subject invention are capable of achieving high efficiency across a wide range of output power levels.

WO 2009/070730 A2

DESCRIPTION

METHOD AND APPARATUS FOR HIGH EFFICIENCY SCALABLE NEAR-FIELD WIRELESS POWER TRANSFER

CROSS-REFERENCE TO RELATED APPLICATION

The present application claims the benefit of U.S. Provisional Application Serial No. 60/990,377, filed November 27, 2007, which is hereby incorporated by reference herein in its entirety, including any figures, tables, or drawings.

BACKGROUND OF INVENTION

Portable electronic devices such as laptop computers, LCD digital photo frames, mobile phones, and mp3 players require power to operate. Often, these devices use rechargeable batteries to provide power. The batteries are typically recharged by plugging a charger into the portable device or by removing the battery from the portable device and separately recharging the battery using a wired charger.

The cables that once restricted electronic devices are gradually being rendered unnecessary by wireless communication technology, and as the circuits that constitute the electronic devices shrink, only the power cords and batteries continue to restrict the portability of mobile electronic devices.

Current trends are leading towards going completely wireless. This means that portable devices can remain portable and can avoid having to 'plug-in' for power charging. Electro-magnetic inductive charging uses a coil to create an electromagnetic field across a charging station surface. The device then converts power from the electromagnetic field back into usable electricity, which is put to work charging the battery. Two coils are brought close to each other and when current is passed through one, the generated magnetic flux causes electromotive force to be generated in the other.

For charging of portable electronic devices and/or powering the portable electronic devices at close proximity, radio frequency charging appears to be a viable option. Recent improvements in efficiency have made it possible to consider radio frequency charging technology for commercial applications. However, there still exists a need in the art for a high efficiency low cost wireless power charging platform and components.

BRIEF SUMMARY

Embodiments of the present invention relate to a method and apparatus for wireless power transfer. Wireless power transfer systems in accordance with embodiments of the invention are capable of transmitting power to devices under charge through radio frequency charging. According to an embodiment, the wireless power transfer system of the present invention includes a switching mode amplifier and a low-Q output network. Embodiments of the present invention can follow class E amplifier designs while focusing on providing a low Q load.

In one embodiment of the present invention, a wireless transmitter includes a transistor block, a supply voltage block, and a transmitting coil block. The transistor block can include a transistor, load network capacitor, and a load network inductor. The transistor block can be scalable. The supply voltage block can include an RF choke inductor and a DC supply voltage. The supply voltage can provide tunable output power. The transmitting coil block can include a second load network capacitor, a transmitting coil, and a coil shunt capacitor.

In a further embodiment, multiple transistor blocks sharing the same load current can be used. Advantageously, the multiple transistor blocks can use a single DC supply voltage and transmitting coil block. Additional RF choke inductors can be added for each transistor block. The multiple transistor blocks can be incorporated to increase output power.

In yet another embodiment, the DC supply voltage can be tuned to achieve high efficiency for a wide range of load conditions through adaptive power supply tuning. This can be accomplished though a feedback circuit involving a current monitor network, a voltage monitor network, and a receiver for receiving charging data of a device under charge.

BRIEF DESCRIPTION OF DRAWINGS

Figure 1 shows a planar wireless power transfer system according to an embodiment of the present invention.

Figure 2 shows ideal voltage and current waveforms to prevent simultaneous high voltage and high current in a transistor of a planar wireless power transfer system according to an embodiment of the present invention.

Figure 3 shows a schematic of a transmitter according to an embodiment of the present invention with three scalable transistor blocks.

Figure 4 shows a power delivery and efficiency plot of a two-scalable transistor block transmitter according to an embodiment of the present invention at 24 V, 36 V and 48 V supply voltage.

Figure 5 shows a power delivery and efficiency plot of a two-scalable transistor block transmitter according to an embodiment of the present invention with adaptive power control.

Figure 6A shows a system block diagram of a planar wireless power transfer system using adaptive power supply tuning according to an embodiment of the present invention.

Figure 6B shows a planar wireless power transfer system according to an embodiment of the present invention including current and voltage monitoring.

Figures 7A and 7B show measurement results with different loads according to an embodiment of the present invention, where Figure 7A shows efficiency vs. power delivered to a load, and Figure 7B shows power supply voltage vs. power delivered to a load.

Figure 8 shows the efficiency at different power levels at a fixed power supply voltage of 48 V according to an embodiment of the present invention with no adaptive power supply tuning.

DETAILED DISCLOSURE

Embodiments of the present invention provide a method and apparatus for wireless power transfer. Specific embodiments of the present invention utilize radio frequency (RF) charging techniques. According to an embodiment of the present invention, a wireless power transmitter includes a transistor block providing a switching mode amplifier, and a low-Q output network including a tunable supply voltage. The combination of a switching mode amplifier and the low-Q output network provides high efficiency for power charging.

Potential uses for the subject invention include, but are not limited to, high efficiency and low cost wireless power charging platform for all portable devices such as laptop computers, LCD digital photo frames, mobile phones, mp3 players. With its high efficiency, energy loss via heating could be reduced. Further, embodiments of the subject invention can be implemented at home, in airports, and in hotel rooms. This would bring great convenience to general consumers, especially frequent travelers, as it would provide a universally charging interface and eliminate the need to bring multiple chargers. In addition, the scalable transmitter and power control design enables the transmitter system to be more flexible and adaptive to wider range of environmental situations while maintaining high energy efficiency.

Further, impedance tuning design enables the transmitter system to be more flexible and adaptive to wider range of loading conditions while maintaining high efficiency.

Figure 1 shows a schematic for a planar wireless power transfer system according to an embodiment of the present invention. Referring to Figure 1, a transmitter of the power transfer system includes a transistor block and a low Q load network. A transmitting coil with shunt capacitor is connected to the low Q load network to transmit the signal from the power transfer system to a device under charge. The device under charge should include a receiving coil for receiving the signal transmitted by the transmitter. Advantageously, the low Q network is insensitive to a change in load. In particular, the transmitter is capable of providing a range of frequencies for the transmitted signal without causing a problem with respect to load matching.

The transistor block can receive a driving power source as input. For example, a wall outlet providing ac power can provide the driving power to the transistor block. In an embodiment, the transistor block can receive the input at the gate of a transistor. A variety of transistors can be utilized in the transistor block in accordance with embodiments of the invention. In a specific embodiment, the transistor can be an n-channel metal oxide semiconductor field effect transistor (NMOSFET). In another specific embodiment, an NPN BJT transistor can be used. The transistor block can amplify an input signal and operate as an on/off switch. The output network shapes the voltage and current waveforms to prevent a simultaneous high voltage and high current in the transistor. Figure 2 shows the ideal voltage and current waveforms to prevent simultaneous high voltage and high current in the transistor. This effect minimizes power dissipation, especially during the switching transitions. Unlike a conventional amplifier using a high-Q output network to reduce harmonics in wireless communications systems, embodiments of the present invention can use a low-Q output network to provide successful operation regardless of load condition. In a specific embodiment, the low-Q output network can also be dynamic.

According to embodiments of the present invention, a maximum value for Q can be about 10. In one embodiment, Q can be selected to be less than or equal to 10. In other embodiments, Q can be selected to be less than or equal to 8 or less than or equal to 5. In further embodiments, Q can be a range of 10-20, 2-4, or 1.8-5.

In one embodiment a single transistor is used for the transistor block. The single-transistor amplifier can be used to significantly reduce the circuit complexity. Although a single transistor is illustrated in the embodiment shown in Figure 1, multiple transistors can also be used.

Embodiments of the subject invention may ensure that a low cost amplifier is able to operate at a wide dynamic range of frequency and load impedance. This is important because the load condition of the planar wireless power transfer system varies over a huge range depending on the device or devices it is powering/charging as well as the charging stage. Embodiments of the present invention have very simple hardware architecture and are able to work for a wide dynamic range of load impedance variation.

Embodiments of the subject invention provide a planar wireless power transfer system that is capable of charging portable devices as well as powering them at close proximity attaining 78% efficiency while delivering 68 W of power to an ideal load. A peak power of 132 W has been achieved using a planar wireless power transfer system according to an embodiment of the present invention. Another specific embodiment has achieved a peak power of 300W. Further, an embodiment can incorporate litz wire and achieve an efficiency of about 80%. Current existing products have not been able to achieve this level of efficiency and power output. The power capability with high efficiency makes the wireless power transmission system of the present invention suitable for charging laptop computers wirelessly.

According to an embodiment of the present invention, as illustrated in Figure 1, the transmitter is designed in a hybrid class-E parallel amplifier topology.

The class E amplifier design equations provide:

$$P = 0.576801\left(\frac{V_{CC}^{\,2}}{R}\right)\left(1.001245 - \frac{0.451759}{Q_L} - \frac{0.402444}{Q_L^{\,2}}\right)$$

$$R = 0.576801\left(\frac{V_{CC}^{\,2}}{P}\right)\left(1.001245 - \frac{0.451759}{Q_L} - \frac{0.402444}{Q_L^{\,2}}\right)$$

$$C1 = \frac{1}{34.2219 fR}\left(0.99866 + \frac{0.91424}{Q_L} - \frac{1.03175}{Q_L^{\,2}}\right) + \frac{0.6}{(2\pi f)^2 L1}$$

$$C2 = \frac{1}{2\pi f R}\left(\frac{1}{Q_L - 0.104823}\right)\left(1.00121 + \frac{1.01468}{Q_L - 1.7879}\right) + \frac{0.2}{(2\pi f)^2 L1}$$

$$L2 = \frac{Q_L R}{2\pi f}$$

$$Q_L = \frac{2\pi f L2}{R},$$

where P is the power delivered to the load of resistance R, R is the load resistance (related to the transmitting coil and shunt capacitor), V_{CC} is the supply voltage, Q_L is the load quality factor, and f is the operating frequency.

Assuming that the amplifier is ideal,

$$P = V_{CC} \times I_{DD}$$

$$PeakV_{DS} = \frac{3.56 \times V_{CC}}{SF}$$

$$PeakI_{DS} = \frac{2.86 \times I_{DD}}{SF}$$

$$\frac{t_{f/r}}{T} = \frac{\sqrt{12\left(\frac{P_{Dissipiation\ of\ f}}{P}\right)}}{2\pi\left(1+\frac{0.82}{Q_L}\right)},$$

where I_{DD} is the supply current, V_{CC} is the supply voltage, Q_L is the load quality factor, SF is the safety factor for the off nominal load condition range (typical value 0.75), $\frac{P_{Dissipiation\ of\ f}}{P}$ is the fraction of turn off power dissipation (typically < 1%), T is the period of the operating frequency, $PeakV_{DS}$ is the peak drain to source voltage (the rating of the transistor), $PeakI_{DS}$ is the peak drain to source current (the rating of the transistor), and $t_{f/r}$ is the fall time and rise time of the transistor (typically of similar value or order of magnitude).

Depending on the requirements, the transmitter can be designed for various maximum power output. In addition, by varying the supply voltage, the transmitter can tune its instantaneous power output to maintain its high efficiency while powering smaller devices such as PDA or cellular phone.

In a further embodiment of the present invention, the transistor block and low Q network of the power transfer system can be modified to achieve higher power output. In particular, the transistor block and low Q network can be provided as a scalable transistor block, tunable supply voltage block, and coil block, where the scalable transistor block ɔck and C1 and L2 from the low Q load network, the tunable supply voltage block incorporates the DC supply and L1 from the low Q load network, and the coil block incorporates C2 from the low Q load network, the coil shunt capacitor, and the transmitting coil. Although the coil block is described as including the capacitor C2 from the low Q load network, the capacitor C2 can be considered part of the load network. The transmitting coil block, not including a receiver coil for a device under charge, can be viewed as part of the load network of the transmitter.

According to embodiments of the present invention, the power supplied from the transmitter can be increased by increasing the number of scalable transistor blocks while lower cost can be achieved by decreasing the number of scalable transistor blocks. Advantageously, a single tunable supply voltage block and coil block can be used for multiple scalable transistor blocks. For example, 2-5 scalable transistor blocks can be used, sharing the load current, with a single tunable supply voltage, load network capacitor (C2), transmitting coil and coil shunt capacitor.

Figure 3 shows a schematic of a transmitter with three scalable transistor blocks according to an embodiment of the present invention. Each scalable transistor block includes a transistor, a load capacitor C1, and a load inductor L1. The transistor can be a NMOSFET. In one embodiment the NMOSFET can be an active device IRFP264NPbF. The load capacitor C1 can have a value of, for example, 3.3 nF, and the load inductor can have a value of, for example, 100 μH. A single variable DC supply can be used for tunable output power. The tunable supply voltage block includes the DC supply and an inductor for each scalable transistor block acting as an RF choke. The inductor acting as an RF choke can have a value of, for example, 500 μH.

A single coil block is connected to the three scalable transistor blocks. The coil block includes a transmitting coil, a coil shunt capacitor, and a load network capacitor C2. The load network capacitor C2 and the coils shunt capacitor can each have a value of, for example, 100 nF. In the embodiment shown in Figure 3, a buffer and input clock oscillator can be included as input to the scalable transistor blocks.

Although each scalable transistor block illustrated in Figure 3 is indicated as having an RF choke inductor, embodiments of the present invention are not limited thereto. For example, a single RF choke inductor can be shared by the scalable transistor blocks to reduce cost.

As indicated by Figure 3, the DC supply can provide tunable output power. The can be affected by the voltage of the DC supply voltage. Figure 4 shows a power delivery and efficiency plot of a two scalable block transmitter at 24 V, 36 V and 48 V supply voltage. In particular, in Figure 4, the lower curve shows the 48V supply voltage results, the middle curve shows the 36V supply voltage results, and the upper left curve shows the 24V supply voltage results. As illustrated by the plots of Figure 4, an efficiency of about 78% is possible over a range of supply voltage. Figure 5 shows a power delivery and efficiency plot of a two scalable block transmitter according to an embodiment of the present invention with tunable power control. As illustrated by the plot shown in Figure 5, a high efficiency is possible over a large range of power.

This system has simple hardware architecture and each scalable block is a duplicate of each other that maintains the architecture of the transmitter. In addition, the power control enables high efficiency across a wide band power level. Therefore, it is able to achieve low cost for future custom design to meet various needs. Embodiments can also incorporate blocks that have different parameters, and that can be utilized together in various combinations. In this way, power delivery can be unevenly distributed.

Further embodiments of the present invention can include adaptive power supply tuning. The adaptive power supply tuning can be performed on a modified switching-mode amplifier with low-Q output network. According to one embodiment of the present invention, the power supply voltage of a wireless transmitter is tuned based on the feedback from a receiver load, which is the device under charge, to optimize the efficiency. Referring back to Figures 1 and 3, since the power supply (labeled as DC supply) has a direct path to the transmitting coil via the switching mode amplifier, the input impedance of the transmitter seen by the power supply is related to the impedance of the transmitting coil. The impedance of the transmitting coil is related to the coupling coefficient of the coils as well as the receiver load impedance. The receiver load impedance varies with the input voltage to the receiver load's voltage regulator. By adjusting the load impedance, by tuning the power supply voltage, the load impedance seen by the transmitter can be closer to an optimal value. Accordingly, varying the DC supply in the embodiments shown in Figures 1 and 3, the voltage at the input of the receiver's voltage regulator will change, so as to change the impedance looking into the input port of the voltage regulator. According to embodiments of the present invention incorporating an adaptive power supply, the tuning system measures the voltage and current delivered to the receiver load and attempts to tune its impedance via varying the power supply voltage. Therefore, by tuning the power supply voltage, power control as well as impedance tuning can be achieved. The adaptive tuning method can be implemented using a low cost microprocessor as well as a programmable switching regulator.

A block diagram of a planar wireless power transfer system using adaptive power supply tuning according to an embodiment of the present invention is shown in Figure 6A. Referring to Figure 6A, a wireless charger according to any embodiment of the present invention can be connected to an adaptive power control circuit. The adaptive power control circuit can include a programmable regulator and feedback network to set programmable regulator output voltage; a voltage monitor network and current monitor network that each receive the output voltage as input; and a microprocessor that receives input from the current monitor network and the voltage monitor network, and provides feedback to the feedback network. The microprocessor can also receive charging data from a device under charge. Examples of charging data that can be provided from the device under charge includes, but is not limited to, charge status, voltage provided to receiver, and/or current into the receiver. The device under charge can provide receiver charging data to a wireless transmitter. In other embodiments, other types of data links than wireless can be used. The charging data can then be received by a receiver of the adaptive power supply tuning circuit and provided to the microprocessor. The current monitor network can provide the power supply to the wireless charger. As examples, the output of the current monitor network in Figure 6A can be the DC supply of Figure 1 or the DC supply variable of Figure 3.

In an embodiment, tuning of the transmitter can be accomplished using voltage and current monitoring. For example, as illustrated in Figure 6B, the voltage across the drain of the transistor and the voltage and current across the transmitting coil can be monitored. These monitored values can be used as input in a feedback system for tuning control. In another specific embodiment, the supply DC current and the voltage across the transmitting coil can be monitored and provided for tuning of the power supply, such as a DC power supply. The power supply can then be tuned by, for example, inputting the monitored data into a lookup table created based on a determination of desired efficiency and the output of the power supply tuned accordingly.

Figures 7A and 7B show measurement results for the embodiment of Figure 6A with different loads. Referring to Figure 7A, it is possible to maintain efficiency at over 75% in all conditions. Figure 7B shows power supply voltage vs. power delivered to the load. Power supply voltage is tuned to achieve high efficiency at each load condition.

As a comparison, Figure 8 shows the efficiency at different power levels when no adaptive power supply tuning is included. Here the power supply voltage is fixed at 48 V.

Previous planar wireless power transfer systems tend to suffer low efficiency at low output power, even though the efficiency at higher output power is higher. In contrast, embodiments of the present invention are capable of maintaining a high efficiency of over 75% across a wide range of output power levels. Current existing products cannot achieve this level of efficiency and power output. The adaptive power supply tuning according to embodiments of the present invention enables high efficiency across a wide range of output power levels. Therefore, embodiments of the present invention are able to achieve low cost for future custom designs to meet various needs.

All patents, patent applications, provisional applications, and publications referred to or cited herein are incorporated by reference in their entirety, including all figures and tables, to the extent they are not inconsistent with the explicit teachings of this specification.

It should be understood that the examples and embodiments described herein are for illustrative purposes only and that various modifications or changes in light thereof will be suggested to persons skilled in the art and are to be included within the spirit and purview of this application.

CLAIMS

What is claimed is:

1. A wireless power transfer system, comprising:

a transistor, wherein the transistor has an on state and an off state;

an output network, wherein the output network is low Q; and

a transmitting coil block, wherein the transmitting coil block is coupled to the output network, wherein the transmitting coil block comprises a transmitting coil,

wherein when the transistor is in the on state the transistor supplies current to the output network and the output network drives the transmitting coil, wherein when the transistor is in the on state the transistor has a low voltage and high current across the transistor, wherein when the transistor is in the off state the transistor has a low current and high voltage across the transistor.

2. The system according to claim 1, wherein Q is less than 20.

3. The system according to claim 1, wherein Q is less than 10.

4. The system according to claim 1, wherein Q is less than 8.

5. The system according to claim 1, wherein Q is in the range of 10-20.

6. The system according to claim 1, wherein Q is in the range of 1.8-5.

7. The system according to claim 1, wherein Q is in the range of 2-4.

8. The system according to claim 1, further comprising:

a receiver coil, wherein the receiving coil is inductively coupled to the transmitting coil, wherein the receiver coil is capable of delivering power to a load.

9. The system according to claim 1, further comprising: a first choke inductor coupled between the drain of the transistor and a power supply, wherein the output network comprises:

a first load network capacitor coupled between a drain of the transistor and ground;

a first load network inductor, wherein the first load inductor is in series between the drain of the transistor and the transmitting coil block.

10. The system according to claim 9, wherein the power supply is a DC power supply.

11. The system according to claim 9, wherein the transmitting coil block comprises a series capacitor in series with the transmitting coil.

12. The system according to claim 11, wherein the transmitting block comprises a shunt capacitor in parallel with the transmitting coil.

13. The system according to claim 9, wherein the output network comprises a series capacitor in series with the first load network inductor.

14. The system according to claim 10, wherein the DC power supply is a variable DC supply.

15. The system according to claim 14, further comprising a tuning circuit capable of receiving charging data from a device under charge for adaptive power supply tuning of the variable DC supply.

16. The system according to claim 9, wherein the output network further comprises:

a second transistor, wherein the second transistor has a second on state and a second off state;

a second load network capacitor coupled between a drain of the second transistor and ground; and a second load network inductor, wherein the second load network inductor is in series between the drain of the second transistor and the transmitting coil block

wherein when the second transistor is in the second on state the second transistor supplies current to the output network and the output network drives the transmitting coil, wherein when the second transistor is in the second on state the second transistor has a low voltage and high current across the second transistor, wherein when the second transistor is in the second off state the second transistor has a low current and high voltage across the second transistor.

17. The system according to claim 16, further comprising:

a second choke inductor coupled between the drain of the second transistor and the DC supply.

18. The system according to claim 9, wherein the output network further comprises:

at least one additional transistor, wherein the at least one additional transistor has an at least one additional on state and an at least one additional off state;

at least one additional load network capacitor coupled between a drain of a corresponding at least one additional transistor and ground; and

at least one additional load network inductor, wherein the at least one additional load network is in a series between the drain of the at least one additional transistor and the transmitting coil block,

wherein when the at least one additional transistor is in the at least one additional on state the at least one additional transistor supplies current to the output network and the output network drives the transmitting coil, wherein when the at least one additional transistor is in the at least one additional on state the at least one additional transistor has a low voltage and high current across the at least one additional transistor, wherein when the at least one additional transistor is in the at least one additional off state, the at least one additional transistor has a low current and high voltage across the at least one additional transistor.

19. The system according to claim 18, further comprising:

at least one additional choke inductor coupled between the drain of the corresponding at least one additional transistor and the DC supply.

20. The system according to claim 1, wherein the transmitting coil block further comprises:

a coil shunt capacitor coupled in parallel with the transmitting coil.

21. The system according to claim 1, further comprising a tuning circuit capable of receiving charging data from a device under charge for adaptive power supply tuning.

22. The system according to claim 1, further comprising a tuning circuit capable of receiving voltage and current data from across the transmitting coil and voltage data from across the transistor as input for adaptive power supply tuning.

23. The system according to claim 9, further comprising a tuning circuit capable of receiving output current of the power supply and voltage across the transmitting coil for adaptive supply tuning.

24. The system according to claim 1, further comprising:

a driving source coupled to a gate of the transistor.

25. The system according to claim 24, wherein the driving source is an ac power source.

26. A method for wireless power transfer, comprising:

providing a transistor, wherein the transistor has an on state and an off state;

providing an output network, wherein the output network is low Q;

providing a transmitting coil block, wherein the transmitting coil block is coupled to the output network, wherein the transmitting coil block comprises a transmitting coil; and

driving a gate of the transistor with a driving source, wherein when the transistor is in the on state the transistor supplies current to the output network and the output network drives the transmitting coil, wherein when the transistor is in the on state the transistor has a low voltage and high current across the transistor, wherein when the transistor is in the off state the transistor has a low current and high voltage across the transistor.

27. The method according to claim 26, wherein Q is less than 20.

28. The method according to claim 26, wherein Q is less than 10.

29. The method according to claim 26, wherein Q is less than 8.

30. The method according to claim 26, wherein Q is in the range of 10-20.

31. The method according to claim 26, wherein Q is in the range of 1.8-5.

32. The method according to claim 26, wherein Q is in the range of 2-4.

33. The method according to claim 26, further comprising:

providing a receiver coil, wherein the receiving coil is inductively coupled to the transmitting coil, wherein the receiver coil is capable of delivering power to a load.

34. The method according to claim 26, further comprising: providing a first choke inductor coupled between the drain of the transistor and a power supply, wherein the output network comprises:

a first load network capacitor coupled between a drain of the transistor and ground;

a first load network inductor, wherein the first load inductor is in series between the drain of the transistor and the transmitting coil block.

35. The method according to claim 34, wherein the power supply is a DC power supply.

36. The method according to claim 34, wherein the transmitting coil block comprises a series capacitor in series with the transmitting coil.

37. The method according to claim 36, wherein the transmitting block comprises a shunt capacitor in parallel with the transmitting coil.

38. The method according to claim 34, wherein the output network comprises a series capacitor in series with the first load network inductor.

39. The method according to claim 35, wherein the DC power supply is a variable DC supply.

40. The method according to claim 39, further comprising providing a tuning circuit capable of receiving charging data from a device under charge for adaptive power supply tuning of the variable DC supply; and

tuning the variable DC supply base on the received charging data.

41. The method according to claim 34, wherein the output network further comprises:

providing a second transistor, wherein the second transistor has a second on state and a second off state;

providing a second load network capacitor coupled between a drain of the second transistor and ground;

providing a second load network inductor, wherein the second load network inductor is in series between the drain of the second transistor and the transmitting coil block; and

driving a second gate of the second transistor with the driving source,

wherein when the second transistor is in the second on state the second transistor supplies current to the output network and the output network drives the transmitting coil, wherein when the second transistor is in the second on state the second transistor has a low voltage and high current across the second transistor, wherein when the second transistor is in the second off state the second transistor has a low current and high voltage across the second transistor.

42. The method according to claim 41, further comprising:

providing a second choke inductor coupled between the drain of the second transistor and the DC supply.

43. The method according to claim 34, wherein the output network further comprises:

providing at least one additional transistor, wherein the at least one additional transistor has an at least one additional on state and an at least one additional off state;

providing at least one additional load network capacitor coupled between a drain of a corresponding at least one additional transistor and ground;

providing at least one additional load network inductor, wherein the at least one additional load network is in a series between the drain of the at least one additional transistor and the transmitting coil block; and

driving the at least one additional gate of the at least one additional transistor with the driving source,

wherein when the at least one additional transistor is in the at least one additional on state the at least one additional transistor supplies current to the output network and the output network drives the transmitting coil, wherein when the at least one additional transistor is in the at least one additional on state the at least one additional transistor has a low voltage and high current across the at least one additional transistor, wherein when the at least one additional transistor is in the at least one additional off state, the at least one additional transistor has a low current and high voltage across the at least one additional transistor.

44. The method according to claim 43, further comprising:

providing at least one additional choke inductor coupled between the drain of the corresponding at least one additional transistor and the DC supply.

45. The method according to claim 26, wherein the transmitting coil block further comprises:

a coil shunt capacitor coupled in parallel with the transmitting coil.

46. The method according to claim 26, further comprising providing a tuning circuit capable of receiving charging data from a device under charge for adaptive power supply tuning; and

tuning the power supply based on the received charging data.

47. The method according to claim 26, further comprising providing a tuning circuit capable of receiving voltage and current data from across the transmitting coil and voltage data from across the transistor as input for adaptive power supply tuning; and

tuning the power supply based on the voltage and current data from across the transmitting coil and voltage data from across the transistor.

48. The method according to claim 34, further comprising providing a tuning circuit capable of receiving output current of the power supply and voltage across the transmitting coil for adaptive supply tuning; and

tuning the power supply based on the output current of the power supply and voltage across the transmitting coil.

49. The method according to claim 26, wherein the driving source is an ac power source.

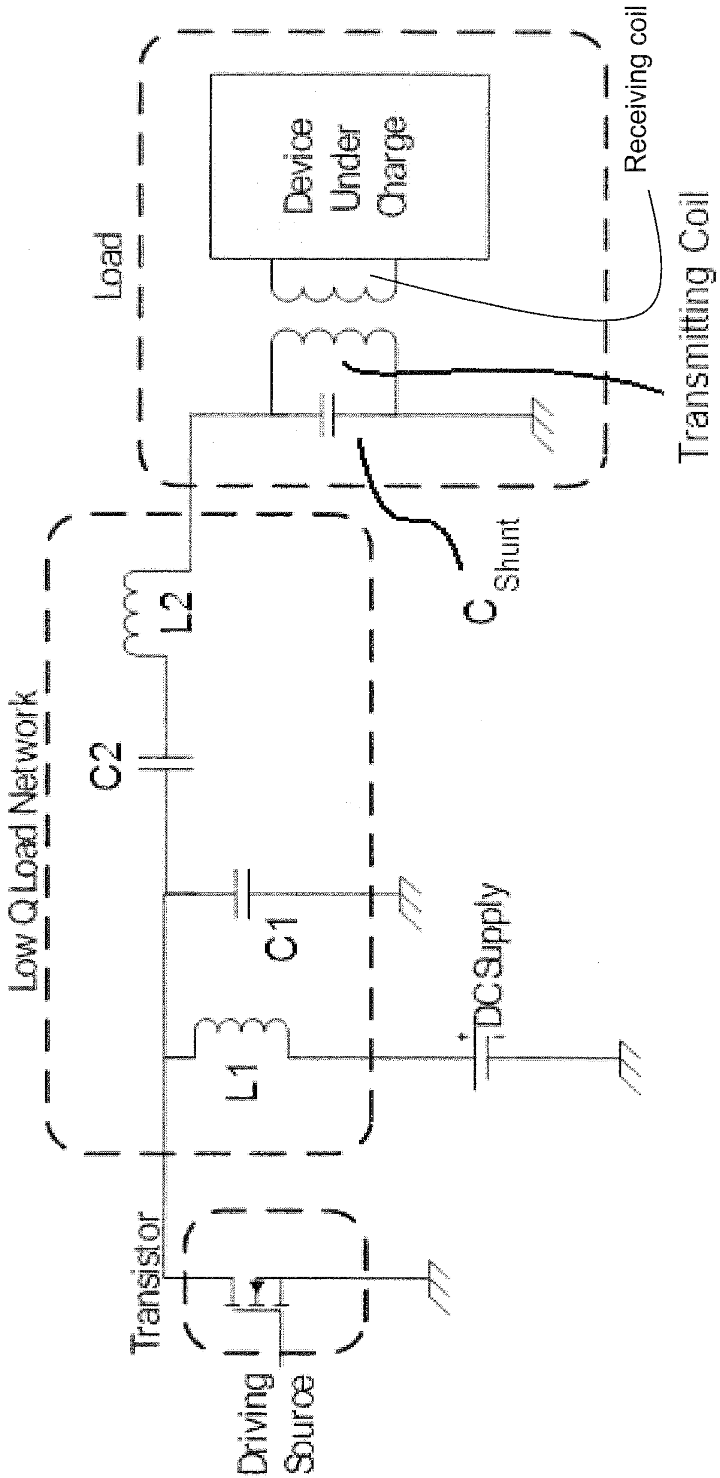

Load
Device Under Charge
Receiving coil
Transmitting Coil
C_{Shunt}
L2
C2
Low Q Load Network
C1
L1
DC Supply
Transistor
Driving Source

FIG. 1

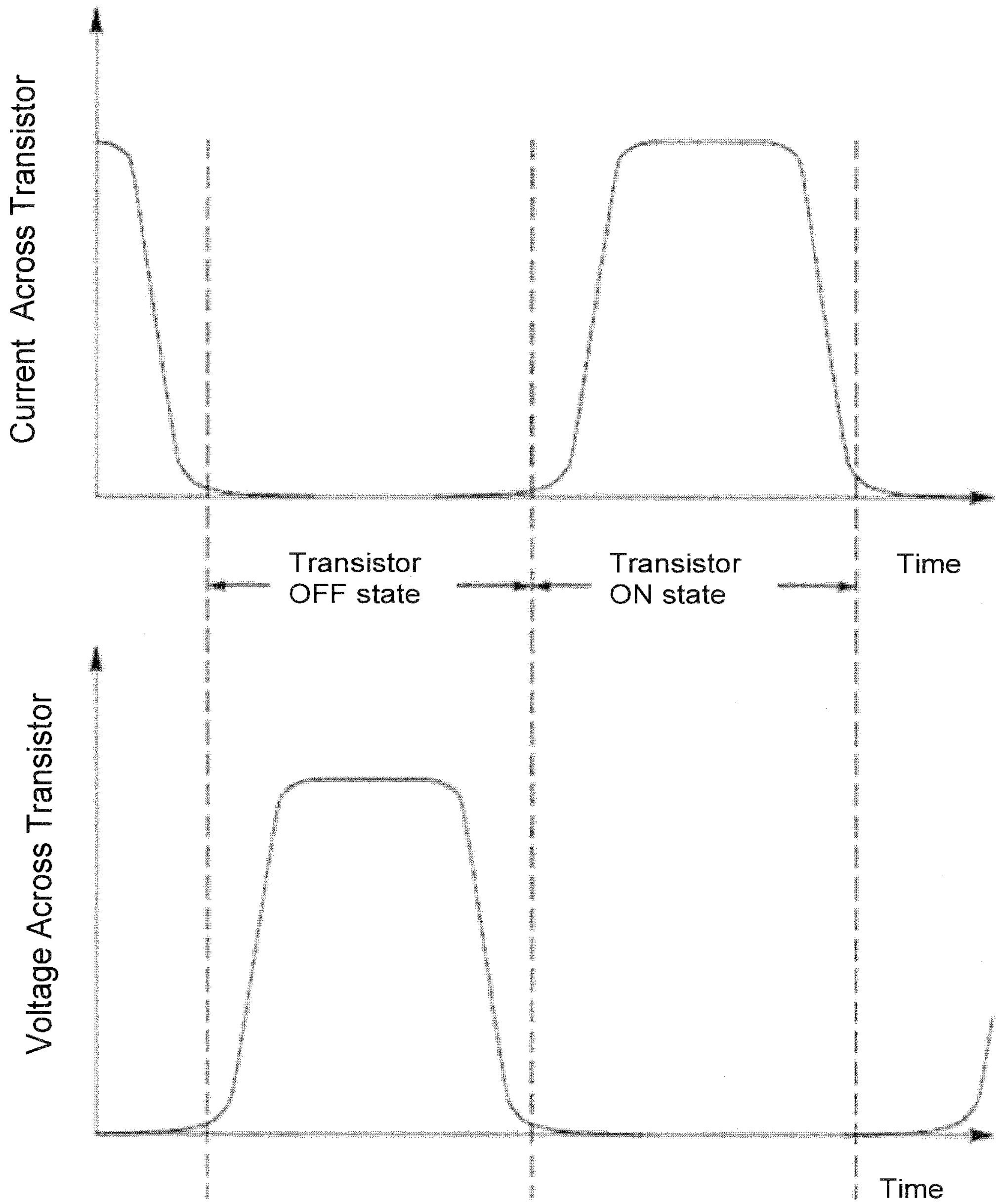

Current Across Transistor
Transistor OFF state
Transistor ON state
Time
Voltage Across Transistor
Time

FIG. 2

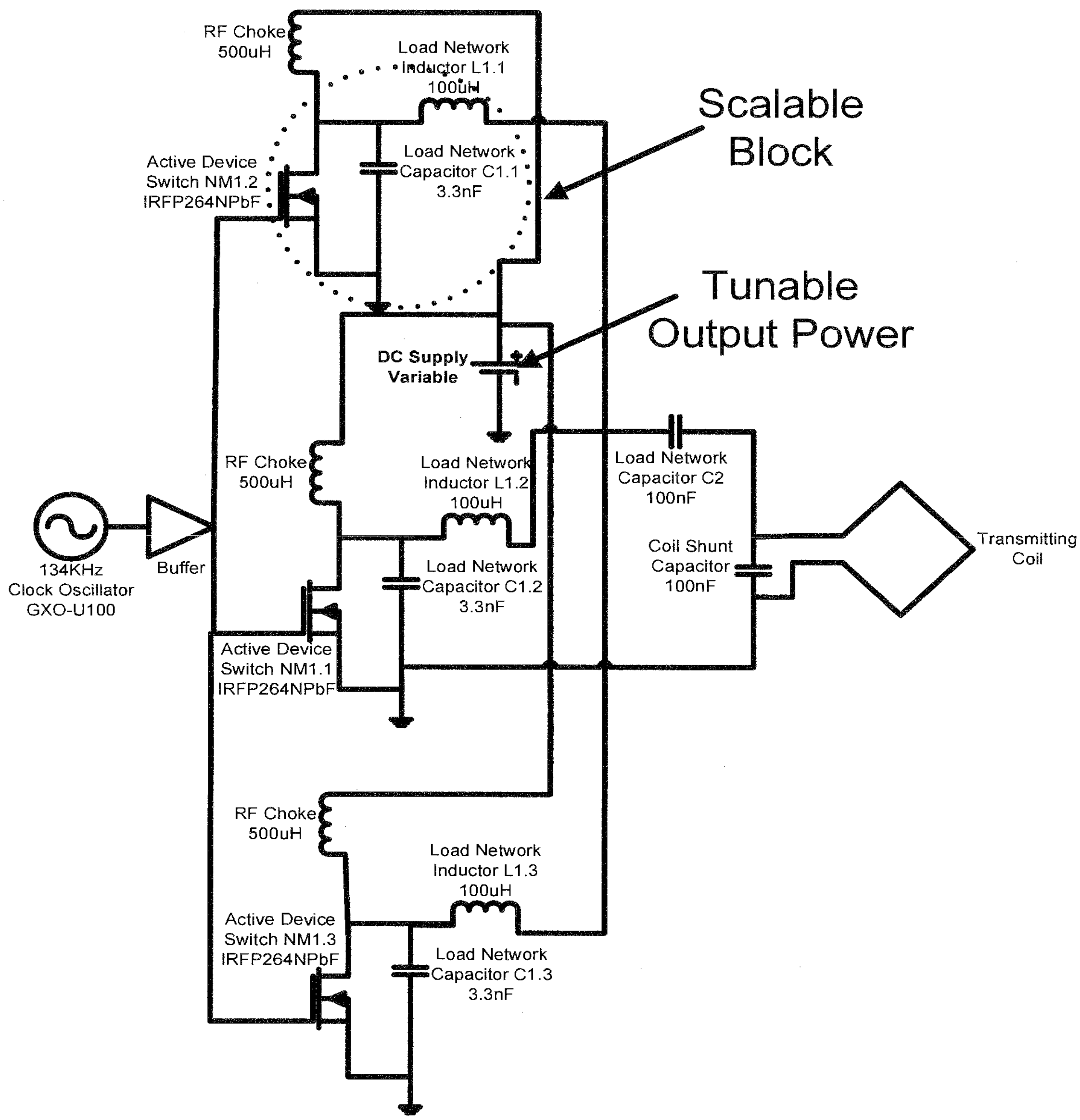

RF Choke
500uH
Load Network
Inductor L1.1
100uH
Scalable
Block
Load Network
Capacitor C1.1
3.3nF
Active Device
Switch NM1.2
IRFP264NPbF
Tunable
Output Power
DC Supply
Variable
RF Choke
500uH
Load Network
Inductor L1.2
100uH
Load Network
Capacitor C2
100nF
Coil Shunt
Capacitor
100nF
Transmitting
Coil
Load Network
Capacitor C1.2
3.3nF
134KHz
Clock Oscillator
GXO-U100
Buffer
Active Device
Switch NM1.1
IRFP264NPbF
RF Choke
500uH
Load Network
Inductor L1.3
100uH
Active Device
Switch NM1.3
IRFP264NPbF
Load Network
Capacitor C1.3
3.3nF

FIG. 3

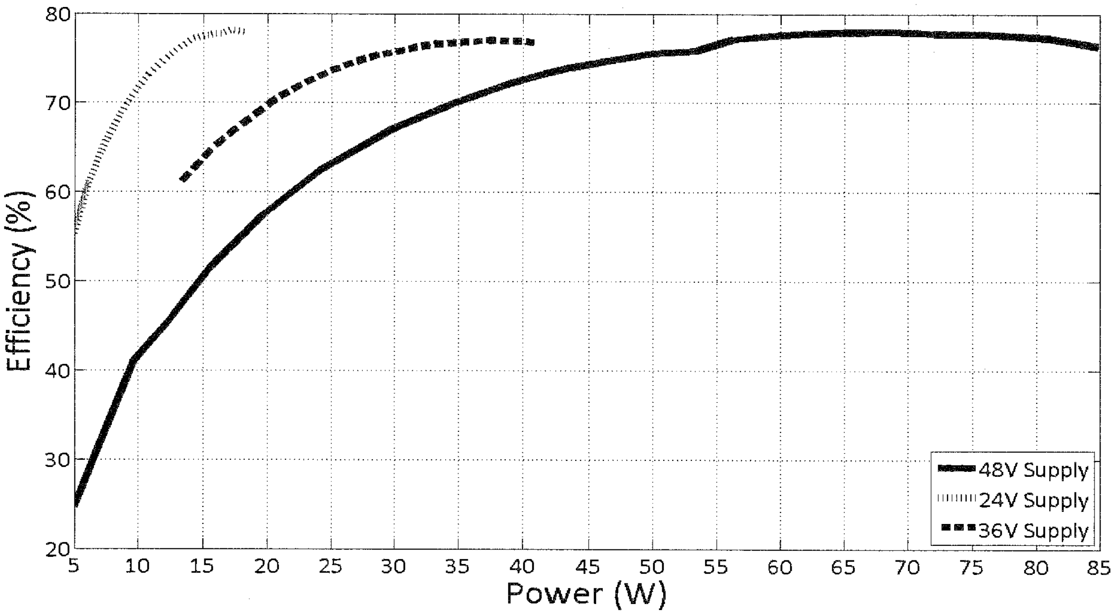

Efficiency (%)
Power (W)
48V Supply
24V Supply
36V Supply

FIG. 4

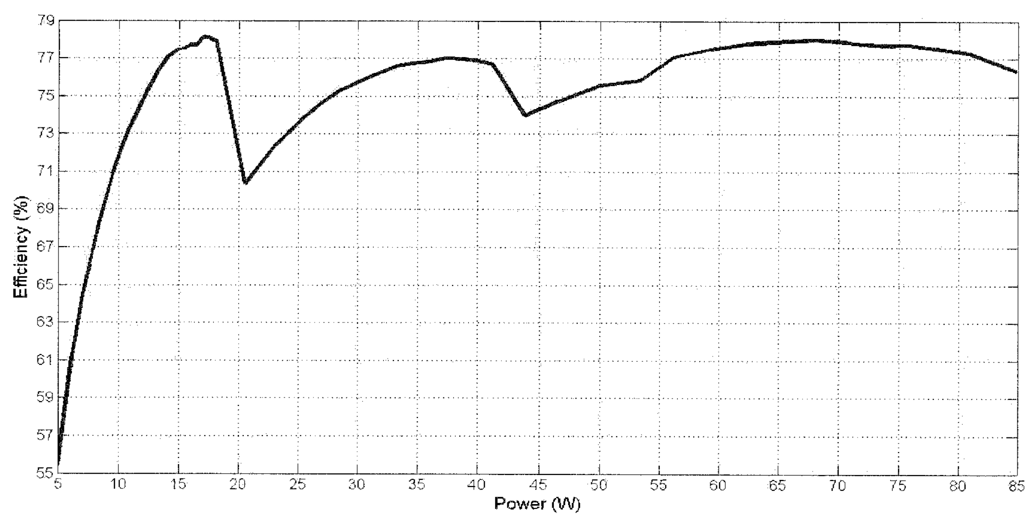

Efficiency (%)
Power (W)

FIG. 5

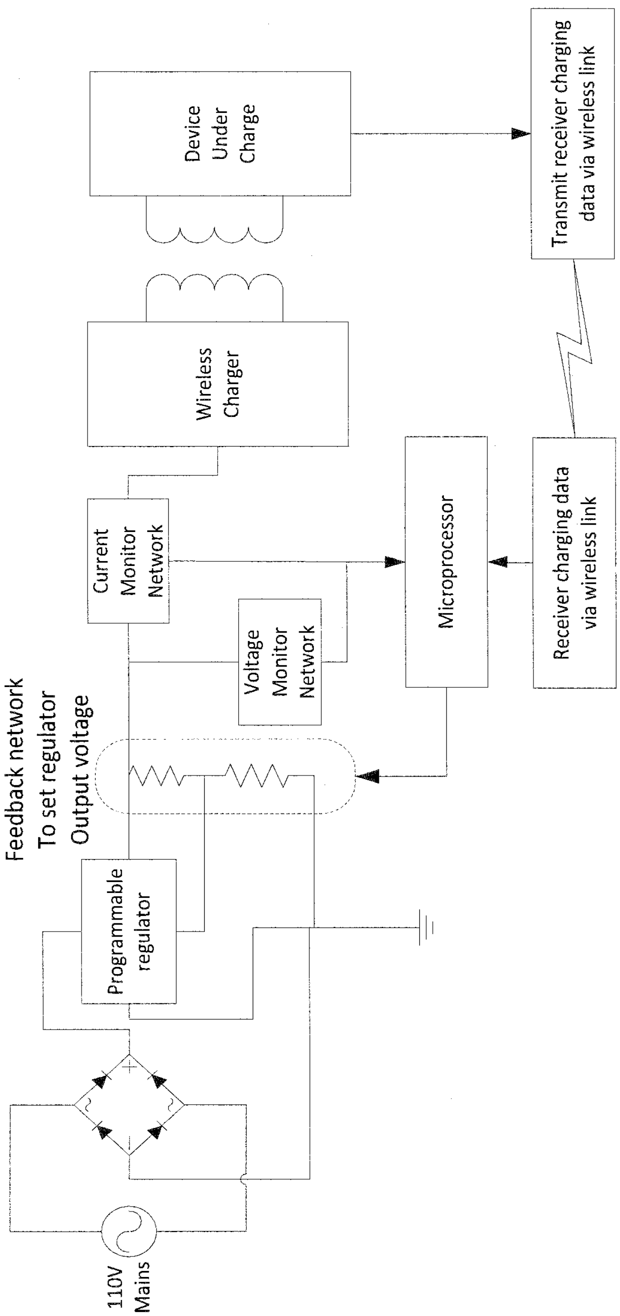

110V
Mains
Programmable
regulator
Feedback network
To set regulator
Output voltage
Voltage
Monitor
Network
Current
Monitor
Network
Microprocessor
Receiver charging data
via wireless link
Wireless
Charger
Device
Under
Charge
Transmit receiver charging
data via wireless link

FIG. 6A

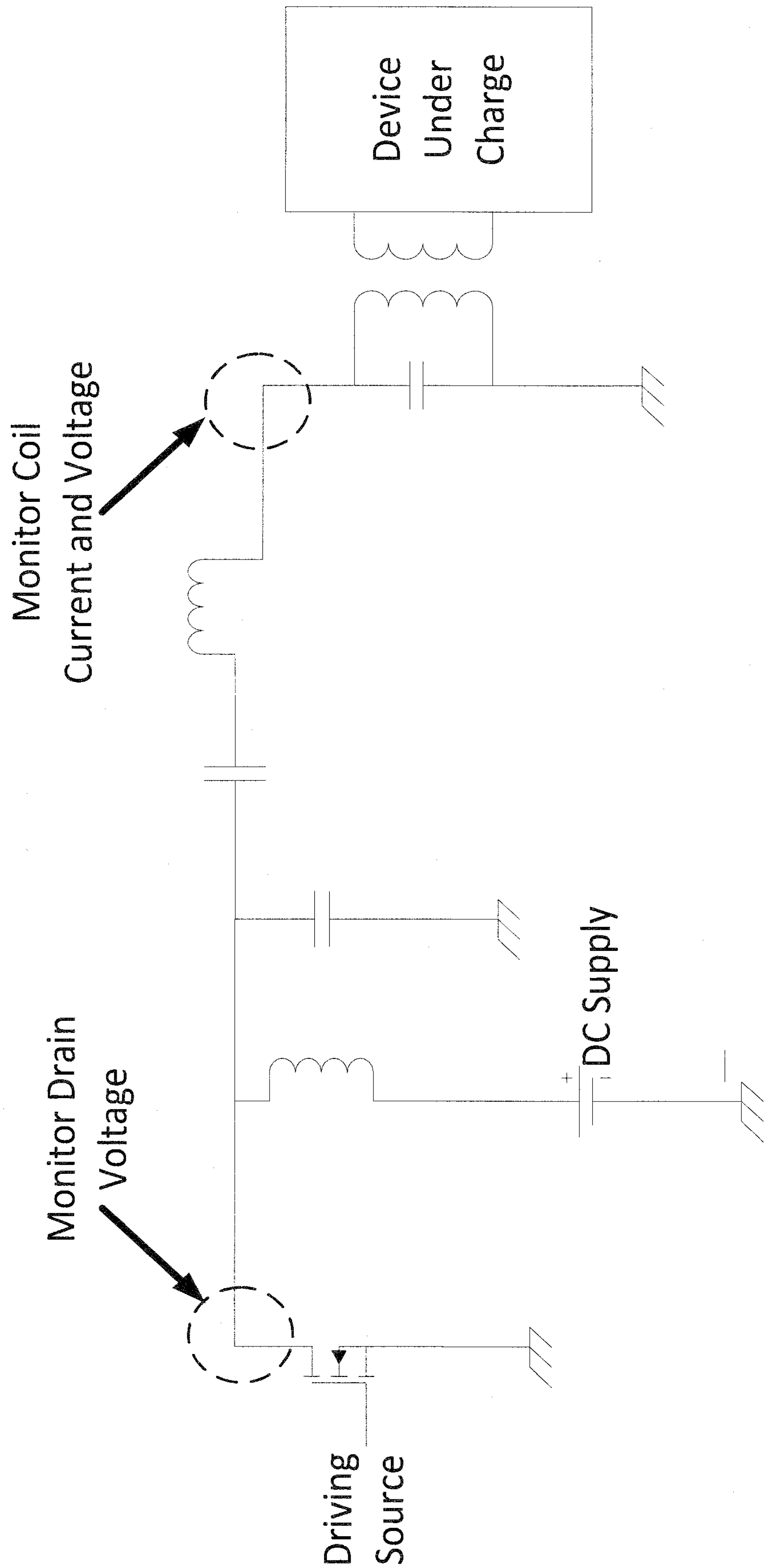

Device
Under
Charge
Monitor Coil
Current and Voltage
DC Supply
Monitor Drain
Voltage
Driving
Source

FIG. 6B

FIG. 7A

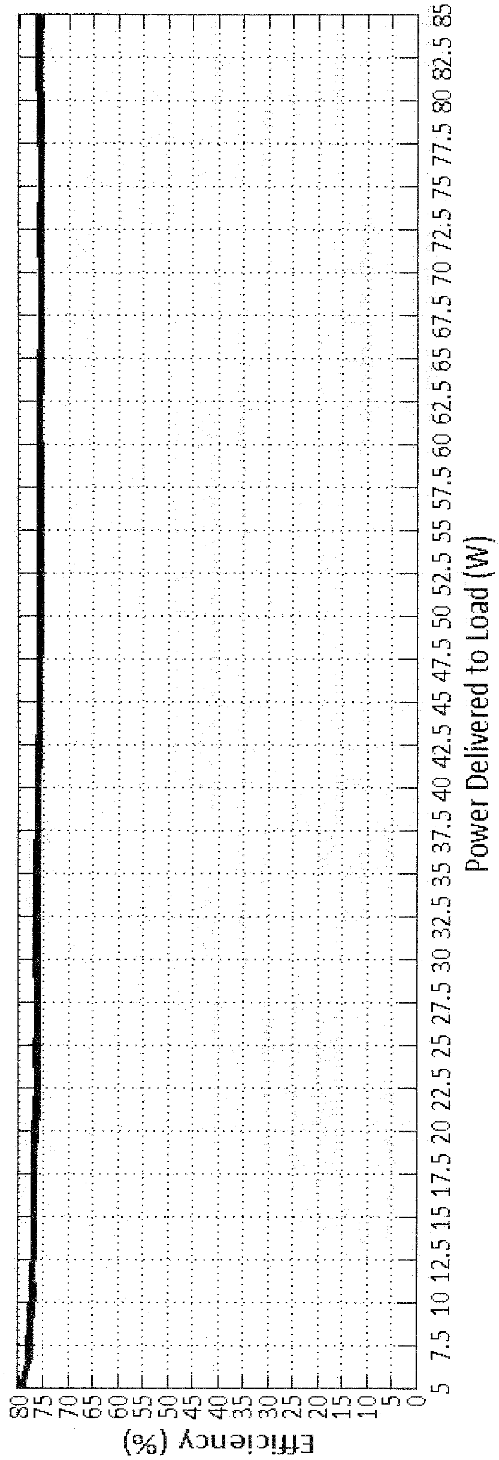
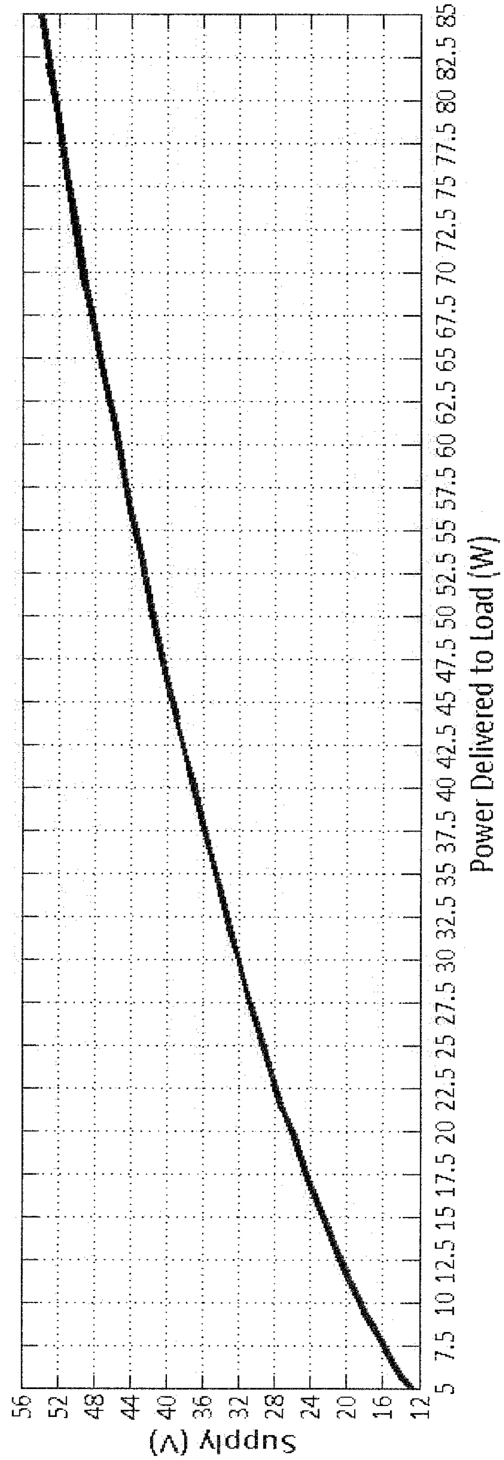

Efficiency (%)
80 75 70 65 60 55 50 45 40 35 30 25 20 15 10 5 0
5 7.5 10 12.5 15 17.5 20 22.5 25 27.5 30 32.5 35 37.5 40 42.5 45 47.5 50 52.5 55 57.5 60 62.5 65 67.5 70 72.5 75 77.5 80 82.5 85
Power Delivered to Load (W)

FIG. 7B


Supply (V)
56 52 48 44 40 36 32 28 24 20 16 12
5 7.5 10 12.5 15 17.5 20 22.5 25 27.5 30 32.5 35 37.5 40 42.5 45 47.5 50 52.5 55 57.5 60 62.5 65 67.5 70 72.5 75 77.5 80 82.5 85
Power Delivered to Load (W)

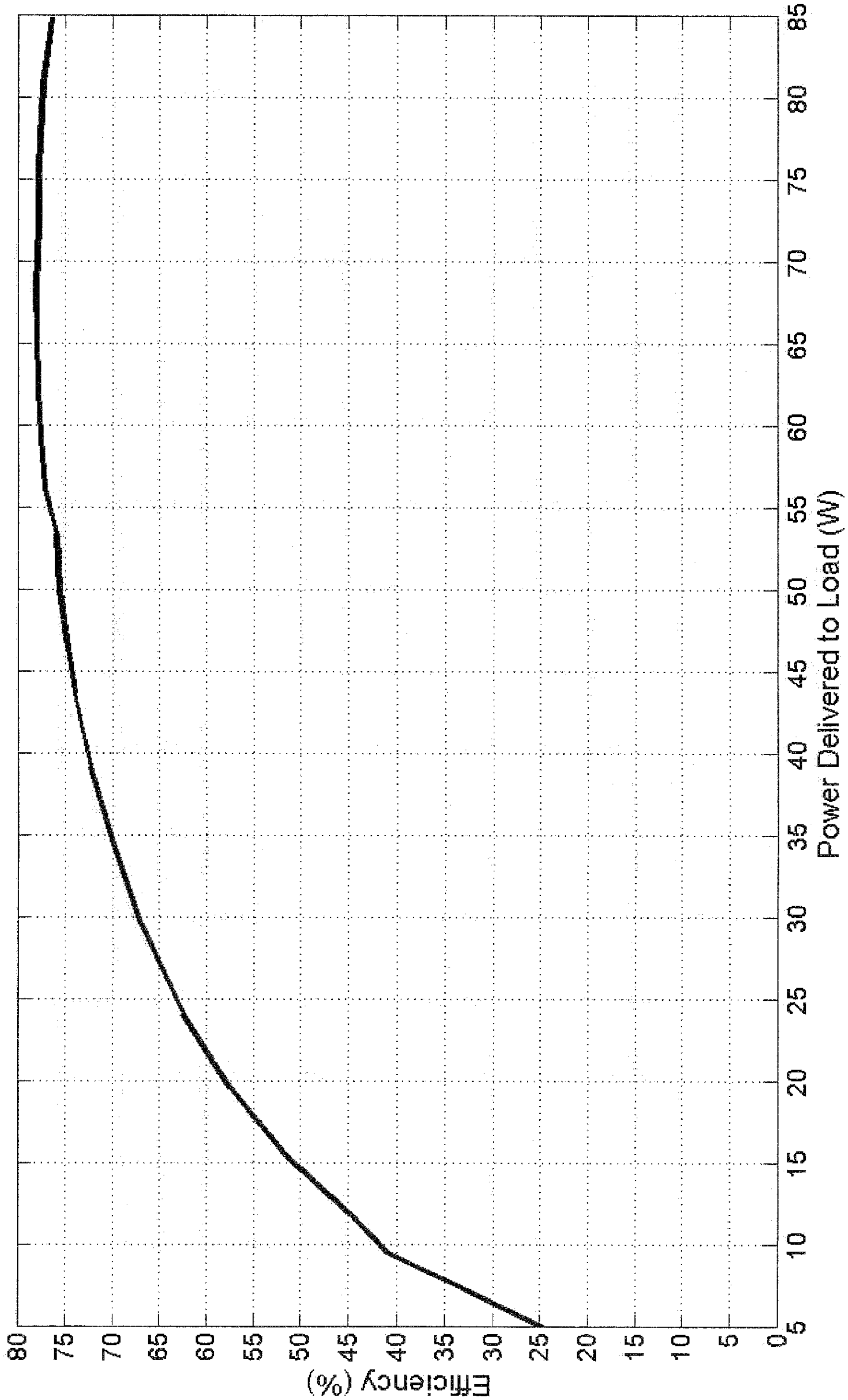

Efficiency (%)
0 5 10 15 20 25 30 35 40 45 50 55 60 65 70 75 80
Power Delivered to Load (W)
5 10 15 20 25 30 35 40 45 50 55 60 65 70 75 80 85

FIG. 8

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(54) Title: POWER CONTROL DUTY CYCLE THROTTLING SCHEME FOR PLANAR WIRELESS POWER TRANSMISSION SYSTEM

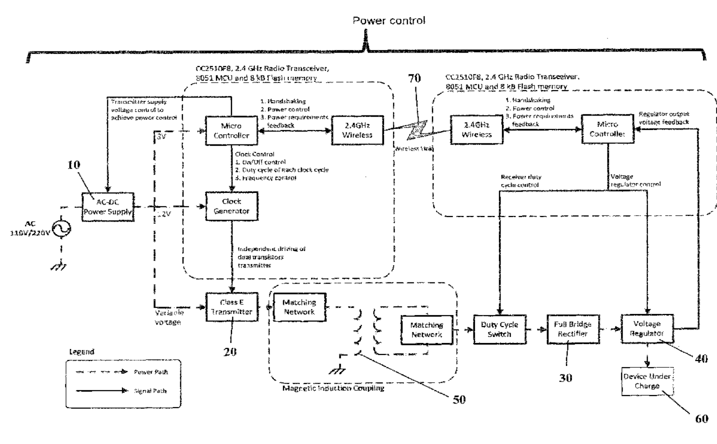


FIG. 1

(57) Abstract: Embodiments of a power transmission system and power control scheme are provided. The power control scheme utilizes the power requirements for each receiver of a device on a charging pad for highly efficient charging of multiple devices. According to an embodiment, each receiver transmits its power requirement to the transmitter, and based on this power requirement, the most power hungry device is used to set the duty cycle of the transmitter. Each individual receiver can continue to monitor its power requirement and make necessary adjustments to ensure efficient power transfer. Once all the devices are fully charged, the transmitter can be powered off or have its duty cycle reduced to performance trickle charging. The transmitter can continue to access the load conditions while in standby to detect any new device being placed on the charging pad.

WO 2009/126963 A2

DESCRIPTION

POWER CONTROL DUTY CYCLE THROTTLING SCHEME FOR PLANAR WIRELESS POWER TRANSMISSION SYSTEM

CROSS-REFERENCE TO RELATED APPLICATION

The present application claims the benefit of U.S. Provisional Application Serial No. 61/044,327, filed April 11, 2008, which is hereby incorporated by reference herein in its entirety, including any figures, tables, or drawings.

BACKGROUND OF INVENTION

Wireless power transmission is an emerging technology as trends move toward completely wireless devices. Currently, power control can be achieved via changing the operating frequency of a transmitter, tuning the transmitter's matching network, and changing the transmitter's supply voltage. However, limited dynamic range can be attained by changing the operating frequency. It can be difficult and costly to find tunable components to operate at such high power level. In addition, system performance tends to degrade when the supply voltage deviates too far from it nominal operating voltage.

BRIEF SUMMARY

Embodiments of the present invention relate to a method and apparatus for power control in wireless power transmission. According to an embodiment, multiple receivers can be used with a single transmitter. In an embodiment, a low cost microprocessor can be used to perform timing control for the transmitter output. Embodiments of the present invention have a significantly wider power dynamic range than other competing solutions. In addition, it is possible to scale the timing control technique to charge a large number of devices concurrently, thus becoming more flexible.

In an embodiment, an intelligent power control technique is implemented that attends to the power requirement for the receiver of each device on a charging pad. According to an embodiment, each receiver transmits the power requirement of the receiver to the transmitter, and based on this power requirement, the most power hungry device is used to set the duty cycle of the transmitter. Each individual receiver can continue to monitor its power requirement and make necessary adjustments to ensure efficient power transfer. Once all the devices are fully charged, the transmitter can be powered off or have its duty cycle reduced to perform trickle charging. The transmitter can continue to access the load conditions while in standby to detect any new device being placed on the charging pad.

BRIEF DESCRIPTION OF DRAWINGS

Figure 1 shows a system overview according to an embodiment of the present invention.

Figure 2 shows a duty cycling throttling scheme according to an embodiment of the present invention.

Figure 3 shows a timing scheme according to an embodiment of the present invention for three receivers of significantly different per-unit area power requirement (75W, 25W, 5W) placed on a 100W transmitting pad.

Figure 4 shows an optimum duty cycling throttling scheme according to an embodiment of the present invention.

DETAILED DISCLOSURE

Embodiments of the present invention provide a method and apparatus for power control in wireless power transmission. Embodiments of the subject invention can be utilized in a planar power transmission system that is capable of charging multiple portable devices as well as powering them at close proximity. Since different devices have different power requirements that can span across a large dynamic range, embodiments of the present invention provide an intelligent power control scheme that allows contemporaneous charging of multiple devices that may span across a large dynamic range. According to an embodiment, multiple receivers can be used with a single transmitter. Certain embodiments of the present invention can attain 81% efficiency while delivering 68W of power to an ideal load. A peak power of 132W has been achieved using a specific embodiment of the present invention. In a further embodiment, a peak power of 300W has been achieved.

Figure 1 illustrates a power transmission system according to an embodiment of the present invention that can utilize one or more power control schemes. The power-related systems of this example are shown, following a power path. Power control techniques according to embodiments of the present invention can be applied to the power-related systems. In the transmitter portion of the system, the amplifier supply voltage **10** and the duty cycle can be tuned. In a specific embodiment, a duty cycle of driving square pulses can be tuned. In addition, in a system where the class E transmitter portion **20** includes two transistors, one of the two transmitting transistors can be shut down to reduce power output.

In one embodiment, two independent clocks can be used to separately drive the two transistors, such that a power control scheme is capable of shutting down one transistor when the power requirement is low. Synchronization between the independent clocks can be provided.

In an embodiment of the power control scheme, the system duty cycle for the transmitter can be controlled. On the receiver end, the system duty cycle can also be controlled as part of a power scheme according to an embodiment of the present invention. In an embodiment, receivers from different devices can have different duty cycles. Receivers from different devices can communicate with each other and coordinate such that the receivers reduce the amount of the duty cycle where no receiver is receiving power and/or the more than one, or more than two, receivers are receiving power, so as to enhance the charging of the two or more devices.

In one embodiment, the system duty cycle of a receiver can be controlled using, for example, discrete rectifying diodes **30** with a voltage regulator **40**. For example, the voltage regulator part LM5574 for 500mA output current and part LM5576 for 3A output current can be used. In another example, voltage regulation can be performed by a microprocessor having an RF receiver such as the Chipcon CC2510F8.

Figure 2 shows a power control scheme according to an embodiment of the present invention. As shown in Figure 2, the transmitter (e.g., reference **20** of Figure 1) excites the coil (e.g., reference **50** of Figure 1) for a short period of time (X/Y seconds) to access a pad's load condition so as to determine if any device (e.g., reference **60** of Figure 1) is on or near the pad and to power down for X seconds if no load is detected. Values of X and Y are determined by the settling time of the system and the response time requirement once the receiver(s) is/are placed on the pad. In a specific embodiment, X can be about one second and X/Y can be from about 500 μsec to about 1 msec, so as to promptly start charging a device placed in appropriate position and avoid wasting power.

The load condition can be monitored, in order to determine when a device to be charged is put in an appropriate position by, for example, comparing one or more of the following with values that occur when no load is present: transmitter supply current, transmitter coil voltage, and transmitter coil current. The transmitter is also able to detect a device on the pad via a near field wireless communication link (e.g., reference **70** of Figure 1) if there is still charge left in the receiver's battery. However, if there is not sufficient charge left in the battery, the transmitter can power up at full 100% duty cycle to transmit power to the power depleted receiver(s). In another embodiment, the transmitter is powered up to at least 50% duty cycle. Upon receiving sufficient power to power the near field communication module, the receiver can then transmit its power requirement to the transmitter. Based on the power requirement of the receiver(s), the transmitter can then determine a duty cycle that is, for example, sufficient to power the most power hungry device without wasting extra energy. In specific embodiments, the selected duty cycle can optimize the efficiency of charging.

When there is a single device to be charged the transmitter can detect the presence of the device to be charged by detecting load conditions, and can then adjust the duty cycle of the transmitter appropriately. In embodiments having two or more devices to be charged the transmitter can detect the devices to be charged by detecting load conditions and then receiving communications from the receiver(s) of the device(s) to be charged as to the charging requirements. In a specific embodiment, the devices to be charged can communicate with each other as to what portions of the transmitters duty cycle to receive power.

The receiver can determine the transmitter duty cycle. In a specific embodiment, the receiver can determine the transmitter duty cycle via the envelope detect technique. In addition, clock synchronization can also be performed during the process of determining the transmitter duty cycle. Depending on the power requirement of the receiver, the receiver can transmit its requirement if insufficient power is being transmitted or perform further duty cycling on the transmitter duty cycled power transmission to obtain the required power level. In a specific embodiment, a device placed on charging pad is recognized by the transmitter as a change in load. If a battery is drained, then the near field communication module is powered-up using power from the transmitter.

Figure 3 shows a duty cycle scheme according to an embodiment of the present invention. In a specific embodiment, a duty cycle scheme for a 100 W per unit area transmitter is provided with examples of three receivers' per unit area power requirement of 75 W, 25 W and 5 W respectively. The shaded portions show when the transmitter is transmitting or when the receivers are receiving, while the dark line shows when the receivers are on. The transmitter is shown as having a 75% duty cycle, as turning the transmitter off for a portion of the cycle can be helpful. Other percentage duty cycles can also be used. Each receiver can determine its duty cycle and turn on charging in accordance with its power requirement. Other than the most power hungry device, which is reflected in the scheme shown by receiver 1 in Figure 3, all other receivers (receiver 2 and receiver 3 of Figure 3) can stop receiving power by disengaging the coil during the beginning of the next cycle of power transmission from the transmitter. Each individual receiver can continue to monitor its power requirement and make necessary adjustments to ensure efficient power transfer. The scheme is capable of seamlessly letting any device receive power even if there are other devices under charge.

In another specific embodiment, the duty cycling scheme can be optimized as shown in Figure 4. The difference between the scheme illustrated by Figure 4 and the scheme illustrated by Figure 3 is the change in the window position of receiver 2. This change in window position can provide improved distribution of power transmission between lower power devices by reducing overlapping of receivers being on. However, synchronization between receivers can be important. If desired, the receivers can be turned off while the transmitter is off. In addition, the window of when the receivers are turned on can be moved to reduce the amount of overlapping of receiver charging in order to further improve charging efficiency.

Then, once all the devices are fully charged, which can be determined via feedback from the receiver(s), the transmitter can be powered off or its duty cycle can be reduced to performance trickle charging. The transmitter can continue to access the load conditions while in standby to detect any new device being placed on the transmitting pad.

Through using a duty cycle scheme according to embodiments of the present invention, it is possible to provide a high efficiency and low cost wireless power charging platform that is capable of concurrent charging/powering of one or more portable devices such as mobile phones, mp3 players, and laptop computers. Embodiments of the present invention can be implemented, for example, at home, at airports, hotel rooms, or any public location. Accordingly, this would bring great convenience to general consumers, especially frequent travelers, as it would provide a universally charging interface and eliminate the need to carry multiple chargers. In addition, the user is able to charge all of his portable devices at the same time without significant impact on charge time.

All patents, patent applications, provisional applications, and publications referred to or cited herein are incorporated by reference in their entirety, including all figures and tables, to the extent they are not inconsistent with the explicit teachings of this specification.

It should be understood that the examples and embodiments described herein are for illustrative purposes only and that various modifications or changes in light thereof will be suggested to persons skilled in the art and are to be included within the spirit and purview of this application.

CLAIMS

What is claimed is:

1. A method for wireless power transmission, comprising:

(a) exciting a transmitter coil of a transmitter of a wireless power transmission system for X/Y seconds, wherein the transmitter coil transmits an RF magnetic field when the transmitter coil is excited;

(b) determining a load condition so as to determine if a device is positioned to be charged, wherein each device to be charged comprises a receiver coil inductively coupled to the transmitter coil when the device is positioned to be charged and a receiver for receiving a receiver current from the receiver coil, wherein if there is no device positioned to be charged, then after a delay of X seconds returning to (a);

(c) if there is a device positioned to be charged, then exciting the transmitter coil so as to transmit power to the device positioned to be charged.

2. The method according to claim 1, further comprising receiving at the transmitter a power requirement for the device positioned to be charged provided by the receiver of the device positioned to be charged.

3. The method according to claim 2, wherein when there are two or more devices positioned to be charged, receiving at the transmitter the power requirement for each device positioned to be charged provided by the receiver of each device positioned to be charged.

4. The method according to claim 2, wherein the receiver comprises a near field communication module for providing the power requirement, wherein when a battery of the device positioned to be charged is not sufficiently charged to power the near field communication module of the receiver to provide the power requirement the receiver of the device positioned to be charged provides the power requirement to the transmitter upon receiving sufficient power from the transmitter to power the near field communication module of the receiver.

5. The method according to claim 3, further comprising:
detecting at the receiver a transmitter duty cycle.

6. The method according to claim 5, wherein the transmitter duty cycle is detected by using an envelope detect technique.

7. The method according to claim 5, wherein each of the two or more devices positioned to be charged detects the transmitter duty cycle and charges according to the power requirement of the device.

8. The method according to claim 1, wherein X is less than or equal to 1 second.

9. The method according to claim 1, wherein X/Y is in the range of 500 μsec to 1 msec.

10. The method according to claim 1, further comprising adjusting a duty cycle of the transmitter based on the load condition.

11. The method according to claim 2, further comprising adjusting a duty cycle of the transmitter based on the power requirement provided by the receiver of the device positioned to be charged.

12. The method according to claim 1, wherein determining the load condition comprises determining a transmitter supply current.

13. The method according to claim 1, wherein determining the load condition comprises determining a transmitter coil voltage.

14. The method according to claim 1, wherein determining the load condition comprises determining a transmitter coil current.

15. A system for wireless power transmission, comprising:

a transmitter having a transmitter coil, wherein the transmitter coil transmits an RF magnetic field when the transmitter coil is excited;

a device having a receiver and a receiver coil, wherein the receiver coil is inductively coupled to the transmitter coil when the device is positioned to be charged, wherein the receiver receives a receiver current from the receiver coil;

a means for exciting the transmitter coil for X/Y seconds; and

a means determining a load condition after exciting the transmitter coil for X/Y seconds so as to determine if the device is positioned to be charged, wherein if the device is not positioned to be charged, then after a delay of X seconds initiating the means for exciting the transmitter coil for X/Y seconds, wherein if the device is positioned to be charged, then exciting the transmitter coil so as to transmit power to the device positioned to be charged.

16. The system according to claim 15, wherein the receiver further comprises a means for transmitting to the transmitter a power requirement for the device.

17. The system according to claim 16, wherein the system comprises two or more devices.

18. The system according to claim 16, wherein the means for transmitting to the transmitter the power requirement comprises a near field communication module for providing the power requirement, wherein when a battery of the device is not sufficiently charged to power the near field communication module to provide the power requirement the receiver of the device positioned to be charged provides the power requirement to the transmitter upon receiving sufficient power from the transmitter to power the near field communication module of the receiver.

19. The system according to claim 17, wherein at least one device further comprises:

a means for detecting a transmitter duty cycle.

20. The system according to claim 19, wherein the transmitter duty cycle is detected by using an envelope detect technique.

21. The system according to claim 19, wherein each of the two or more devices detect the transmitter duty cycle and charges according to the power requirement of the device.

22. The system according to claim 15, wherein X is less than or equal to 1 second.

23. The system according to claim 15, wherein X/Y is in the range of 500 μsec to 1 msec.

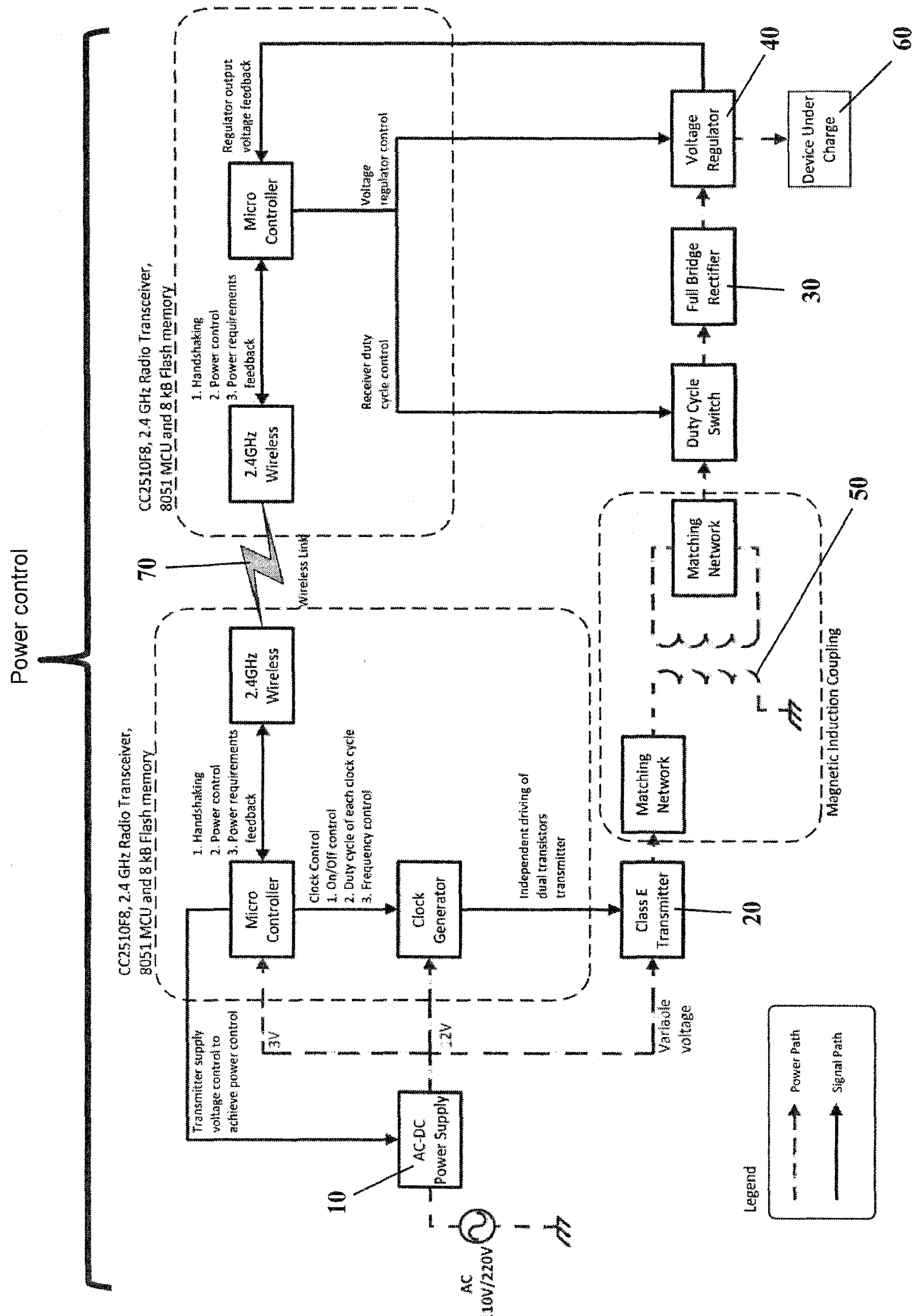
24. The system according to claim 15, wherein the transmitter further comprises a means for adjusting a duty cycle of the transmitter based on the load condition.

25. The system according to claim 16, wherein the transmitter further comprises a means for adjusting a duty cycle of the transmitter based on the power requirement provided by the receiver of the device.

26. The system according to claim 15, wherein the means for determining the load condition comprises a means for determining a transmitter supply current.

27. The system according to claim 15, wherein the means for determining the load condition comprises a means for determining a transmitter coil voltage.

28. The system according to claim 15, wherein the means for determining the load condition comprises a means for determining a transmitter coil current.



Power control
CC2510F8, 2.4 GHz Radio Transceiver, 8051 MCU and 8 kB Flash memory
Micro Controller
2.4GHz Wireless
1. Handshaking
2. Power control
3. Power requirements feedback
Clock Control
1. On/Off control
2. Duty cycle of each clock cycle
3. Frequency control
Clock Generator
Independent driving of dual transistors transmitter
Class E Transmitter
Transmitter supply voltage control to achieve power control
AC-DC Power Supply
AC 110V/220V
3V
12V
Variable voltage
Matching Network
Magnetic Induction Coupling
Wireless Link
Regulator output voltage feedback
Voltage regulator control
Receiver duty cycle control
Duty Cycle Switch
Full Bridge Rectifier
Voltage Regulator
Device Under Charge
Legend
Power Path
Signal Path
10
20
30
40
50
60
70

FIG. 1

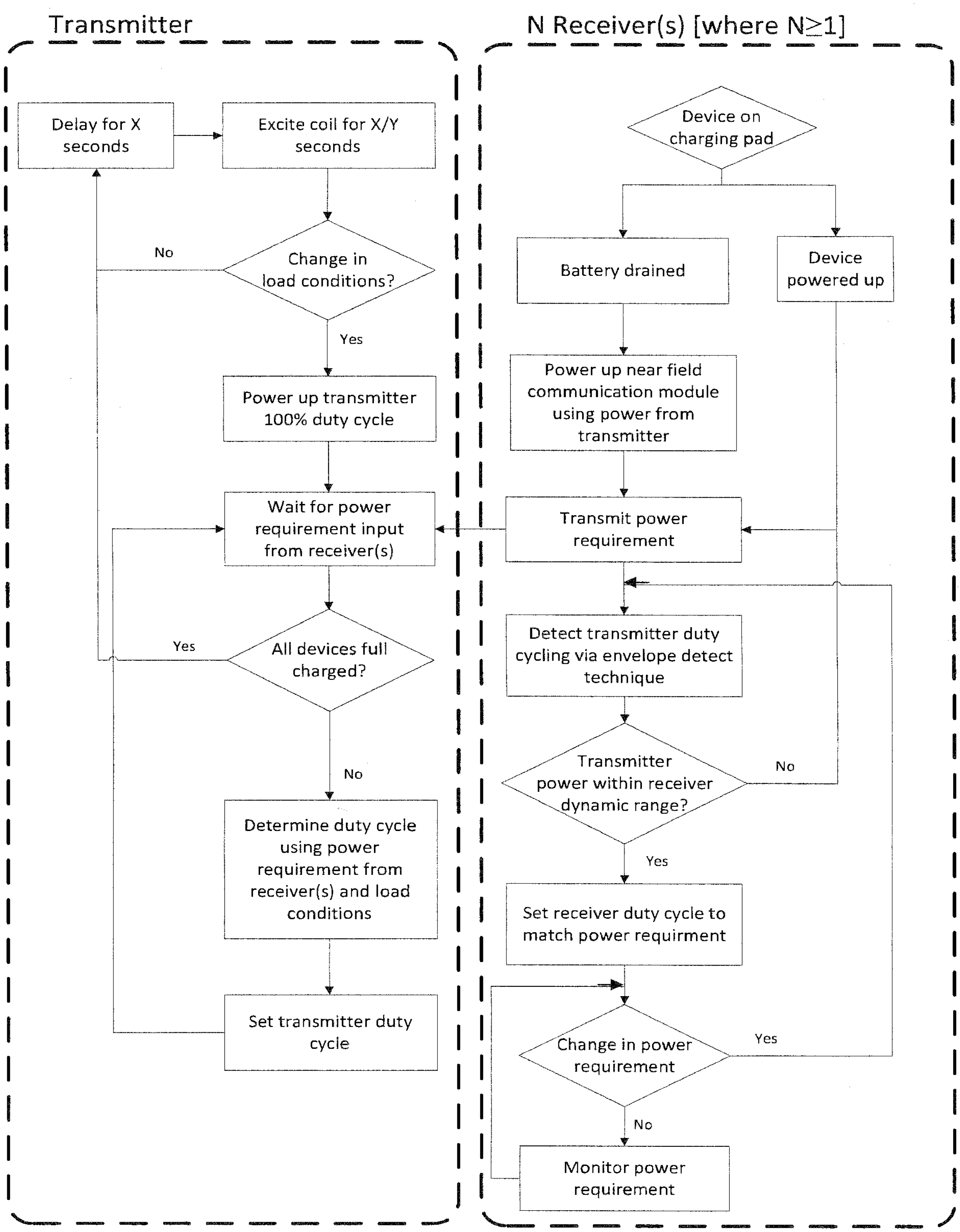

Transmitter
N Receiver(s) [where N≥1]
Delay for X seconds
Excite coil for X/Y seconds
Change in load conditions?
No
Yes
Power up transmitter 100% duty cycle
Wait for power requirement input from receiver(s)
All devices full charged?
Yes
No
Determine duty cycle using power requirement from receiver(s) and load conditions
Set transmitter duty cycle
Device on charging pad
Battery drained
Device powered up
Power up near field communication module using power from transmitter
Transmit power requirement
Detect transmitter duty cycling via envelope detect technique
Transmitter power within receiver dynamic range?
No
Yes
Set receiver duty cycle to match power requirment
Change in power requirement
Yes
No
Monitor power requirement

FIG. 2

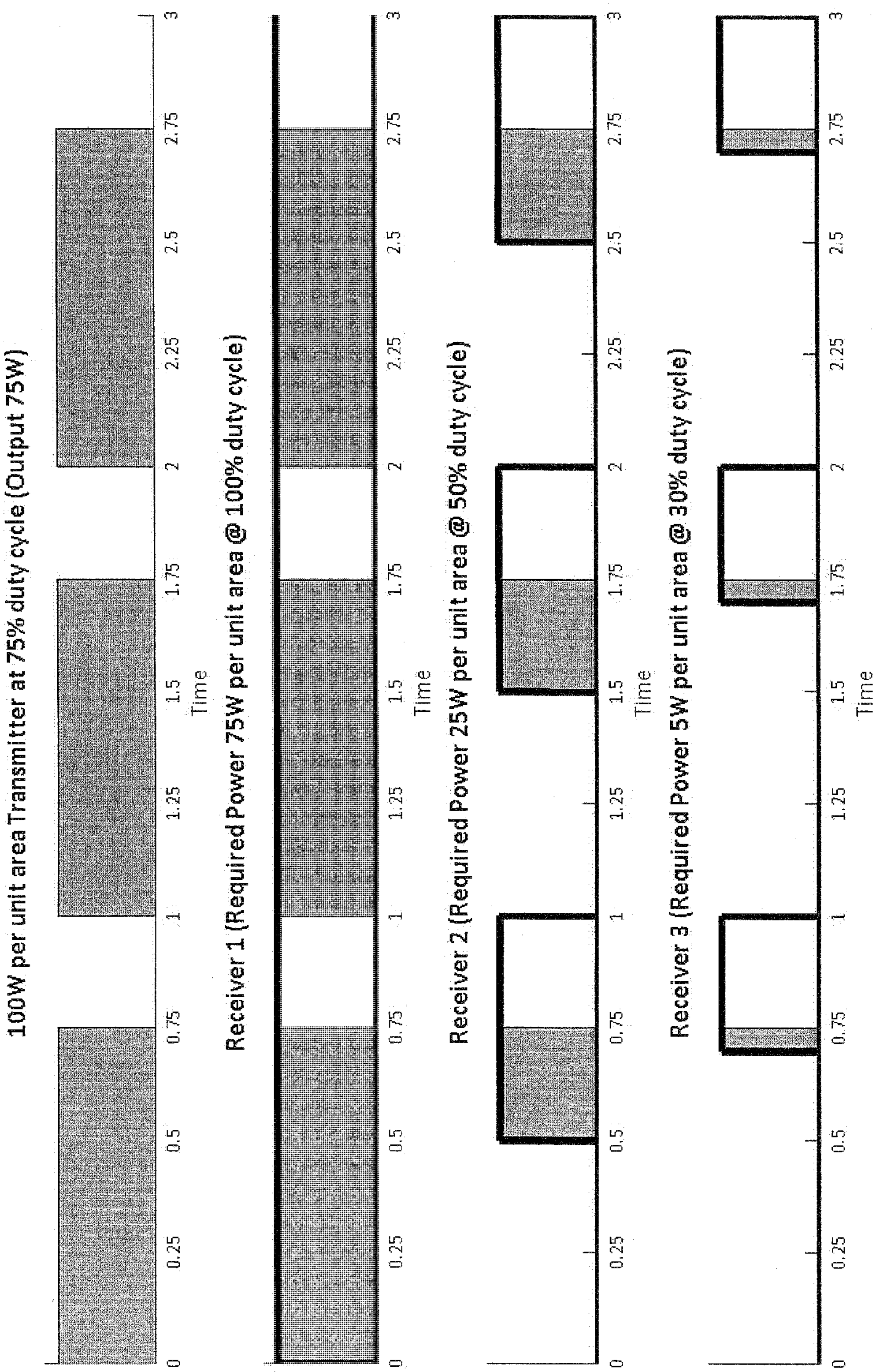

100W per unit area Transmitter at 75% duty cycle (Output 75W)
Receiver 1 (Required Power 75W per unit area @ 100% duty cycle)
Receiver 2 (Required Power 25W per unit area @ 50% duty cycle)
Receiver 3 (Required Power 5W per unit area @ 30% duty cycle)
0
0.25
0.5
0.75
1
1.25
1.5
1.75
2
2.25
2.5
2.75
3
Time

FIG. 3

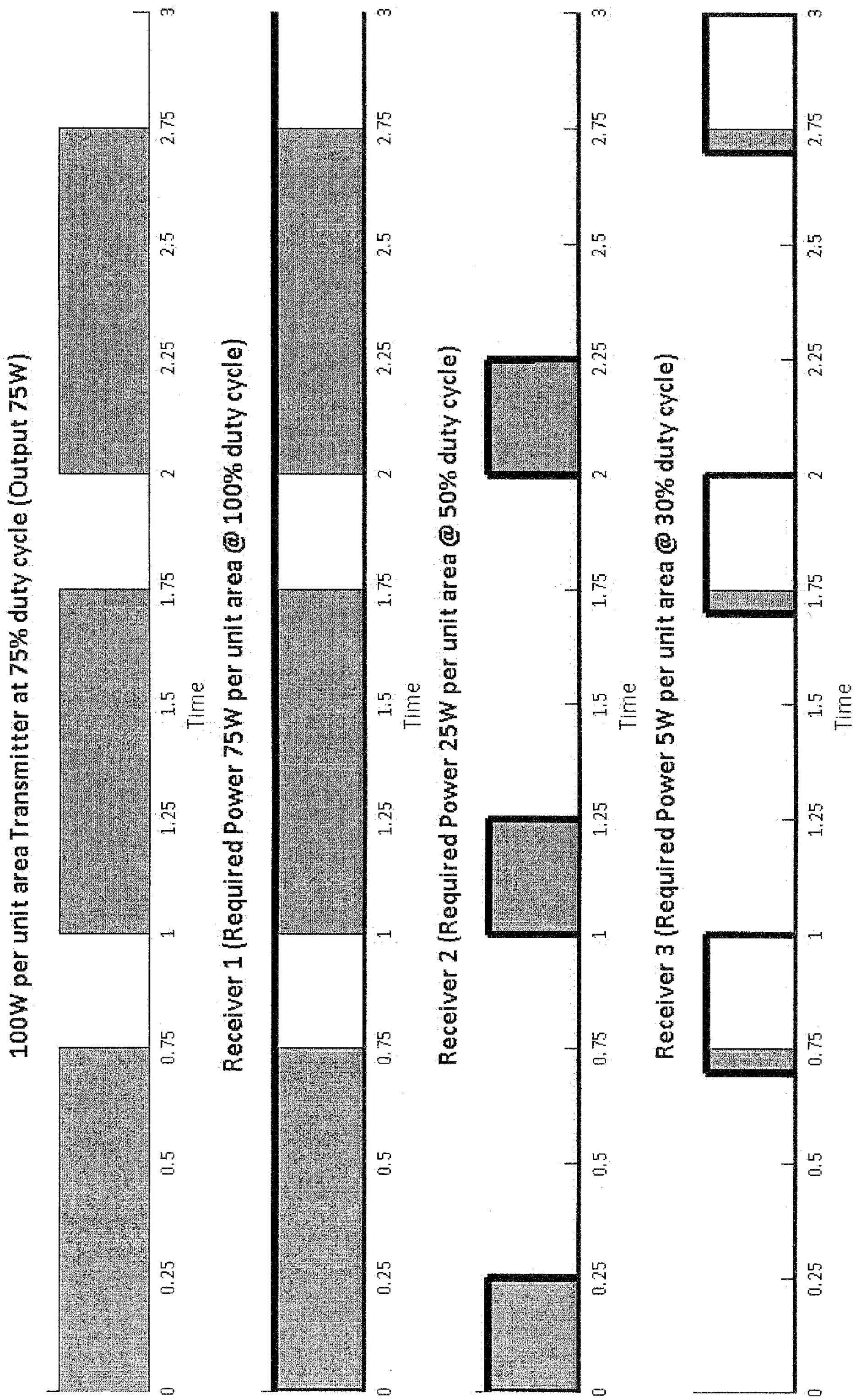

100W per unit area Transmitter at 75% duty cycle (Output 75W)
0 0.25 0.5 0.75 1 1.25 1.5 1.75 2 2.25 2.5 2.75 3
Time
Receiver 1 (Required Power 75W per unit area @ 100% duty cycle)
0 0.25 0.5 0.75 1 1.25 1.5 1.75 2 2.25 2.5 2.75 3
Time
Receiver 2 (Required Power 25W per unit area @ 50% duty cycle)
0 0.25 0.5 0.75 1 1.25 1.5 1.75 2 2.25 2.5 2.75 3
Time
Receiver 3 (Required Power 5W per unit area @ 30% duty cycle)
0 0.25 0.5 0.75 1 1.25 1.5 1.75 2 2.25 2.5 2.75 3
Time

FIG. 4

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(54) Title: METHOD AND APPARATUS FOR PRODUCING SUBSTANTIALLY UNIFORM MAGNETIC FIELD

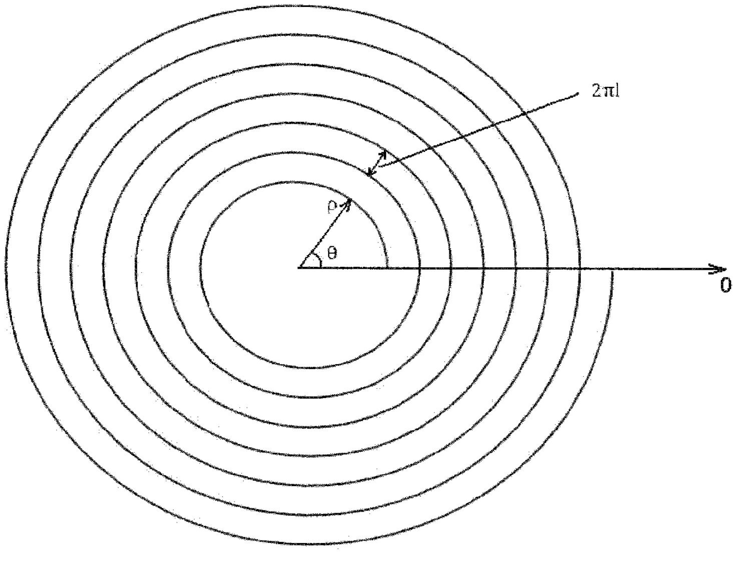

FIG. 1

(57) Abstract: A planar wireless power transmitter coil design and method are provided. A single spiral coil can be used to provide a uniform magnetic field across its surface area for location-independent planar wireless power charging. The spiral coil can be designed to have a non- constant gap between adjacent loops such that the gap between adjacent loops decreases towards the outer loops.

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DESCRIPTION

METHOD AND APPARATUS FOR PRODUCING SUBSTANTIALLY UNIFORM MAGNETIC FIELD

CROSS-REFERENCE TO RELATED APPLICATION

The present application claims the benefit of U.S. Provisional Application Serial No. 61/056,354, filed May 27, 2008, which is hereby incorporated by reference herein in its entirety, including any figures, tables, or drawings.

BACKGROUND OF INVENTION

In recent years, consumer electronics devices such as cell phones, personal digital assistants (PDAs), and laptops are using more wireless components such as a Bluetooth headset, wireless mouse, and wireless LAN. However, the wired power line remains to impair wireless freedom. Many designs and research has been conducted to provide solutions to get rid of this wire. Inductive wireless power transmission appears to be the most promising solution to this problem.

In wireless power charging, AC current passes through a transmitter coil, inducing magnetic flux on and/or above the surface of a power platform. A receiver coil generates voltage when magnetic flux passing through the receiver coil's loop(s) changes. In many cases, the transmitter coil and the receiver coil are not of the same size.

However, when the transmitter coil and receiver coil have significantly different sizes, the voltage generated on the receiving side can be greatly affected by the receiver coil's placement on the surface of the transmitter coil.

Specifically, a typical transmitter coil has a non-uniform magnetic field across its surface area, which may cause voltage variation and impedance matching difficulty.

A normal spiral coil as shown in Figure 1 usually has constant gap between adjacent loops. For example, the circular spiral coil of Figure 1 follows the equation of

$$\rho = \rho_0 + l\theta . \qquad (1)$$

where ρ is the radius, θ is the angle, ρ_0 is the initial radius and l is a constant. In Figure 1, the distance between adjacent wires is a constant 2pl.

The magnetic field near the surface for the coil shown in Figure 1 is shown in Figure 2, which illustrates a high magnetic field strength at the center of the coil. Each loop of the coil contributes magnetic field in the area it encloses, and the magnetic field in the center is the superposition of magnetic field contributions from all the loops.

For a regular coil, the density of magnetic flux generated in the coil has a maximum value at a position closest to the coil, and has a minimum value at a position at the center of the coil. Thus, the charging efficiency may be abruptly deteriorated leading to significant variation in charging efficiency.

One approach to solve this problem is discussed in WO2007/013725A1 (Gwon *et al.*), which discloses a wireless charger having decreased variation of charging efficiency. According to Gwon *et al.*, a smaller coil is placed in the center of an outer coil, which reinforces the magnetic flux density in the center of the outer and inner coils. Thus the entire magnetic flux density is flattened as a whole in comparison to the magnetic flux density formed by only the outer coil. Though the design disclosed in Gwon *et al.* reduces the effect of variation of the magnetic flux density of the outer coil, the variation can still be significant. In addition, when the receiver coil is much smaller than the transmitter coil, the location of the receiver coil can often affect its performance.

In a similar approach, WO2007/019806A1 (Hui *et al.*) provides a design of an auxiliary winding for improved performance of a planar inductive charging platform. According to Hui *et al.* an auxiliary winding is introduced to compensate the magnetic field generated by a principle winding. The design taught by Hui *et al.* uses a similar mechanism as that taught by Gwon *et al.* in that separate coils are used to improve charging efficiency.

Thus, there exists in the art a need for improved inductive wireless power transmission.

BRIEF SUMMARY

Embodiments of the subject invention relate to a method and apparatus for providing a planar spiral transmitter coil that produces a substantially uniform magnetic field over a region of interest near the surface of the coil. Embodiments of the invention provide a planar inductive wireless power transmission system incorporating a planar spiral transmitter coil and a receiver coil.

According to embodiments of the invention, a single coil design can provide improved charging efficiency to a wireless power transmission apparatus.

Coils in accordance with embodiments of the invention can provide for a system that uses near-field coupling to transfer power. Advantageously, embodiments of the invention do not require the alignment of the two axes of two coils. Certain embodiments of the invention provide improved robustness for wireless power transfer.

According to an embodiment of the invention, a single spiral coil can be used to provide a uniform magnetic field across the coil's surface area for location-independent planar wireless power charging. Embodiments of the invention generate a uniform magnetic field across an area that enables uniform wireless power transfer insensitive to the location of the device being charged.

In one embodiment, a circular spiral coil can be used. In another embodiment, a rectangular spiral coil can be used. Other shapes can also be utilized for the coil, such as elliptical, rectangular, hexagonal, and other polygonal shapes. The spiral coil can be designed to have a non-constant gap between adjacent loops such that the gap between adjacent loops decreases towards the outer loops.

BRIEF DESCRIPTION OF DRAWINGS

Figure 1 shows a normal spiral coil having a constant gap between adjacent loops.

Figure 2 shows a plot of the magnetic field of the normal spiral coil shown in Figure 1.

Figure 3 shows a spiral coil according to an embodiment of the invention.

Figure 4 shows a plot of the magnetic field of the spiral coil shown in Figure 3.

Figure 5 shows a rectangular spiral coil according to an embodiment of the invention.

Figure 6 shows a plot of the magnetic field of the rectangular spiral coil shown in Figure 5.

DETAILED DISCLOSURE

Embodiments of the subject invention relate to a spiral coil that can generate a substantially uniform magnetic field near the surface of the coil, across at least a portion of the surface area of the coil. Embodiments provide a location-independent planar wireless power charging system. Embodiments of the spiral coil can generate a substantially uniform magnetic field near the surface of the coil, across a portion of the surface area of the coil. A wireless power transmission system in accordance with an embodiment of the invention can have performance insensitive to the placement of the receiver coil within the substantially uniform magnetic field. The transmitter coil can be driven by a driver. In specific embodiments, the driver is a current source or a voltage source.

Specific embodiments can provide magnetic fields where the magnitude of the magnetic field in a direction perpendicular to the plane of the coil is substantially uniform over the region of interest such that $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.2 over the region of interest, where MAX and MIN are the maximum magnitude, and minimum magnitude, of the magnet field over the region of interest, respectively, and AVERAGE is $\frac{MAX + MIN}{2}$. Further specific embodiments can have the $\frac{MAX - MIN}{AVERAGE}$ of less than or equal to 0.1 over the region of interest and the $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.05 over the region of interest.

In an embodiment of a wireless power charging system, AC current passes through a transmitter coil, inducing magnetic flux on the surface of a power platform. In a specific embodiment, the frequency of the transmitter is between 1kHz and 10MHz. In a preferred embodiment, the frequency of the transmitter is in the range 100 kHz to 400 kHz, and in another embodiment, less than 1MHz. In specific embodiments, the region of interest can be a plane parallel to the plane of the coil offset from the plane of the coil by less than R, less than 30 cm, and/or less than 10 cm. The region of interest can cover a portion of, or all of the area of the coil. A receiver coil generates voltage when magnetic flux passing through the loop of the receiver coil changes. In specific embodiments, the transmitter coil and the receiver coil are not of the same size. A normal coil in accordance with an embodiment of the subject charging system can have a uniform magnetic field across its surface area, which reduces voltage variation and improves impedance matching. In a specific embodiment, the uniformity of the magnetic field can be less than 10% across the surface area of the coil, where the surface of the coil is the area enclosed by the outermost turn of the coil.

According to embodiments of the invention, to generate a more uniform field near the surface of the spiral coil, the distance between two adjacent loops can be adjusted. To reduce the magnetic field density at the center, the density of the inner loops should be less than the outer loops. In a specific embodiment, the gap between two adjacent coils decreases continuously toward the outer loops of the coil. A formula that describes the curve of a circular spiral inductor according to an embodiment of the invention is

$$\rho(\theta) = \rho_0 + l(\theta)\theta \tag{2}$$

where $l(\theta)$ is a function of θ. θ can vary from 0 to $2\pi N$, where N is the number of turns of the coil. In an embodiment, the derivative of $l(\theta)$ is positive and decreases as θ increases. Specific functions allow $l(\theta)$, the distance between adjacent loops, to be adjusted, and can allow the field across the surface of the coil to be substantially uniform. In another embodiment, the derivative of $l(\theta)$ is such that the spacing between adjacent loops can decrease or remain the same as θ increases such that as the coil moves from the innermost radius to the outermost radius the spacing decreases.

According to an embodiment of the invention, a circular spiral coil, which can be used to obtain the formula for $l(\theta)$, is

$$\rho(\theta) = r + (1 - (1 - \frac{\theta}{2\pi N})^4)(R - r) \tag{3}$$

where R is the outermost radius, r is the innermost radius, and N is the total turns of the coil. $l(\theta)$ can be obtained by setting the right side of equation (2) equal to the right side of equation (3) and solving $l(\theta)$, where r has the same meaning as ρ_0. According to one embodiment, r is 1/4 to 1/3 of R. In another embodiment, the coil can be elliptical with appropriate modifications to equations (2) and (3).

Figure 3 shows a coil with non-constant gap between adjacent loops, in accordance with an embodiment of the subject invention. The curvature of the spiral coil of Figure 3 follows equation 3, in which R = 200 mm, r = 50 mm, and N = 8. Figure 4 shows the magnetic field strength in a perpendicular direction across the surface area of the coil of Figure 3. As shown in Figure 4, the uniformity of the magnetic field for the coil of Figure 3 is significantly improved over the uniformity of the magnetic field for the coil of Figure 1.

For a rectangular spiral inductor, narrower gaps can be used between adjacent loops as the loops become farther from the center. According to an embodiment of the present invention, the gap between adjacent loops can be derived from

$$\rho(2n\pi) - \rho[2(n-1)\pi], \text{ n=1, 2,..., N.} \tag{4}$$

where ρ is the same function as Equation 3. Figure 5 shows a rectangular spiral coil according to an embodiment of the invention. The design of the coil of Figure 5 follows equation 4, in which $R = 200$ mm, $r = 50$ mm, and $N = 8$mm. The magnetic field strength in a perpendicular direction for the coil of Figure 5 is shown in Figure 6.

The results, as shown in Figures 4 and 6, demonstrate that a substantially uniform magnetic field of spiral coil can be generated in accordance with embodiments of the invention.

Additional embodiments utilize polygonal coils, such as rectangles, squares, hexagons, and other multisided shapes, to produce the magnetic fields. The spacing between adjacent coils can decrease or stay the same at each corner of the polygon such that the spacing decreases as the coil goes from the innermost radius to the outermost radius. In specific embodiments, the spacing can continuously decrease, the spacing can be the same between two corners (along one side of the polygonal) and decrease from before each corner to after each corner, the spacing can remain the same for a portion or all of a loop (as shown in Figure 5) and have decreases as the coil moves outward, and/or combinations of these changes.

In specific embodiments, a receiver coil can be inductively coupled to the transmitter coil so as to transfer power to the receiver coil. Embodiments can use receiver coils that have areas such that the transmitter coil area is 2 to 12 times as large, 2 to 8 times as large, or 2 to 4 times as large as the receiver coil area.

All patents, patent applications, provisional applications, and publications referred to or cited herein are incorporated by reference in their entirety, including all figures and tables, to the extent they are not inconsistent with the explicit teachings of this specification.

It should be understood that the examples and embodiments described herein are for illustrative purposes only and that various modifications or changes in light thereof will be suggested to persons skilled in the art and are to be included within the spirit and purview of this application.

CLAIMS

What is claimed is:

1. An apparatus for producing a magnetic field, comprising:

a coil, wherein the coil is a planar spiral coil, where the coil has at least two loops, wherein a spacing between adjacent loops decreases continuously from an inner loop toward an outer loop of the coil, and

a driver, wherein the driver drives the coil to produce a magnetic field, wherein a magnitude of the magnetic field in a direction perpendicular to a plane of the coil is substantially uniform over a region of interest.

2. The apparatus according to claim 1, wherein the magnitude of the magnetic field in a direction perpendicular to the plane of the coil is substantially uniform over the region of interest such that $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.2 over the region of interest, where MAX and MIN are the maximum magnitude, and minimum magnitude, of the magnet field over the region of interest, respectively, and AVERAGE is $\frac{MAX + MIN}{2}$.

3. The apparatus according to claim 2, wherein the $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.1 over the region of interest.

4. The apparatus according to claim 2, wherein the $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.05 over the region of interest.

5. The apparatus according to claim 1, wherein the magnetic field is time-varying.

6. The apparatus according to claim 5, wherein the time-varying magnetic field has a frequency in the range 1 kHz to 10 MHz.

7. The apparatus according to claim 5, wherein the time-varying magnetic field has a frequency in the range 100 kHz to 400 kHz.

8. The apparatus according to claim 5, wherein the time varying magnetic field has a frequency less than or equal to 1 MHz.

9. An apparatus according to claim 1, wherein the coil is a planar elliptical spiral coil.

10. The apparatus according to claim 1, wherein the coil is a planar circular spiral coil, wherein the coil follows the equation:

$$\rho(\theta) = \rho_0 + l(\theta)\theta$$

where $\rho(\theta)$ is the radius of the coil, ρ_0 is the initial radius of the coil, θ is the angle with respect to the initial radius of the coil, and $l(\theta)$ is a function of θ.

11. The apparatus according to claim 10, wherein a derivative of $l(\theta)$ is positive.

12. The apparatus according to claim 11, wherein the derivative of $l(\theta)$ decreases as θ increases over at least a portion of the coil.

13. The apparatus according to claim 12, wherein the derivative of $l(\theta)$ decreases as θ increases over the coil.

14. The apparatus according to claim 1, wherein the spiral coil further follows the equation:

$$\rho(\theta) = r + (1 - (1 - \frac{\theta}{2\pi N})^4)(R - r)$$

where R is an outermost radius of the coil, r is the initial radius of the coil, and N is a number of loops of the coil.

15. The apparatus according to claim 1, wherein the region of interest is a second plane parallel to a plane of the coil.

16. The apparatus according to claim 15, wherein the second plane is offset from the plane of the coil by a distance d.

17. The apparatus according to claim 16, wherein d is less than 30 cm.

18. The apparatus according to claim 16, wherein d is less than 10 cm.

19. The apparatus according to claim 16, wherein d is less than R, where R is an outermost radius of the coil.

20. The apparatus according to claim 15, wherein the region of interest is a region covering at least a portion of an area of the coil.

21. The apparatus according to claim 15, wherein the region of interest is a region covering an area of the coil.

22. The apparatus according to claim 1, wherein the coil is a polygonal spiral coil.

23. The apparatus according to claim 1, wherein the coil is a rectangular spiral coil.

24. An apparatus for producing a magnetic field, comprising:

a coil, wherein the coil is a planer polygonal spiral coil wherein the coil has at least two loops, wherein a spacing between adjacent loops either stays the same or decreases at each corner of the polygonal going from an inner loop toward an outer loop of the coil; and

a driver, wherein the driver drives the coil to produce a magnetic field, wherein a magnitude of the magnetic field in a direction perpendicular to a plane of the coil substantially is uniform over a region of interest.

25. The apparatus according to claim 24, wherein the magnitude of the magnetic field in a direction perpendicular to the plane of the coil is substantially uniform over the region of

interest such that $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.2 over the region of interest, where MAX and MIN are the maximum magnitude, and minimum magnitude, of the magnet field over the region of interest, respectively, and AVERAGE is $\frac{MAX + MIN}{2}$.

26. The apparatus according to claim 25, wherein the $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.1 over the region of interest.

27. The apparatus according to claim 25, wherein the $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.05 over the region of interest.

28. The apparatus according to claim 24, wherein the magnetic field is time-varying.

29. The apparatus according to claim 28, wherein the time-varying magnetic field has a frequency in the range 1 kHz to 10 MHz.

30. The apparatus according to claim 25, wherein the time-varying magnetic field has a frequency in the range 100 kHz to 400 kHz.

31. The apparatus according to claim 25, wherein the time varying magnetic field has a frequency less than or equal to 1 MHz.

32. The apparatus according to claim 24, wherein the spacing between adjacent loops follows the equation:

$\rho(2n\pi) - \rho[2(n-1)\pi]$, $n = 1, 2, \dots$, N, where ρ is the function

$$\rho(\theta) = r + (1 - (1 - \frac{\theta}{2\pi N})^4)(R - r),$$

where R is an outermost radius of the coil, r is an innermost radius of the coil, and N is a number of loops of the coil.

33. The apparatus according to claim 24, wherein the region of interest is a second plane parallel to the plane of the coil.

34. The apparatus according to claim 33, wherein the second plane is offset from the plane of the coil by a distance d.

35. The apparatus according to claim 34, wherein d is less than 30 cm.

36. The apparatus according to claim 34, wherein d is less than 10 cm.

37. The apparatus according to claim 34, wherein d is less than R, where R is an outermost radius of the coil.

38. The apparatus according to claim 33, wherein the region of interest is a region covering at least a portion of an area of the coil.

39. The apparatus according to claim 33, wherein the region of interest is a region covering an area of the coil.

40. The apparatus according to claim 24, wherein the polygonal coil is a square coil.

41. The apparatus according to claim 24, wherein the polygonal coil is a hexagonal coil.

42. An apparatus for producing a magnetic field, comprising:

a coil, wherein the coil is a planar spiral coil, where the coil has at least two loops, wherein a spacing between starting points of adjacent loops decreases from an inner loop toward an outer loop of the coil, and

a driver, wherein the driver drives the coil to produce a magnetic field, wherein a magnitude of the magnetic field in a direction perpendicular to the plane of the coil is substantially uniform over the region of interest such that $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.2 over the region of interest, where MAX and MIN are the maximum magnitude, and minimum magnitude, of the magnet field over the region of interest, respectively, and AVERAGE is $\frac{MAX + MIN}{2}$, wherein a magnitude of the magnetic field in a direction perpendicular to a plane of the coil is substantially uniform over a region of interest.

43. A method for producing a magnetic field, comprising:

providing a coil, wherein the coil is a planar spiral coil, where the coil has at least two loops, wherein a spacing between adjacent loops decreases continuously from an inner loop toward an outer loop of the coil, and

driving the coil to produce a magnetic field, wherein a magnitude of the magnetic field in a direction perpendicular to a plane of the coil is substantially uniform over a region of interest.

44. The method according to claim 43, wherein the magnitude of the magnetic field in a direction perpendicular to the plane of the coil is substantially uniform over the region of interest such that $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.2 over the region of interest, where MAX and MIN are the maximum magnitude, and minimum magnitude, of the magnet field over the region of interest, respectively, and AVERAGE is $\frac{MAX + MIN}{2}$.

45. The method according to claim 44, wherein the $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.1 over the region of interest.

46. The method according to claim 44, wherein the $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.05 over the region of interest.

47. The method according to claim 44, wherein the magnetic field is time-varying.

48. The method according to claim 47, wherein the time-varying magnetic field has a frequency in the range 1 kHz to 10 MHz.

49. The method according to claim 47, wherein the time-varying magnetic field has a frequency in the range 100 kHz to 400 kHz.

50. The method according to claim 47, wherein the time varying magnetic field has a frequency less than or equal to 1 MHz.

51. An method according to claim 43, wherein the coil is a planar elliptical spiral coil.

52. The method according to claim 43, wherein the coil is a planar circular spiral coil, wherein the coil follows the equation:

$$\rho(\theta) = \rho_0 + l(\theta)\theta$$

where $\rho(\theta)$ is the radius of the coil, ρ_0 is the initial radius of the coil, θ is the angle with respect to the initial radius of the coil, and $l(\theta)$ is a function of θ.

53. The method according to claim 52, wherein a derivative of $l(\theta)$ is positive.

54. The method according to claim 53, wherein the derivative of $l(\theta)$ decreases as θ increases over at least a portion of the coil.

55. The method according to claim 54, wherein the derivative of $l(\theta)$ decreases as θ increases over the coil.

56. The method according to claim 43, wherein the spiral coil further follows the equation:

$$\rho(\theta) = r + (1 - (1 - \frac{\theta}{2\pi N})^4)(R - r)$$

where R is an outermost radius of the coil, r is the initial radius of the coil, and N is a number of loops of the coil.

57. The method according to claim 43, wherein the region of interest is a second plane parallel to a plane of the coil.

58. The method according to claim 57, wherein the second plane is offset from the plane of the coil by a distance d.

59. The method according to claim 58, wherein d is less than 30 cm.

60. The method according to claim 58, wherein d is less than 10 cm.

61. The method according to claim 58, wherein d is less than R, where R is an outermost radius of the coil.

62. The method according to claim 57, wherein the region of interest is a region covering at least a portion of an area of the coil.

63. The method according to claim 57, wherein the region of interest is a region covering an area of the coil.

64. The method according to claim 43, wherein the coil is a polygonal spiral coil.

65. The method according to claim 43, wherein the coil is a rectangular spiral coil.

66. An method for producing a magnetic field, comprising:

producing a coil, wherein the coil is a planer polygonal spiral coil wherein the coil has at least two loops, wherein a spacing between adjacent loops either stays the same or decreases at each corner of the polygonal going from an inner loop toward an outer loop of the coil; and driving the coil to produce a magnetic field, wherein a magnitude of the magnetic field in a direction perpendicular to a plane of the coil substantially is uniform over a region of interest.

67. The method according to claim 66, wherein the magnitude of the magnetic field in a direction perpendicular to the plane of the coil is substantially uniform over the region of interest such that $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.2 over the region of interest, where MAX and MIN are the maximum magnitude, and minimum magnitude, of the magnet field over the region of interest, respectively, and AVERAGE is $\frac{MAX + MIN}{2}$.

68. The method according to claim 67, wherein the $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.1 over the region of interest.

69. The method according to claim 67, wherein the $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.05 over the region of interest.

70. The method according to claim 66, wherein the magnetic field is time-varying.

71. The method according to claim 70, wherein the time-varying magnetic field has a frequency in the range 1 kHz to 10 MHz.

72. The method according to claim 67, wherein the time-varying magnetic field has a frequency in the range 100 kHz to 400 kHz.

73. The method according to claim 67, wherein the time varying magnetic field has a frequency less than or equal to 1 MHz.

74. The method according to claim 66, wherein the spacing between adjacent loops follows the equation:

$\rho(2n\pi) - \rho[2(n-1)\pi]$, $n = 1, 2, \ldots, N$, where ρ is the function

$$\rho(\theta) = r + (1 - (1 - \frac{\theta}{2\pi N})^4)(R - r),$$

where R is an outermost radius of the coil, r is an innermost radius of the coil, and N is a number of loops of the coil.

75. The method according to claim 66, wherein the region of interest is a second plane parallel to the plane of the coil.

76. The method according to claim 75, wherein the second plane is offset from the plane of the coil by a distance d.

77. The method according to claim 76, wherein d is less than 30 cm.

78. The method according to claim 76, wherein d is less than 10 cm.

79. The method according to claim 76, wherein d is less than R, where R is an outermost radius of the coil.

80. The method according to claim 75, wherein the region of interest is a region covering at least a portion of an area of the coil.

81. The method according to claim 75, wherein the region of interest is a region covering an area of the coil.

82. The method according to claim 66, wherein the polygonal coil is a square coil.

83. The method according to claim 66, wherein the polygonal coil is a hexagonal coil.

84. An method for producing a magnetic field, comprising:

providing a coil, wherein the coil is a planar spiral coil, where the coil has at least two loops, wherein a spacing between starting points of adjacent loops decreases from an inner loop toward an outer loop of the coil, and

driving the coil to produce a magnetic field, wherein a magnitude of the magnetic field in a direction perpendicular to the plane of the coil is substantially uniform over the region of interest such that $\frac{MAX - MIN}{AVERAGE}$ is less than or equal to 0.2 over the region of interest, where MAX and MIN are the maximum magnitude, and minimum magnitude, of the magnet field over the region of interest, respectively, and AVERAGE is $\frac{MAX + MIN}{2}$, wherein a magnitude of the magnetic field in a direction perpendicular to a plane of the coil is substantially uniform over a region of interest.

85. A system for inductive power transfer, comprising:

an apparatus for producing a magnetic field according to any of claims 1-42; and

a receiver coil, wherein when the receiver coil is positioned proximate the apparatus for producing the magnetic field, power is inductively transfer to the receiver coil.

86. The system according to claim 85, wherein the coil has an area in the range of 2 to 12 times as large as an area of the receiver coil.

87. A method for inductively transferring power, comprising:

implementing the method according to any of claims 43-84; and

providing a receiver coil proximate to the coil such that power is inductively coupled to the receiver coil.

88. The method according to claim 87, wherein the coil has an area in the range of 2 to 12 times as large as an area of the receiver coil.

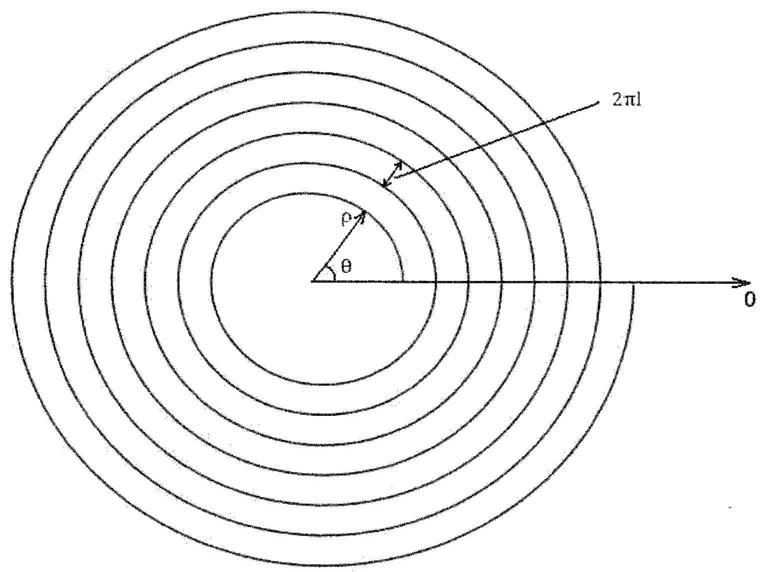
2πl
ρ
θ
0

FIG. 1

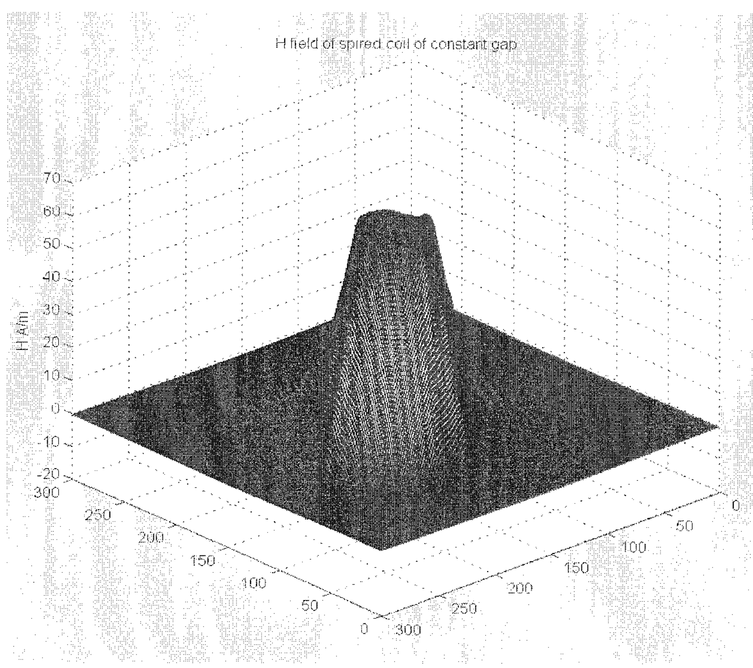
H field of spired coil of constant gap
H A/m
70
60
50
40
30
20
10
0
-10
-20
300
250
200
150
100
50
0
0
50
100
150
200
250
300

FIG. 2

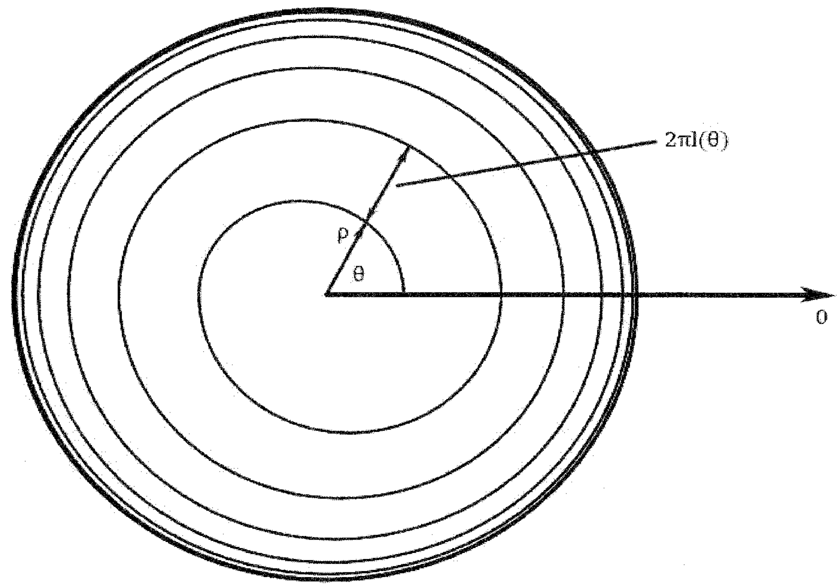

FIG. 3

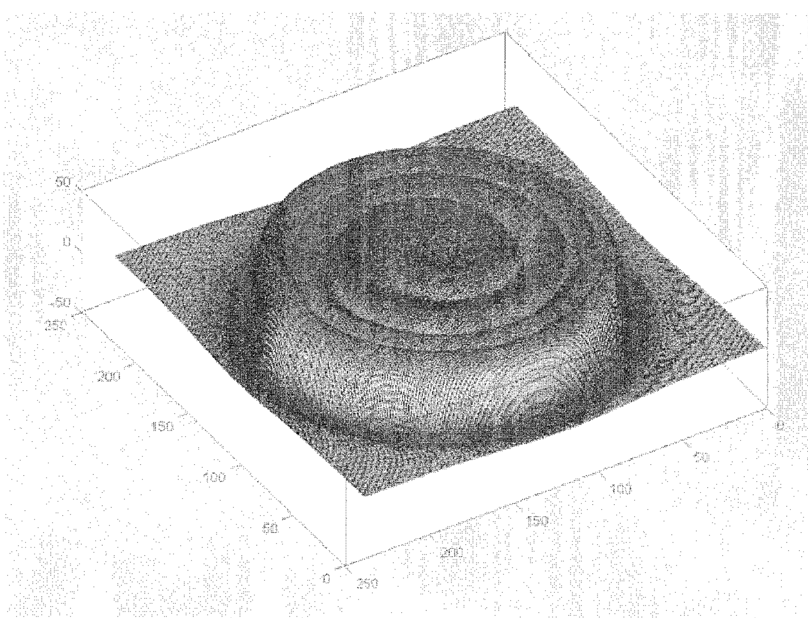

FIG. 4

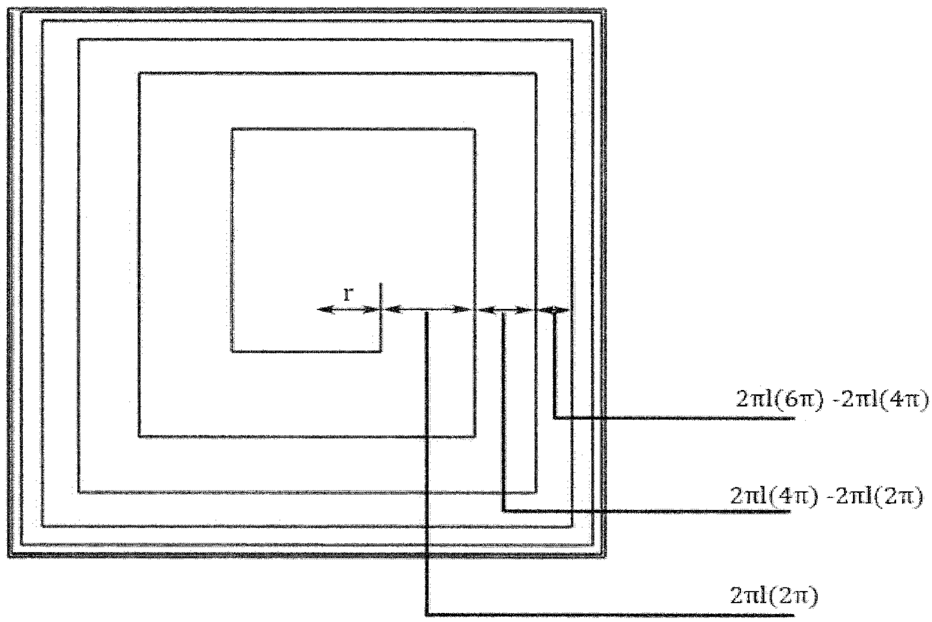
r
2πl(6π) -2πl(4π)
2πl(4π) -2πl(2π)
2πl(2π)

FIG. 5

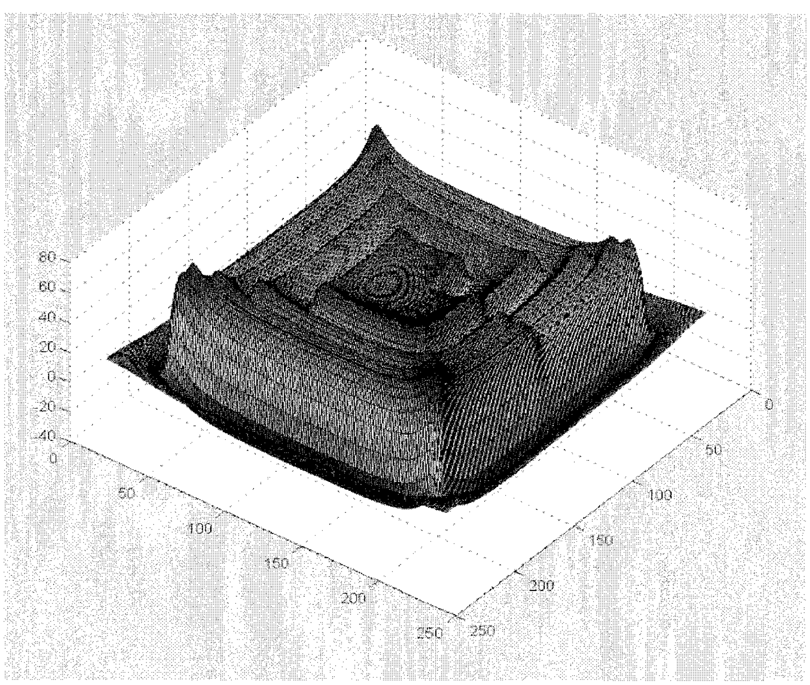
80
60
40
20
0
-20
-40
0
50
100
150
200
250
0
50
100
150
200
250

FIG. 6

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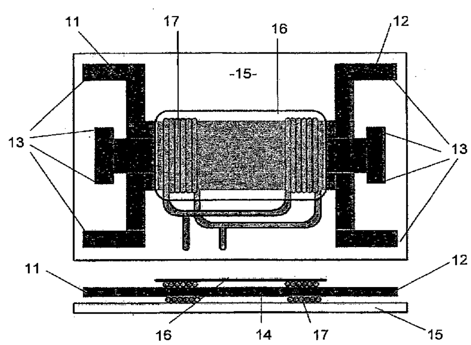

FIGURE 8

(57) Abstract: A magnetic flux pad for generating or receiving magnetic flux has two pole areas (11, 12), a permeable core (14) and a coil (16) wound about the core. The pad allows useable flux to be generated at a significant height above a surface of the pad.

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INDUCTIVE POWER TRANSFER APPARATUS

Field of the Invention

This invention relates to apparatus for generating or receiving magnetic flux. The invention has particular, but not sole, application to a low profile, substantially flat device, such as a pad, for power transfer using an Inductive Power Transfer (IPT) system.

Background

IPT systems, and use of a pad including one or more windings that may comprise the primary or secondary windings for inductive power transfer, are introduced in our published international patent application WO 2008/140333, the contents of which are incorporated herein by reference. One particular application of IPT power transfer pads is electric vehicle charging. IPT power transfer pads are used both in the vehicle as a power "pick-up" device (i.e. the secondary side winding of the IPT system), and at a stationary location such as a garage floor as the "charging pad" (i.e. the primary side winding) from which power is sourced.

In the development of pick-ups for inductively charging electric vehicles a problem of some concern is the clearance available under the vehicle. With conventional pick-up circuits power in sufficient quantities can be provided at distances up to perhaps 100 mm at which time the coupling factor becomes so small that it becomes impractical.

It is generally conceded that the power required to charge a typical electric vehicle overnight is about 2.0 kW, so that in an overnight charging mode some 24 kWH can be transferred. With modern electric vehicles this is enough energy to travel more than 100 km and is ideal for small vehicles used for tasks such as dropping children at schools, running errands, short commutes and the like.

Inductively coupled chargers commonly use two power pads that are circular in shape and may have dimensions of 400 mm diameter by 25 mm thick as shown in Figure 1. However, to use an inductive charger such as this the vehicle must be positioned relatively accurately over the charging pad - typically within 50 mm of perfect alignment - and the separation between the power pad on the vehicle and the power pad on the ground must be closely controlled. In principle inductive power transfer may be accomplished for vertical spacings between 0 mm and 100 mm but if the system is set up for 100 mm it will have quite a large reduction in power at

120 mm and will be inoperable below 50 mm. This state of affairs occurs because both the self inductance and the mutual inductance of the power pads vary widely as the distance between the pads changes. The self inductance and the mutual inductance as a function of the separation for two identical circular pads that have the construction of Figure 1, are shown in Figure 2. Thus at 100 mm the power pad receiver or pick-up may have a pick-up voltage of 100 V and a short circuit current of 5.0 A for a power rating of 500 W. If the IPT system electronics operates with a Q factor of 4, then 2 kW can be transferred to the battery though there are still difficulties to overcome in producing the power needed at the appropriate battery voltage.

The induced voltage in the pick-up pad (i.e. the vehicle mounted power pad) is very separation sensitive – corresponding to the variation in mutual inductance shown in Figure 2 - so that at 120 mm it is reduced by approximately 40% while at 50 mm it is increased by a factor of 2. A reduction in power means that the vehicle does not get fully charged in the usual time, but the more challenging situation is that at smaller separations the power transferred may be so high that the components of the circuit are overloaded. Also, as the separation is reduced the self inductance of the pick-up coil also changes so that the circuit operates off-frequency putting extra stress on the power supply. As the separation gets smaller still this stress on the power supply caused by the non-tuned pick-up on the primary side cannot be sustained and the system must be shut down. In practice it is feasible to operate with a separation between 40 and 100 mm but a larger range is too difficult.

A range of separation from 40 to 100 mm is quite small. If the vehicle has a relatively high ground clearance then either the power pad on the vehicle has to be lowered or the power pad on the ground has to be raised. Automatic systems for doing this compromise the reliability of the charging system. Alternatively the pad on the ground can be on a fixed but a raised platform but such a pad is a tripping hazard when a car is not being charged and this situation is generally to be avoided in a garage or other location involving vehicles and pedestrians.

The known power pad construction of Figure 1 comprises an aluminium case 1 containing typically eight ferrite bars 2 and a coil 3. Current in the coil causes magnetic flux in the ferrite bars and this flux has flux lines that start on the ferrite bars and propagate to the other end of the bar in a path containing the coil that may be thought of as a semi-elliptical shape. The flux lines 4 for a single bar are shown in Figure 3. The flux lines leave the ferrite in an upward direction and propagate to the other end of the bar, entering it at right angles. No flux goes out the back of the pad as the solid aluminium backing plate 1 prevents it. In the actual pad the eight bars give a flux pattern shown approximately in cross section in Figure 4. A simulation of the actual flux pattern is shown in Figure 4A.

From Figure 4A it can be seen that at the highest point the flux lines are essentially horizontal. Therefore, to get the maximum separation possible between the primary pad and the secondary pad it would be advantageous to detect this horizontal flux. However, the horizontal flux is still relatively close to the pad (extending from the pad approximately one quarter of the diameter of the pad) and there is no horizontal flux at all at the very centre of the power pad. Thus at the very point where maximum flux density would be ideal – the centre – the actual usable horizontal flux component is zero.

Summary

It is an object of the invention to provide an improved apparatus or method for generating or receiving magnetic flux, or an improved IPT power transfer pad, or to at least provide a useful alternative.

Accordingly in one aspect the invention provides a magnetic flux pad having a front face and a back face for generating or receiving magnetic flux, the pad comprising:

two pole areas for sending or receiving flux;

a magnetically permeable core magnetically connecting the pole areas;

a coil wound about the core; and

whereby the flux enters the pad at one of the pole areas and exits the pad at the other pole area.

In some embodiments a flux shaping means is provided such that flux is directed into a space beyond the front face of the pad. The flux shaping means may be located adjacent to the back face of the pad and may advantageously comprise a member, such as a plate, constructed from a flux repelling material.

In some embodiments a flux shaping means is provided such that flux is substantially prevented from escaping from the core. The flux shaping means may comprise a flux repelling member located adjacent to the front face of the pad. It may further comprise a flux repelling member located adjacent to the rear face of the pad.

The coil may comprise a plurality of coils. The coils may be connected electrically in parallel and/or magnetically in series.

In another aspect the invention provides a magnetic flux pad having a front face and a back face for generating a magnetic flux in a space beyond the front face of the pad, the pad comprising:

two pole areas for sending or receiving flux;

a magnetically permeable core magnetically connecting the pole areas;

a coil wound about the core;

a flux repelling means provided adjacent to a rear face; and

whereby the flux enters the pad at one of the pole areas and exits the pad at the other pole area.

In a further aspect the invention provides an IPT power transfer pad including: a magnetic flux carrying member having a high magnetic permeability and two ends, each end being substantially adjacent a peripheral edge of the pad; and one or more windings provided about at least a part of the flux carrying member; and wherein said pad is configured such that magnetic flux exits or enters the flux carrying member substantially only at or adjacent to the ends.

In a still further aspect the invention provides an IPT system including a first magnetic flux pad or IPT power transfer pad for connection to a power supply and a second magnetic flux pad or IPT power transfer pad for connection to a load, the first and second magnetic flux pads or IPT power pads constructed according to any one of the above-described aspects and having one or more windings with the same number of turns, and wherein the number of turns is selected dependent on a required operating frequency

In another aspect the invention provides an IPT system including a magnetic flux pad according to any one of the preceding statements.

In some embodiments the system supplies power to an electric vehicle, such as an electric vehicle charging system.

Further aspects of the invention will become apparent from the following description.

Drawing Description

One or more embodiments are described below by way of example with reference to the accompanying drawings, in which:

Figure 1 is a perspective view of part of a known form of IPT power transfer pad;

Figure 2 is a graph of inductance measurements and flux linkage efficiency with respect to height displacement for a pad such as that of Figure 1;

Figure 3 is a diagrammatic elevation in cross section of a part of the pad of Figure 1 showing flux lines;

Figure 4 is a plan view and elevation of a cross section of the pad of Figure 1 showing flux lines;

Figure 4A is an elevation in cross section of a computer generated simulation of the magnetic field (indicated by flux lines) of the pad of Figure 1;

Figure 5 is a plan view and elevation in cross section of an embodiment of a new IPT power transfer pad;

Figure 6 is a diagrammatic view of the pad of Figure 5 showing one example of a winding arrangement;

Figure 7 is a diagrammatic elevation in cross section of the pad of Figure 5, and showing flux lines;

Figure 7A is an elevation in cross section of a computer generated simulation of the magnetic field (indicated by flux lines) of the pad of Figure 6;

Figure 8 is a plan view of another embodiment of a new pad based on the design of the pad of Figure 5;

Figure 9 is a graph of inductance measurements and flux linkage efficiency with respect to height displacement for a pad such as that of Figure 7;

Figure 10 is a graph of inductance measurements and flux linkage efficiency with respect to height displacement for both the pad of Figure 1 (referred to as the Circular pad) and the pad of Figure 7 (referred to as the Polarised pad);

Figure 11 is an isometric view of two separated ferrite cores showing an arrangement used for the purpose of simulating their performance in a power transfer system;

Figure 12 is a computer generated flux plot in a pad as shown in Figure 11 with 25 A current in windings provided about the ferrite core;

Figure 13 is a computer generated plot showing flux density in the ferrite core of the pad referred to in Figure 12 taken through an XY plane half way through the thickness (Z axis) of the ferrite core;

Figure 14 is a plan view of the arrangement of Figure 11 illustrating the position of a cut plane through the XZ axis at a point half way through the width (Y axis) of the ferrite cores of the assembly;

Figure 15 is a computer generated flux plot on the cut plane of Figure 14 for a 100mm separation between the pads;

Figure 16 is a computer generated flux plot on the cut plane of Figure 14 for a 200mm separation between the pads, and;

Figure 17 is a computer generated plot showing flux density in the cut plane of Figure 14 for a 200mm separation between the pads.

Description of One or More Embodiments

A new concept in IPT power transfer arrangements is now disclosed. The embodiments described in this document relate to flux transmission and reception apparatus. These are commonly (although not necessarily) provided in the form of a discrete unit which may conveniently be referred to as power transfer pads i.e. arrangements that may be portable and which typically have a greater extent in two dimensions relative to a third dimension so that they may be used in applications such as electric vehicle charging where one pad is provided on a ground surface (such as a garage floor) and another in the vehicle. However, the disclosed subject matter may also be provided in other arrangements including permanent structures such as a roadway for example, and does not need to take the form of a pad.

Referring to the arrangement of Figure 5, a pad is shown which combines three leakage flux control techniques to produce a much enhanced performance. In this regard it uses a novel "flux pipe", generally referenced 10, to connect two separated pole area that provide flux transmitter/receiver portions that comprise pole areas 11 and 12. The flux pipe provides a generally elongate region of high permeance allowing a high flux concentration from which ideally no flux escapes. The flux pipe 10 has a core 14 of a material such as ferrite to attract flux to stay in the core. A back-plate 15 of aluminium is provided adjacent to a rear face of the pad and acts to 'frighten' or repel flux from leaking from the core 14. Above the core 14 there may be a separate aluminium plate 16 adjacent to a front face of the pad to complete the same 'frightening' or shaping task. Magnetic flux is attracted to the ferrite, and it is repulsed by the aluminium. With electric circuits there is a large difference between the conductivity of conductors, typically 5.6×10^7 for copper; and air – in the order of 10^{-14} – but this situation does not pertain with magnetic fields where the difference in permeability between ferrite and air is only the order of 10,000 : 1. Thus in magnetic circuits leakage flux in air is always present and this has to be controlled to get the best outcome.

The ends of the core 14 comprise the transmitter/receiver portions 11 and 12. The top plate 16 does not cover the end portions 11 and 12, so the flux is directed upwardly from the ends to provide flux in the space beyond the front face of the pad as will be seen further below.

Plate 16 cannot be electrically connected to the backing plate 15 or the combination would constitute a short circuited turn. There is a winding on the flux core to electrically connect to the pick-up and the third flux control technique concerns this winding. It is well known that long toroidal windings have zero or very small leakage flux outside them. In the situation here a toroidal winding covering the full length of the flux pipe would have too much inductance but the winding can be partitioned into several windings 17 that are magnetically in series but electrically in parallel, as shown in Figure 6. In practice two windings in magnetic series-electrical parallel placed with one at or toward each end of the flux pipe is a good approximation to a continuous winding and in some circumstances may outperform a single winding.

The provision of a winding arrangement that covers substantially the full length of the core 14 means that little flux escapes from the core. For example, in the embodiment having two windings connected electrically in parallel (magnetically in series), the flux linkages in each winding must be the same so essentially no flux can escape from the core. Thus, plate 16, in this embodiment, is not essential.

The flux paths from a pick-up as in Figure 5 are shown diagrammatically in Figure 7 by flux lines 20. In Figure 7A a computer generated simulation of the magnetic field (indicated by flux lines 20) of the pad of Figure 6 is shown. As before they are approximately semi-elliptical but they are from a much larger base than the power pad ferrites of Figure 1 and therefore can operate over much larger separations. At the centre of the pick-up the flux paths are horizontal as required. A practical pick-up is shown in Figure 8, and measured self inductance and mutual inductance for this pick-up are illustrated in Figure 9. A performance comparison of the circular pad of figure 1 and the new pick up of figure 8 is shown in figure 10. The pad design of Figures 5 and 8 is polarised so that the ends 11 and 12 must be aligned, but that is relatively easy to implement.

As shown in Figure 8, some embodiments may include pole areas 11 and 12 that include finger portions 13. These allow the flux to be distributed more widely while using a minimal quantity of permeable material, thus lowering weight and cost.

A useful feature of the new pad design disclosed herein is that the winding number of the primary and secondary coils may in some embodiments be kept the same. This is quite different from the conventional IPT system setup, which normally has an elongated loop of one turn on the primary side and has a winding with multiple turns on the secondary side. This setup has two significant features, 1) the magnetic structure of both the primary and the secondary of the charger pads are the same, and 2) the induced voltage and uncompensated power at the secondary output (i.e. the pick-up pad) are independent of the operating frequency by varying the number of turns in relation to the frequency change.

The uncompensated power (S_u) and induced voltage (V_{oc}) of an IPT pick-up are commonly known and are expressed in equation 1 and 2, where I_1 is the primary track current, L_1 is the primary track inductance and N_1 and N_2 are the number of turns in the primary and secondary respectively. N_1 is equal to N_2 in this new pad design.

Under these conditions the rated uncompensated power for the pick-up S_u, the mutually coupled voltage V_{oc} and the terminal voltage on the primary V_1 are given by

$$\begin{aligned} S_u &= \frac{\omega \cdot M^2 \cdot I_1^2}{L_2} \\ &\propto \frac{f \cdot (N_1 N_2)^2 \cdot I_1^2}{N_2^{\,2}} \\ &\propto f \cdot N^2 \cdot I_1^2 \end{aligned} \qquad (1)$$

$$\begin{aligned} V_{oc} &= j\omega \cdot N_1 N_2 \cdot I_1 \\ &\propto f \cdot N^2 \cdot I_1 \end{aligned} \qquad (2)$$

And

$$\begin{aligned} V_1 &= j\omega \cdot L_1 \cdot I_1 \\ &\propto f \cdot N^2 \cdot I_1 \end{aligned} \qquad (3)$$

Note that the short circuit current is proportional to M/L and is independent of the number of turns

$$I_{SC} = I\frac{M}{L_2} = I \cdot k \qquad (4)$$

where k is the magnetic coupling factor between the primary and the secondary. As mentioned earlier, the pick-up induced voltage and the uncompensated power are to be the same for a different operating frequency. This also means that the terminal voltage and the short circuit current are also equal. Equations 1 and 2 can be rewritten as shown in equations 5 and 6 respectively for the same uncompensated power and induced voltage but different operating frequency.

$$f_a N_a^{\,2} I_a^{\,2} = f_b N_b^{\,2} I_b^{\,2} \qquad (5)$$

$$f_a N_a^{\,2} I_a = f_b N_b^{\,2} I_b \qquad (6)$$

From equation 5:

$$N_b I_b = N_a I_a \sqrt{\frac{f_a}{f_b}} \tag{7}$$

Using equation 6 and 7:

$$\frac{N_b}{N_a} = \frac{f_a N_a I_a}{f_b N_b I_b} = \frac{f_a}{f_b}\sqrt{\frac{f_b}{f_a}} = \sqrt{\frac{f_a}{f_b}} \tag{8}$$

Using equation 3 and 6:

$$\frac{I_b}{I_a} = \sqrt{\frac{f_a N_a^{\,2}}{f_b N_b^{\,2}}} = \sqrt{\frac{f_a}{f_b} \cdot \frac{f_b}{f_a}} = 1 \tag{9}$$

Equation 5 to 9 indicate that the pick-up uncompensated power and V_{oc} will be the same for different frequency while the primary current is kept the same and the winding turns are varied according to equation 8. For example, a charger pad with 15 turns on both primary and secondary, designed to operate at 38.4 kHz, would need to have the number of turns increased to 21 at 20 kHz in order to keep the pick-up V_{oc} and uncompensated power the same. In other words, this feature enables charger pads with the same magnetic design to be used at a different frequency, and the pick-up output characteristic can be maintained the same simply by scaling the turns number accordingly. However, as shown in equation 10, the core flux is proportional to the number of turns and current, thus keeping the current constant and varying the number of turns will vary the core flux, and hence the flux density. By substituting equation 8 into equation 10, it can be shown that the flux in the core is varying proportional to $\sqrt{(f_a/f_b)}$, which is equivalent to equation 8. Thus, if the operating frequency is scaled down, the cross sectional area of the ferrite core may need to be increased to avoid ferrite saturation. An increase of cross sectional area is preferably done by increasing the thickness of the ferrite core so the magnetic reluctance path of the charger pad remains nearly identical.

$$\begin{aligned} \phi &= \frac{L \cdot I}{N} = \frac{N^2 \cdot I}{N \cdot R_m} \\ &\propto N \cdot I \end{aligned} \tag{10}$$

where R_m is the magnetic reluctance of the flux path.

The eddy current loss (P_e) and hysteresis loss (P_h) equations for the core are shown in equation 11 and 12 in units of W/m^3. If the ferrite core cross sectional area are kept the same, the ratio of the eddy current loss and hysteresis loss for two different operating frequencies are given by equations 13 and 14.

$$P_e \propto B^2 f^2 \propto \frac{\phi^2 f^2}{A^2} \propto \frac{N^2 I^2 f^2}{A^2} \qquad (11)$$

$$P_h \propto f \cdot B^n \propto f \cdot \left(\frac{\phi}{A}\right)^n \propto f \cdot \left(\frac{N \cdot I}{A}\right)^n \qquad (12)$$

where n is the Steinmetz coefficient for the material and is normally in the range of 1.6 – 2.

$$\begin{aligned}\frac{P_{e.b}}{P_{e.a}} &= \frac{N_b^{\,2} I^2 f_b^{\,2}}{N_a^{\,2} I^2 f_a^{\,2}} = \left(\frac{N_b f_b}{N_a f_a}\right)^2 = \left(\sqrt{\frac{f_a}{f_b}} \cdot \frac{f_b}{f_a}\right)^2 \\ &= \frac{f_b}{f_a}\end{aligned} \qquad (13)$$

$$\begin{aligned}\frac{P_{h.b}}{P_{h.a}} &= \frac{f_b (N_b I_b)^2}{f_a (N_a I_a)^2} = \frac{f_b N_b^{\,2}}{f_a N_a^{\,2}} \\ &= 1\end{aligned} \qquad (14)$$

The above expressions suggest that for the same cross sectional area and volume, the hysteresis loss of the core will remain constant regardless of the frequency but the eddy current loss in the core will decrease proportionally to the decrease of operating frequency. As the overall power loss in a ferrite core is dominated by its hysteresis loss, most of the attributes, apart from the core flux density, of the charger pad will remain approximately the same with the operating frequency scaling process.

However, as discussed earlier the trade off of operating at a lower frequency is the increase of flux density in the core by $\sqrt{(f_a/f_b)}$. Thus to accommodate the higher flux density the ferrite cross sectional area should be increased in order to keep the flux density the same. With this increased volume of ferrite and keeping flux density constant, the power loss density in the ferrite core is expected to be lower as shown below. Equation 11 and 12 express the eddy current loss and hysteresis loss in terms of watt per m^3, thus the total eddy current and hysteresis loss should take into account the ferrite volume (A*L) shown in equation 15 and 16 respectively.

$$\begin{aligned}\frac{P_{e.b}}{P_{e.a}} &= \frac{\frac{\phi_b^{\,2}}{A_b^{\,2}} \cdot f_b^{\,2} \cdot A_b \cdot L}{\frac{\phi_a^{\,2}}{A_a^{\,2}} \cdot f_a^{\,2} \cdot A_a \cdot L} = \frac{N_b^{\,2} I^2 f_b^{\,2}}{N_a^{\,2} I^2 f_a^{\,2}} \cdot \frac{A_a}{A_b} \\ &= \frac{f_b}{f_a} \cdot \frac{A_a}{A_b}\end{aligned} \qquad (15)$$

where L is the length of the charger pad ferrite core length and is kept constant.

$$\frac{P_{h.b}}{P_{h.a}} = \frac{f_b \cdot \left(\frac{\phi_b}{A_b}\right)^2}{f_a \cdot \left(\frac{\phi_a}{A_a}\right)^2} \cdot \frac{A_b \cdot L}{A_a \cdot L} = \frac{f_b \cdot \left(\frac{N_b I}{A_b}\right)^2}{f_a \cdot \left(\frac{N_a I}{A_a}\right)^2} \cdot \frac{A_b}{A_a} = \frac{f_b}{f_a} \cdot \left(\frac{N_b}{N_a}\right)^2 \cdot \frac{A_a}{A_b} \qquad (16)$$

$$= \frac{A_a}{A_b}$$

Referring to the example discussed earlier where a charger pad operating frequency was scaled from 38.4 kHz to 20 kHz, the ferrite area will need to be increased by a factor of 1.385 √(38.4 kHz/20 kHz) in order to keep the flux density the same. Thus the eddy current and hysteresis loss of the charger pad, operating at 20 kHz, will be reduced by 37.59% and 72.17% respectively, compared with operating at 38.4 kHz at the same core flux density.

A simulated example

Referring now to Figures 11 to 17 a simulation of coupled power pads according to the invention will be described to provide an example of a possible embodiment and its use. In this example a coupled system of power pads is simulated with the pick-up winding open circuited. Figure 11 shows the arrangement of the ferrite core which is essentially 93 x 28 x 16 mm blocks of ferrite ground to give very close fitting, and then glued together. The ferrite is surrounded by an aluminium wall with an 8 mm gap between the ferrite and the aluminium, and is 5 mm above an aluminium backing plate. A flux plot for the driven pad (i.e. the pad connected to a power supply) is shown in Figure 12 for the situation where there are two coils driven magnetically in series, electrically in parallel with a current of 23 A. In these circumstances the flux density midway through the ferrite is shown in Figures 12 and 13. As shown the "flux pipe" is very effective in carrying the flux from one end of the pad to the other. Also, it can be seen from Figures 15 and 16 that there is essentially no leakage flux beyond the region between the pads.

For coupled pads a cut-plane is shown in Figure 14 and the other Figures use measurements along this cut-plane to illustrate the performance of the system. The flux lines at 100 mm spacing between pads are given in Figure 15 and for 200 mm spacing in Figure 16. The flux density in the ferrites is shown in Figure 18. The constant flux density in the ferrites of Figure 18 shows that the flux pipe efficiently carries flux from one end of the pad to the other and thereby provides good magnetic coupling between the two pads. The maximum flux density in the driven pad (in the ferrite) is approximately 0.2 T which is safely below saturation for this ferrite. The flux density in the pick-up pad is lower but will increase substantially to about the same as the transmitter pad when the pick-up is resonated.

Therefore, the invention provides a low profile device, referred to herein as a pad, which can be used as a magnetic flux generator that can be used to generate useful flux a significant distance from the device. The device can also be used as a receiver of flux to thereby produce electric energy from the received field. The ability of the pad to generate or receive flux over a significant distance is particularly useful for charging or energising an electric vehicle. Although certain examples and embodiments have been disclosed herein it will be understood that various modifications and additions that are within the scope and spirit of the invention will occur to those skilled in the art to which the invention relates. All such modifications and additions are intended to be included in the scope of the invention as if described specifically herein.

The word "comprise" and variations such as "comprising", unless the context clearly requires the contrary, is intended to be interpreted in an inclusive sense (i.e. as meaning "including, but not limited to").

CLAIMS:

1. A magnetic flux pad having a front face and a back face for generating or receiving magnetic flux, the pad comprising:
   two pole areas for sending or receiving flux;
   a magnetically permeable core magnetically connecting the pole areas;
   a coil wound about the core; and
   whereby the flux enters the pad at one of the pole areas and exits the pad at the other pole area.

2. A magnetic flux pad as claimed in claim 1 including a flux shaping means such that flux is directed into a space beyond the front face of the pad.

3. A magnetic flux pad as claimed in claim 2 wherein the flux shaping means is located adjacent to the back face of the pad.

4. A magnetic flux pad as claimed in claim 2 or claim 3 wherein the flux shaping means comprises a member constructed from a flux repelling material.

5. A magnetic flux pad as claimed in claim 1 including a flux shaping means such that flux is substantially prevented from escaping from the core.

6. A magnetic flux pad as claimed in claim 5 wherein the flux shaping means comprises a flux repelling member located adjacent to the front face of the pad.

7 A magnetic flux pad as claimed in claim 6 wherein the flux shaping means further comprises a flux repelling member located adjacent to the rear face of the pad.

8. A magnetic flux pad as claimed in any one of the preceding claims wherein the coil comprises a plurality of coils.

9. A magnetic flux pad as claimed in claim 8 wherein the coils are connected electrically in parallel.

10. A magnetic flux pad as claimed in claim 8 wherein the coils are connected magnetically in series.

11. A magnetic flux pad as claimed in any one of the preceding claims wherein each pole area includes a plurality of fingers.

12. A magnetic flux pad having a front face and a back face for generating a magnetic flux in a space beyond the front face of the pad, the pad comprising:
    two pole areas for sending or receiving flux;
    a magnetically permeable core magnetically connecting the pole areas;
    a coil wound about the core;
    a flux repelling means provided adjacent to a rear face; and
    whereby the flux enters the pad at one of the pole areas and exits the pad at the other pole area.

13. An IPT power transfer pad including: a magnetic flux carrying member having a high magnetic permeability and two ends, each end being substantially adjacent a peripheral edge of the pad; and one or more windings provided about at least a part of the flux carrying member; and wherein said pad is configured such that magnetic flux exits or enters the flux carrying member substantially only at or adjacent to the ends.

14. An IPT system including a first magnetic flux pad or IPT power transfer pad for connection to a power supply and a second magnetic flux pad or IPT power transfer pad for connection to a load, the first and second magnetic flux pads or IPT power pads constructed according to any one of claims 1-13 and having one or more windings with the same number of turns, and wherein the number of turns is selected dependent on a required operating frequency.

15. An IPT system including a magnetic flux pad or IPT power transfer pad according to any one of claims 1-13.

16. An IPT system as claimed in claim 14 or claim 15 wherein the system supplies power to an electric vehicle.

17. A magnetic flux pad substantially as herein described with reference to the accompanying drawings.

18. An IPT system substantially as herein described with reference to the accompanying drawings.

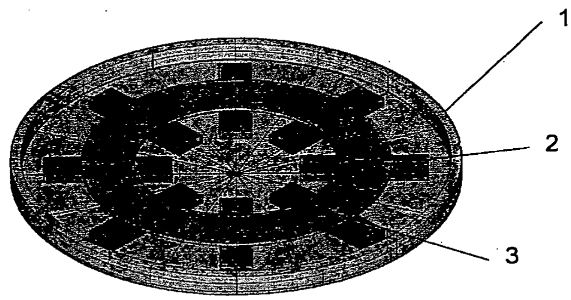

FIGURE 1

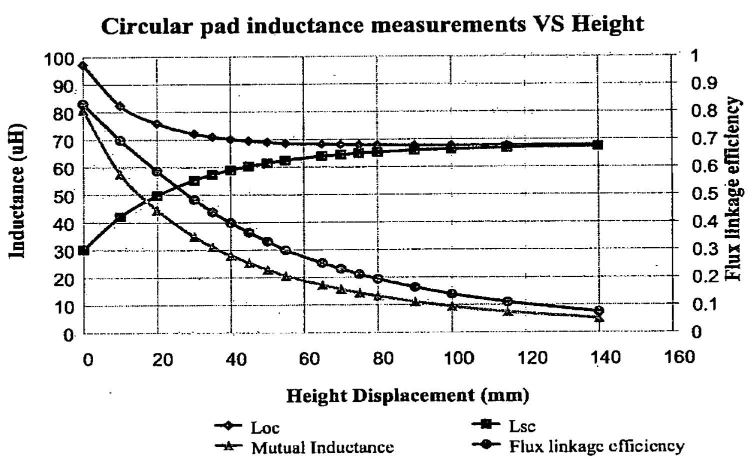

FIGURE 2

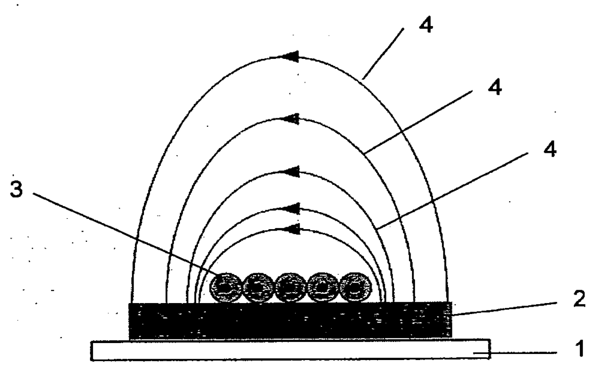

FIGURE 3

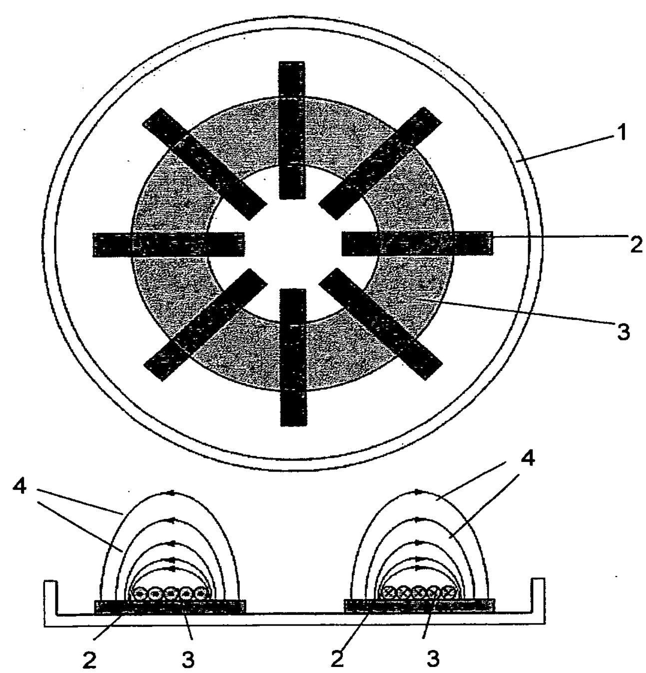

FIGURE 4

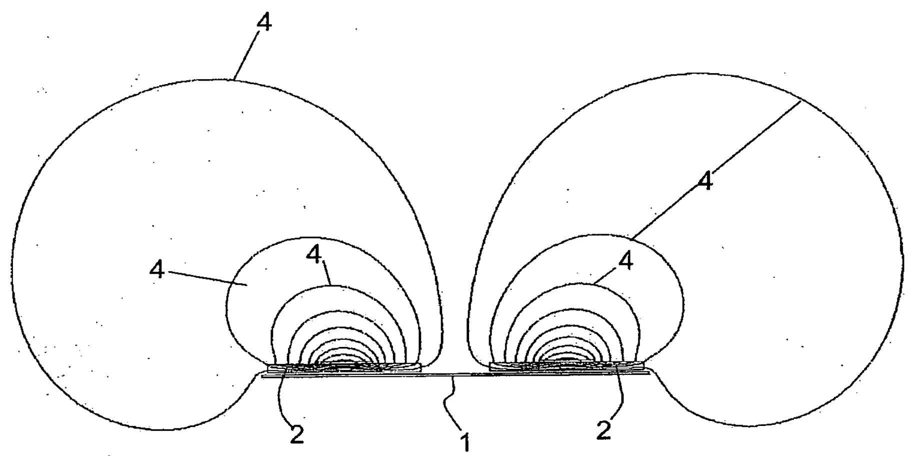

FIGURE 4A

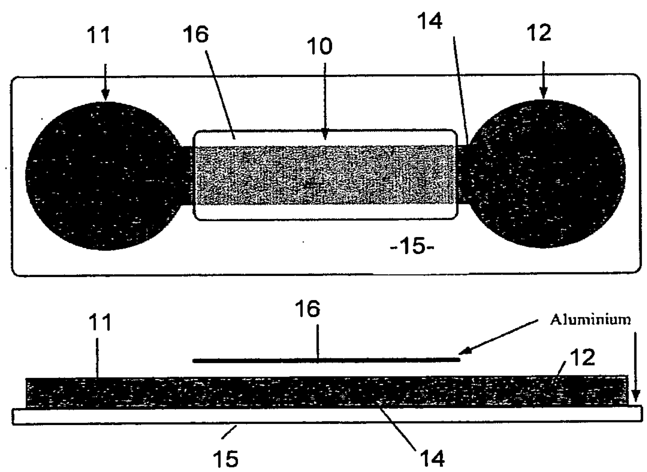

FIGURE 5

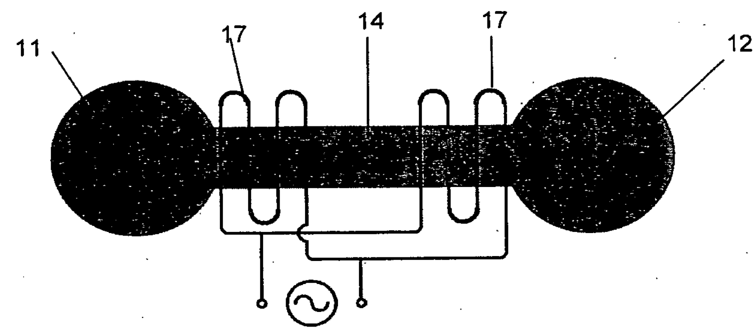

FIGURE 6

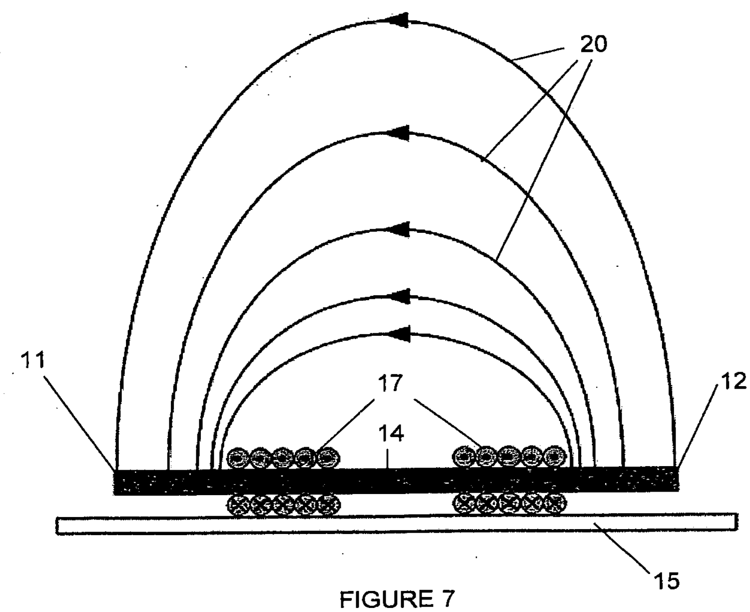

FIGURE 7

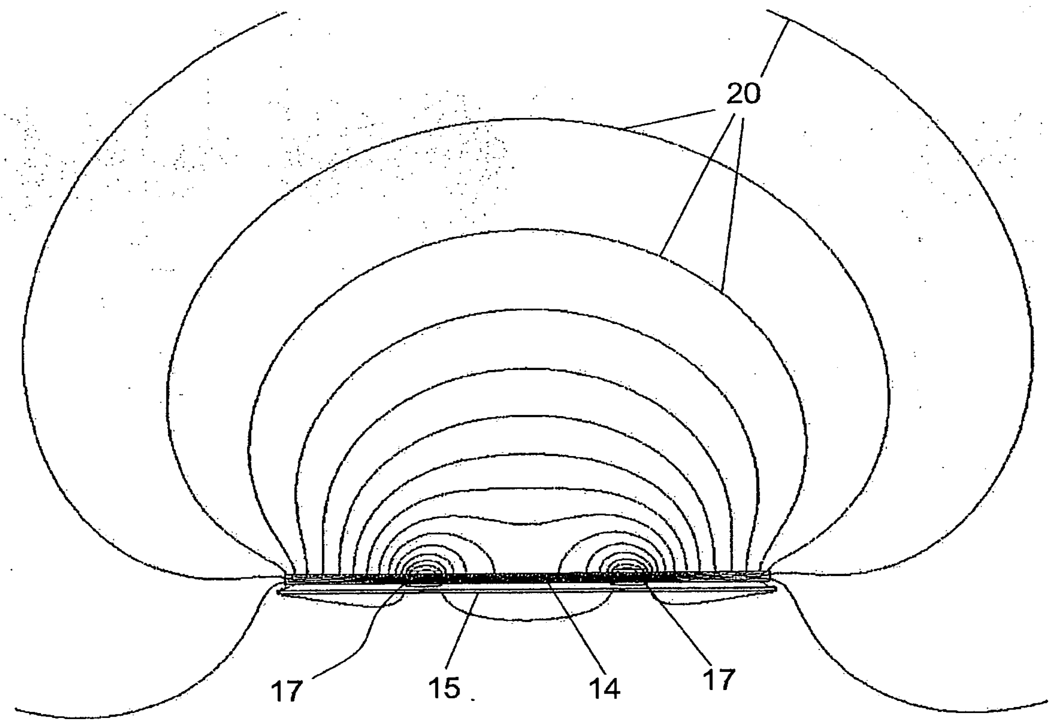

FIGURE 7A

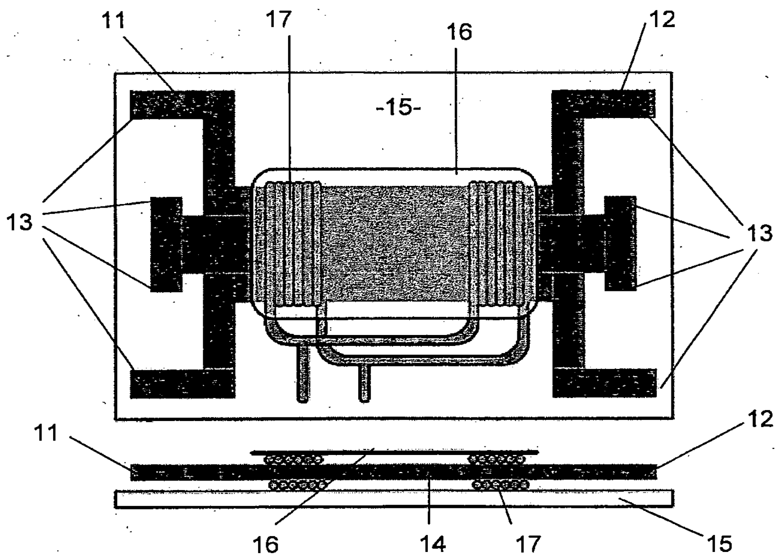

FIGURE 8

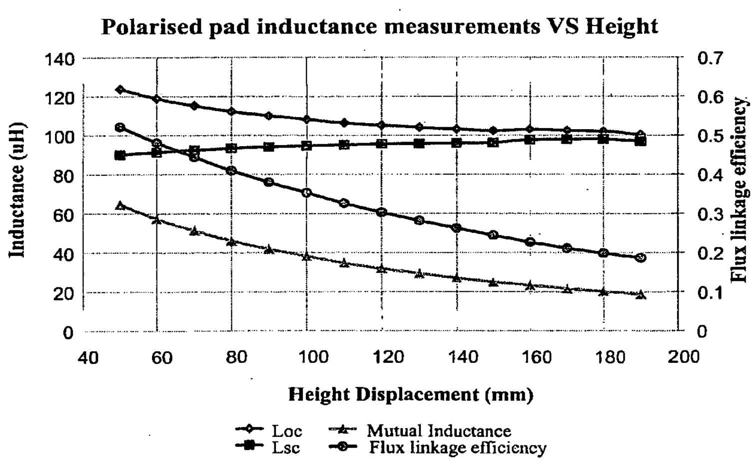

FIGURE 9

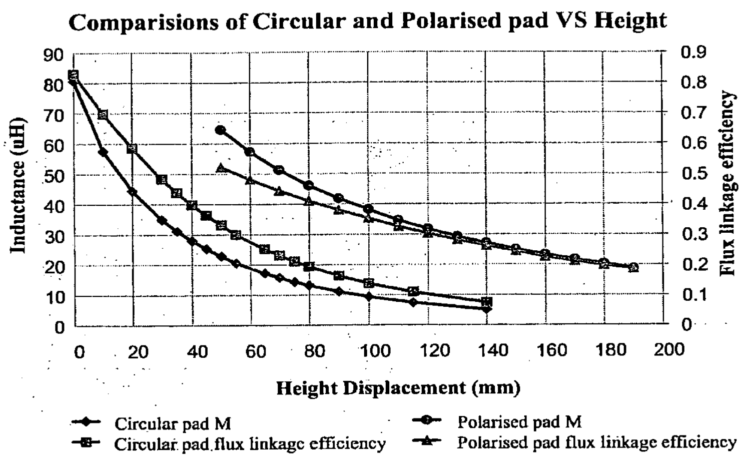
Comparisions of Circular and Polarised pad VS Height
90
80
70
60
50
40
30
20
10
0
0.9
0.8
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0
Inductance (uH)
Flux linkage efficiency
0
20
40
60
80
100
120
140
160
180
200
Height Displacement (mm)
Circular pad M
Circular pad flux linkage efficiency
Polarised pad M
Polarised pad flux linkage efficiency

FIGURE 10

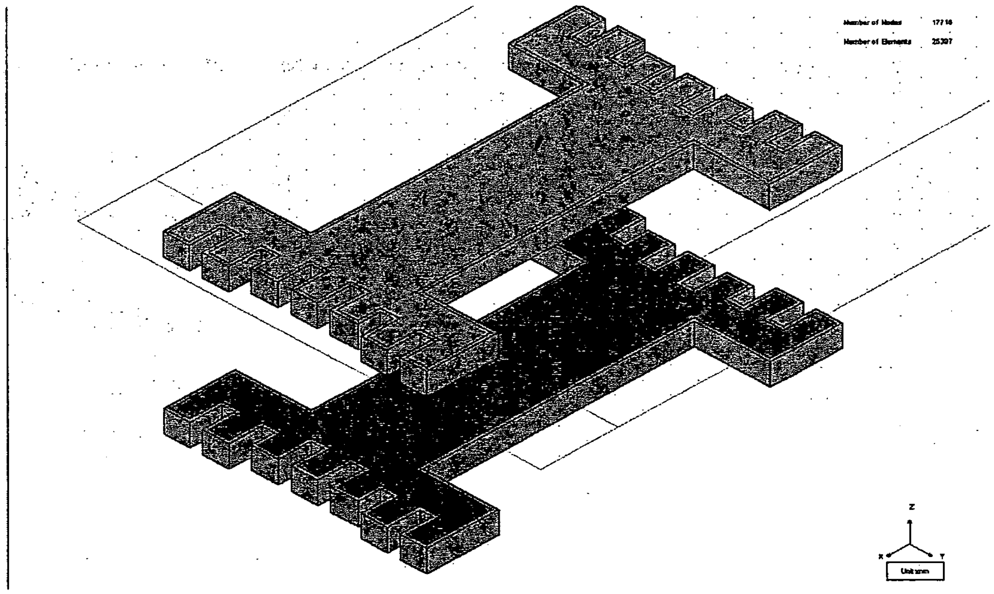

FIGURE 11

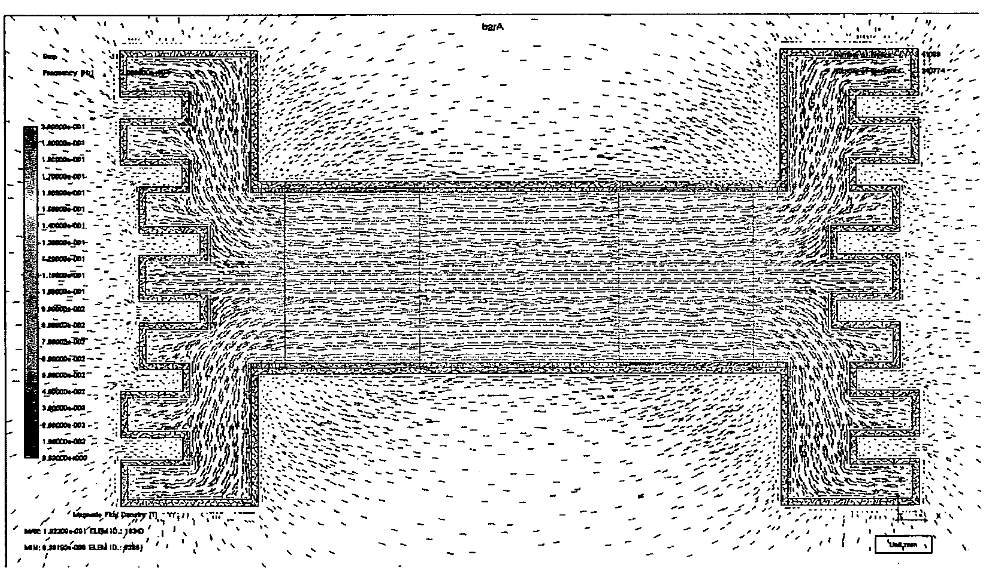

FIGURE 12

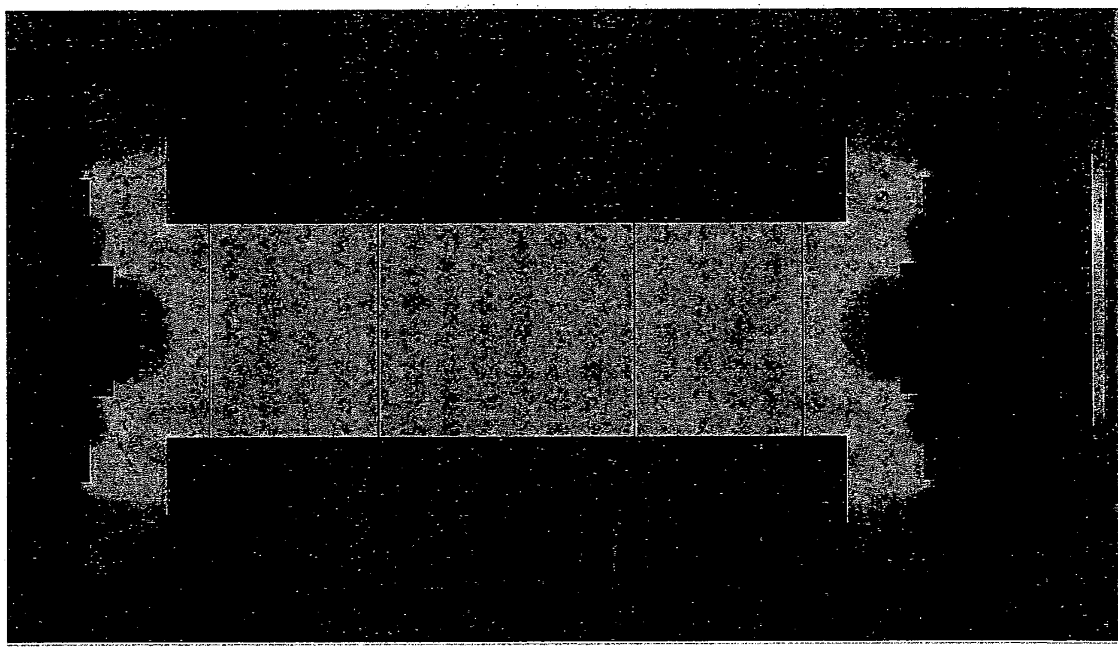

FIGURE 13

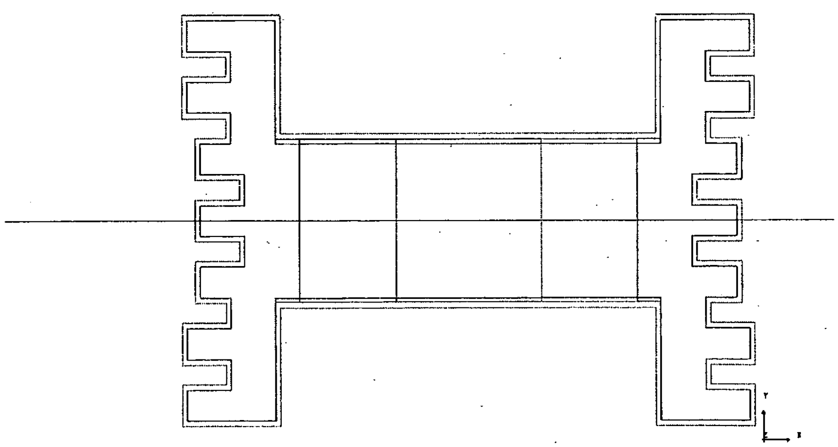

FIGURE 14

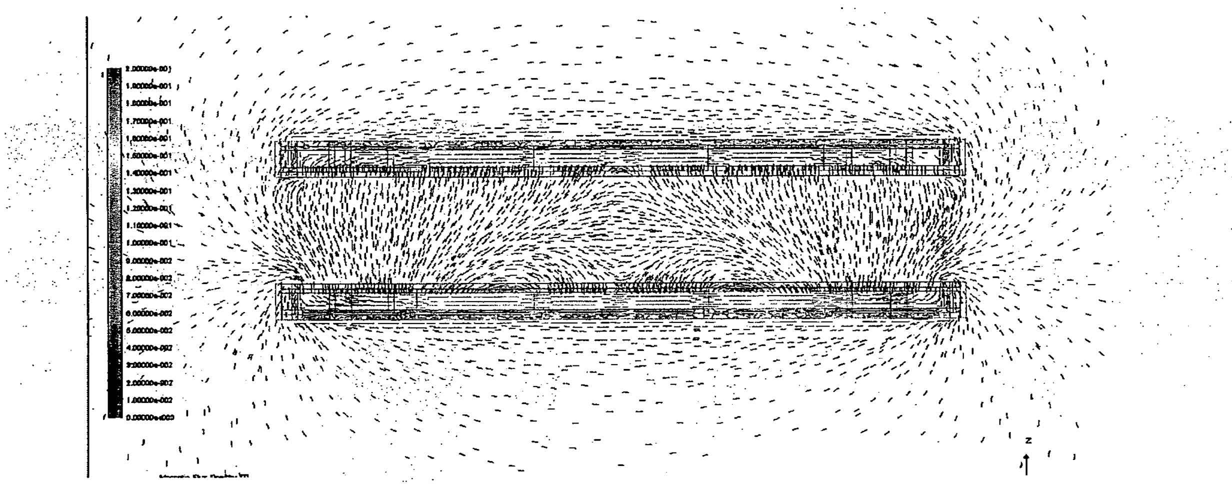

FIGURE 15

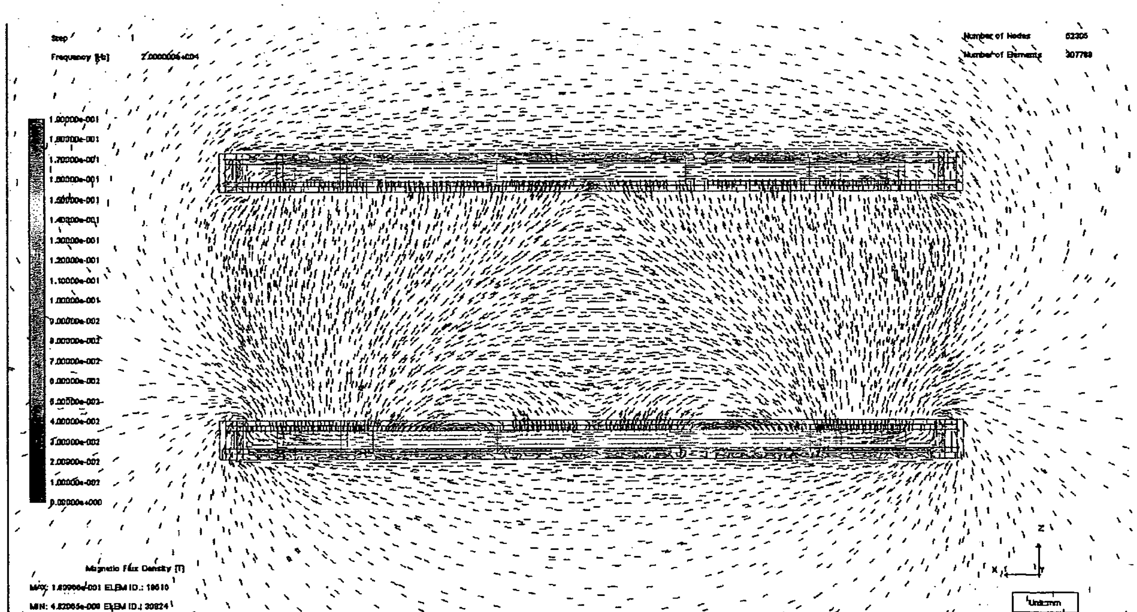

FIGURE 16

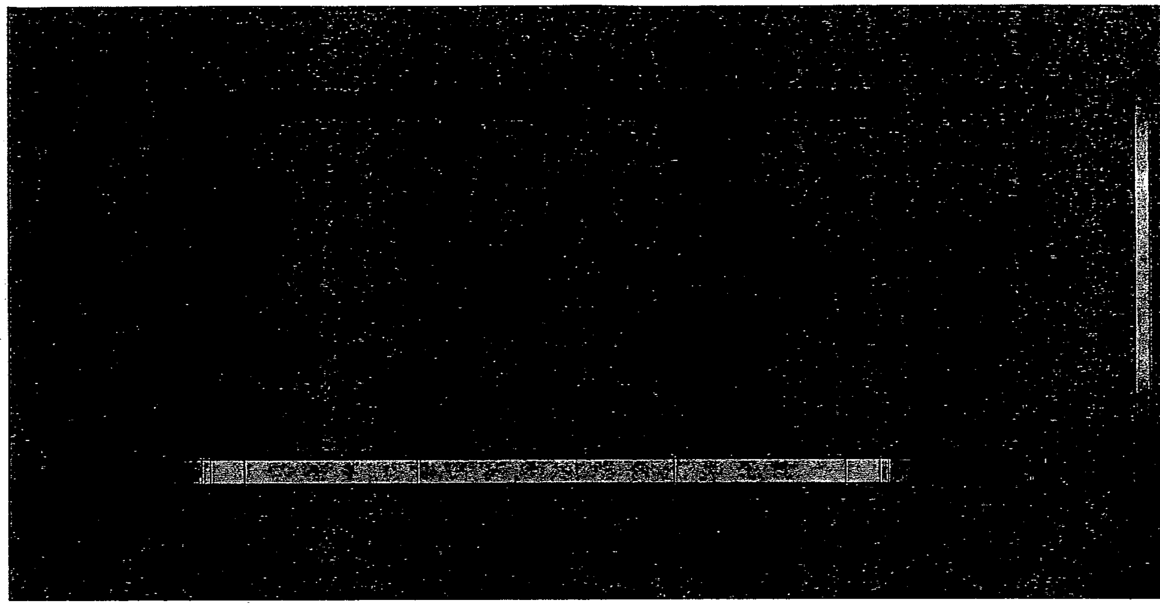

FIGURE 17

INTERNATIONAL SEARCH REPORT

International application No.
PCT/NZ2010/000017

A. CLASSIFICATION OF SUBJECT MATTER

Int. Cl.

H02J 3/00 (2006.01)

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)
WPI: magnetic, flux, pad, inductive, power, transfer and similar terms
Google Patents& Esp@ce: inductive, power, transfer and similar terms

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X Y	US 7,042,196 B2 (KA-LAI et al.) 9 May 2006 Figs. 1b, 4a, 4c; column 18, lines 21-22, column 18, lines 30-40, column 19, lines 13-17, column 19, lines 18-24,	1-3, 5, 8-10, 13, 14, 15 1-7, 12, 13, 15
Y	WO 2005/024865 A2 (SPLASHPOWER LIMITED) 17 March 2005 Fig. 7; page 4, lines 25-30, page 10, lines 6-16	1-7, 12, 13, 15
	WO 2008/140333 A2 (AUKLAND UNISERVICES LIMITED) 20 November 2008 whole document	
	WO 2007/126321 A1 (AUKLAND UNISERVICES LIMITED) 8 November 2007 whole document	

☐ Further documents are listed in the continuation of Box C ☒ See patent family annex

* Special categories of cited documents:

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier application or patent but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
31 May 2010	24 JUN 2010

Name and mailing address of the ISA/AU	Authorized officer
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Form PCT/ISA/210 (second sheet) (July 2009)

INTERNATIONAL SEARCH REPORT	International application No. PCT/NZ2010/000017

Box No. II Observations where certain claims were found unsearchable (Continuation of item 2 of first sheet)

This international search report has not been established in respect of certain claims under Article 17(2)(a) for the following reasons:

1. [X] Claims Nos.: **17;18**
   because they relate to subject matter not required to be searched by this Authority, namely:
   The claims do not comply with Rule 6.2(a) because they rely on references to the description and/or drawings.

2. [ ] Claims Nos.:
   because they relate to parts of the international application that do not comply with the prescribed requirements to such an extent that no meaningful international search can be carried out, specifically:

3. [ ] Claims Nos.:
   because they are dependent claims and are not drafted in accordance with the second and third sentences of Rule 6.4(a)

Box No. III Observations where unity of invention is lacking (Continuation of item 3 of first sheet)

This International Searching Authority found multiple inventions in this international application, as follows:

1. [ ] As all required additional search fees were timely paid by the applicant, this international search report covers all searchable claims.
2. [ ] As all searchable claims could be searched without effort justifying additional fees, this Authority did not invite payment of additional fees.
3. [ ] As only some of the required additional search fees were timely paid by the applicant, this international search report covers only those claims for which fees were paid, specifically claims Nos.:
4. [ ] No required additional search fees were timely paid by the applicant. Consequently, this international search report is restricted to the invention first mentioned in the claims; it is covered by claims Nos.:

Remark on Protest

[ ] The additional search fees were accompanied by the applicant's protest and, where applicable, the payment of a protest fee.

[ ] The additional search fees were accompanied by the applicant's protest but the applicable protest fee was not paid within the time limit specified in the invitation.

[ ] No protest accompanied the payment of additional search fees.

INTERNATIONAL SEARCH REPORT

Information on patent family members

International application No.
PCT/NZ2010/000017

This Annex lists the known "A" publication level patent family members relating to the patent documents cited in the above-mentioned international search report. The Australian Patent Office is in no way liable for these particulars which are merely given for the purpose of information.

Patent Document Cited in Search Report		Patent Family Member					
US	7042196	AU	2003233895	AU	2003240999	AU	2003282214
		AU	2008255158	CN	1653669	CN	101699708
		CN	101699709	CN	101699710	CN	101699711
		EP	1506554	EP	1506605	GB	2398176
		GB	2388715	GB	2388716	GB	2399225
		GB	2399226	GB	2399227	GB	2399228
		GB	2399229	GB	2399230	JP	2009010394
		US	2003210106	US	6906495	US	2005140482
		US	2005135122	US	7239110	US	2006076922
		US	7248017	US	2005116683	US	7525283
		US	2006061323	US	7622891	US	2009189565
		US	7714537	US	2009096414	WO	03096361
		WO	03096512	WO	2004038888	ZA	200408863
WO	2005/024865	EP	1665299	US	2007064406		
WO	2008/140333	AU	2008251143	CA	2687060	CN	101689761
		EP	2156532	KR	20100017582	NZ	555128
		US	2010109604				
WO	2007/126321	CN	101461114	EP	2013957	NZ	546955
		US	2009303749				

Due to data integration issues this family listing may not include 10 digit Australian applications filed since May 2001.

END OF ANNEX

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
12 August 2010 (12.08.2010)

PCT

(10) International Publication Number
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(51) International Patent Classification:
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(21) International Application Number:
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(22) International Filing Date:
5 February 2010 (05.02.2010)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
574677 5 February 2009 (05.02.2009) NZ
576137 8 April 2009 (08.04.2009) NZ

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(72) Inventors; and

(75) Inventors/Applicants *(for US only)*: **BOYS, John, Talbot** [NZ/NZ]; 41A Dominion Street, Takapuna, Auckland 1309 (NZ). **COVIC, Grant Anthony** [NZ/NZ]; 28 Haverstock Road, Sandringham, Auckland 1004 (NZ). **HUANG, Chang-Yu** [NZ/NZ]; 393 Lake Road, Takapuna, Auckland 0622 (NZ). **BUDHIA, Mickel Bipin** [NZ/NZ]; 58 Golf Road, Green Bay, Auckland 0600 (NZ).

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(81) Designated States *(unless otherwise indicated, for every kind of national protection available)*: AE, AG, AL, AM, AO, AT, AU, AZ, BA, BB, BG, BH, BR, BW, BY, BZ, CA, CH, CL, CN, CO, CR, CU, CZ, DE, DK, DM, DO, DZ, EC, EE, EG, ES, FI, GB, GD, GE, GH, GM, GT, HN, HR, HU, ID, IL, IN, IS, JP, KE, KG, KM, KN, KP, KR, KZ, LA, LC, LK, LR, LS, LT, LU, LY, MA, MD, ME, MG, MK, MN, MW, MX, MY, MZ, NA, NG, NI, NO, NZ, OM, PE, PG, PH, PL, PT, RO, RS, RU, SC, SD, SE, SG, SK, SL, SM, ST, SV, SY, TH, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VC, VN, ZA, ZM, ZW.

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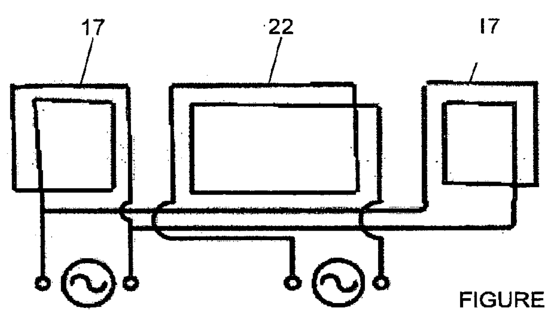


FIGURE 6

(57) Abstract: A magnetic flux pad for receiving or generating magnetic flux. The pad includes two pole areas (11, 12) associated with a magnetically permeable core (14). Coils (17) define the pole areas. The pad allows useable flux to be generated at a significant height above a surface of the pad.

INDUCTIVE POWER TRANSFER APPARATUS

Field of the Invention

This invention relates to apparatus for generating or receiving magnetic flux. The invention has particular, but not sole, application to a low profile, substantially flat device, such as a pad, for power transfer using an Inductive Power Transfer (IPT) system.

Background

IPT systems, and use of a pad including one or more windings that may comprise the primary or secondary windings for inductive power transfer, are introduced in our published international patent application WO 2008/140333, the contents of which are incorporated herein by reference. One particular application of IPT power transfer pads is electric vehicle charging. IPT power transfer pads are used both in the vehicle as a power "pick-up" device (i.e. the secondary side winding of the IPT system), and at a stationary location such as a garage floor as the "charging pad" (i.e. the primary side winding) from which power is sourced.

In the development of pick-ups for inductively charging electric vehicles a problem of some concern is the clearance available under the vehicle. With conventional pick-up circuits power in sufficient quantities can be provided at distances up to perhaps 100 mm at which time the coupling factor becomes so small that it becomes impractical.

It is generally conceded that the power required to charge a typical electric vehicle overnight is about 2.0 kW, so that in an overnight charging mode some 24 kWH can be transferred. With modern electric vehicles this is enough energy to travel more than 100 km and is ideal for small vehicles used for tasks such as dropping children at schools, running errands, short commutes and the like.

Inductively coupled chargers commonly use two power pads that are circular in shape and may have dimensions of 400 mm diameter by 25 mm thick as shown in Figure 1. However, to use an inductive charger such as this the vehicle must be positioned relatively accurately over the charging pad - typically within 50 mm of perfect alignment - and the separation between the power pad on the vehicle and the power pad on the ground must be closely controlled. In principle inductive power transfer may be accomplished for vertical spacings between 0 mm and 100 mm but if the system is set up for 100 mm it will have quite a large reduction in power at 120 mm and will be inoperable below 50 mm. This state of affairs occurs because both the self inductance and the mutual inductance of the power pads vary widely as the distance between the pads changes. The self inductance and the mutual inductance as a function of the separation for two identical circular pads that have the construction of Figure 1,are shown in Figure 2. Thus at 100 mm the power pad receiver or pick-up may have a pick-up voltage of 100 V and a short circuit current of 5.0 A for a power rating of 500 W. If the IPT system electronics operates with a Q factor of 4, then 2 kW can be transferred to the battery though there are still difficulties to overcome in producing the power needed at the appropriate battery voltage.

The induced voltage in the pick-up pad (i.e. the vehicle mounted power pad) is very separation sensitive – corresponding to the variation in mutual inductance shown in Figure 2 - so that at 120 mm it is reduced by approximately 40% while at 50 mm it is increased by a factor of 2. A reduction in power means that the vehicle does not get fully charged in the usual time, but the more challenging situation is that at smaller separations the power transferred may be so high that the components of the circuit are overloaded. Also, as the separation is reduced the self inductance of the pick-up coil also changes so that the circuit operates off-frequency putting extra stress on the power supply. As the separation gets smaller still this stress on the power supply caused by the non-tuned pick-up on the primary side cannot be sustained and the system must be shut down. In practice it is feasible to operate with a separation between 40 and 100 mm but a larger range is too difficult.

A range of separation from 40 to 100 mm is quite small. If the vehicle has a relatively high ground clearance then either the power pad on the vehicle has to be lowered or the power pad on the ground has to be raised. Automatic systems for doing this compromise the reliability of the charging system. Alternatively the pad on the ground can be on a fixed but a raised platform but such a pad is a tripping hazard when a car is not being charged and this situation is generally to be avoided in a garage or other location involving vehicles and pedestrians.

The known power pad construction of Figure 1 comprises an aluminium case 1 containing typically eight ferrite bars 2 and a coil 3. Current in the coil causes magnetic flux in the ferrite bars and this flux has flux lines that start on the ferrite bars and propagate to the other end of the bar in a path containing the coil that may be thought of as a semi-elliptical shape. The flux lines 4 for a single bar are shown in Figure 3. The flux lines leave the ferrite in an upward direction and propagate to the other end of the bar, entering it at right angles. No flux goes out the back of the pad as the solid aluminium backing plate 1 prevents it. In the actual pad the eight bars give a flux pattern shown approximately in cross section in Figure 4. A simulation of the actual flux pattern is shown in Figure 4A.

From Figure 4A it can be seen that at the highest point the flux lines are essentially horizontal. Therefore, to get the maximum separation possible between the primary pad and the secondary pad it would be advantageous to detect this horizontal flux. However, the horizontal flux is still relatively close to the pad (extending from the pad approximately one quarter of the diameter of the pad) and there is no horizontal flux at all at the very centre of the power pad. Thus at the very point where maximum flux density would be ideal – the centre – the actual usable horizontal flux component is zero.

Summary

It is an object of the invention to provide an improved apparatus or method for inductive power transfer, or an improved IPT power transfer pad, or to at least provide a useful alternative.

Accordingly in one aspect the invention provides a magnetic flux pad having a front face and a back face for generating or receiving magnetic flux in or from a space beyond the front face, the pad comprising:

two pole areas for sending or receiving flux,

a magnetically permeable core,

two coils magnetically associated with the core,

whereby the flux enters the pad at one of the pole areas and exits the pad at the other pole area.

The coils are flat coils in one embodiment, and each coil may define one of the pole areas.

The coils may be spiral wound coils.

In one embodiment the coils are located on a side of the core nearest to the front face of the pad, and the coils and the core together form a flux path in the pad.

The core may comprise lengths of permeable material, such as ferrite.

In one embodiment the turns of the coils are spread between the pole areas. In another embodiment the turns of the coils are concentrated at areas outside the region between the pole areas, such as the areas at the ends of the pad. The coils may each be asymmetric, the combination of windings of a coil being wider between the pole areas than at the periphery of the pad.

In one embodiment the coils are located immediately adjacent to each other in the region between the pole areas. The coils may touch each other at the region between the pole areas.

The coils may be shaped to provide the poles areas and a flux pipe between the pole areas. The coils may also be in substantially the same plane.

In a further aspect the invention provides a magnetic flux pad having a front face and back face for generating a magnetic flux in a space beyond the front face of the pad, the pad comprising:

two pole areas for sending or receiving flux,

a magnetically permeable core,

two coils associated with the core and being provided on a side of the core adjacent to the front face of the pad,

whereby the pad produces an arch shaped flux in the space such that essentially each line flux starts on one of the pole areas, arches to the second pole area and joins on itself through the core with essentially no flux present at the back face of the pad.

In a still further aspect the invention provides a magnetic flux pad having a front face and back face for receiving a magnetic flux from a space beyond the front face of the pad, the pad comprising:

two pole areas for sending or receiving flux,

a magnetically permeable core,

two coils associated with the core and being provided on a side of the core adjacent to the front face of the pad and adapted to receive a horizontal flux component,

and a further coil magnetically associated with the core and adapted to receive a vertical flux component.

In yet a further aspect the invention provides an IPT system including a magnetic flux pad according to any one of the preceding statements. The system may also include a transmitter pad and a receiver pad according to the preceding statements.

The IPT system may supply power to an electric vehicle.

In a further aspect the invention provides an IPT power transfer magnetic flux transmitter or receiver including:
a magnetic flux carrying member having a relatively high magnetic permeability and having two ends, two windings electromagnetically associated with the flux carrying member, and the flux carrying member having two flux transmission or reception regions, one region being provided adjacent to each end, whereby magnetic flux exists or enters the flux carrying member substantially only at or adjacent to the transmission or reception regions.

Further aspects of the invention will become apparent from the following description.

Drawing Description

One or more embodiments are described below by way of example with reference to the accompanying drawings, in which:
Figure 1 is a perspective view of part of a known form of IPT power transfer pad;
Figure 2 is a graph of inductance measurements and flux linkage efficiency with respect to height displacement for a pad such as that of Figure 1;
Figure 3 is a diagrammatic elevation in cross section of a part of the pad of Figure 1 showing flux lines;
Figure 4 is a plan view and elevation of a cross section of the pad of Figure 1 showing flux lines;
Figure 4A is an elevation in cross section of a computer generated simulation of the magnetic field (indicated by flux lines) of the pad of Figure 1;
Figure 5A is a plan view of an embodiment of inductive power transfer apparatus which may be provided in the form of a pad;
Figure 5B is a side elevation of the apparatus of Figure 19A;
Figure 5C is a view of Figure 5B but also showing flux lines;
Figure 5D is a computer simulation of s cross section through a pad according to Figures 5A-5C illustrating the flux lines generated by such a pad in use;
Figure 6 is a diagrammatic illustration of an electrical wiring diagram for a further embodiment of inductive power transfer apparatus including a centre, or quadrature coil;
Figure 7A is an isometric view from below of a flux transmitter and flux receiver (oriented above the flux transmitter).
Figure 7B is an isometric view from above of the arrangement of Figure 7A;

Figure 8A shows flux lines based on a simulation of the arrangement of Figures 7A and 7B when the transmitter and receiver are aligned with a 200mm separation between the transmitter and receiver;
Figure 8B shows flux lines based on a simulation of the arrangement of Figures 7A and 7B when the transmitter and receiver are misaligned in the X axis direction;
Figure 9 is a diagram of power against displacement in the X axis direction for the arrangement of Figures 7A and 7B;
Figure 10 is a diagram of power against displacement in the Y axis direction for the arrangement of Figures 7A and 7B;
Figure 11 is an illustrative diagram of a winding arrangement for coils of a pad of the preceding figures;
Figure 12 is a plan view of a former or support plate for construction of a pad according to the preceding figures;
Figure 13 is an isometric view of the former of figure 12, and;
Figure 14 is an isometric view of a backing plate adapted for attachment to the rear of the former shown in figures 12 and 13.

Description of Embodiments

A new concept in IPT power transfer arrangements is now disclosed. The embodiments described in this document relate to flux transmission and reception apparatus. These are commonly (although not necessarily) provided in the form of a discrete unit which may conveniently be referred to as power transfer pads i.e. arrangements that may be portable and which typically have a greater extent in two dimensions relative to a third dimension so that they may be used in applications such as electric vehicle charging where one pad is provided on a ground surface (such as a garage floor) and another in the vehicle. However, the disclosed subject matter may also be provided in other arrangements including permanent structures such as a roadway for example, and does not need to take the form of a pad. Like reference numerals refer to like features throughout the description.

Referring to the arrangement of Figure 5A, a pad is shown which uses a novel "flux pipe", generally referenced 10, to connect two separated flux transmitter/receiver regions comprising pole areas 11 and 12. The flux pipe provides a generally elongate region of high flux concentration from which ideally no flux escapes. The flux pipe 10 in this embodiment has a core 14 which includes a magnetically permeable material such as ferrite to attract flux to stay in the core. With electric circuits there is a large difference between the conductivity of conductors – typically 5.6 x 10^7 for copper; and air – in the order of 10^{-14} – but this situation does not pertain with magnetic fields where the difference in permeability between ferrite and air is only the order of 10,000 : 1 or less. Thus in magnetic circuits leakage flux in air is always present and this has to be controlled to get the best outcome.

Flat coils or windings 17 sit on top of the core 14 to provide the flux pipe. There is no *straight* path through the flux pipe that passes through the coils 17. Instead, the arrangement of the coils 17 means that flux entering the pad through one of the areas 11 or 12 propagates through the relevant coil 17 into the core 14 from where it propagates along the core, then exits the pad out through the other area 12 or 11, and completes its path through air back to the first area 11 or 12 to form a complete curved flux path. The flux path so formed is essentially completely above a front surface of the pad and extends into a space beyond the front surface. The arrangement of coils 17 also means that there is essentially no flux extending beyond a rear face of the pad. Thus, the orientation of the windings 17 ensures that the flux path is directed in a curve out into a space in front of the front surface of the pad, and the spread or distributed nature of the coils 17 across the upper surface of the core 14 ensures that the flux in the centre of the pad is primarily constrained within the core. The coils 17 also define the spaced apart pole areas so that the flux is guided into and out of the pad via the pole areas and forms an arch shaped loop in the space beyond the front surface of the pad to provide a significant horizontal flux component at a significant distance above the front surface of the pad.

In a preferred embodiment there are two coils 17 in close proximity to each other. The coils 17 are spiral wound. In the diagrammatic embodiment illustrated in Figures 5A to 5C the coils 17 take the form of Archimedean spirals, and are touching along the centre line 17A. The flux pipe 10, comprising core 14, extends to the ends of the coils 17. The coils 17 are substantially planar and are arranged in substantially the same plane on one side of the core 14. The actual length of the core 14 is not critical – in one embodiment it should include the centre line of the coils 17 and should extend past the hole in the centre of each coil to at least the position indicated by A. The core 14 may extend under the coil 17 to position B or even further. The holes in the coils 17 define the pole areas 11 and 12 which function as flux receiver/transmitter regions for the pad.

In one embodiment the core 14 is made of ferrite bars in strips or lengths (not shown in Figures 5A-C, but illustrated in Figures 7A and 7B). Air-gaps are acceptable between the strips, which simplifies manufacture. The ideal flux paths 20 are shown in Fig 5C and are only on one side of the core 14 which is an ideal situation. In principle there is ideally no flux extending out the rear face of the pad (i.e. on the side of the core 14 opposite to the side on which coils 17 are mounted) and therefore no aluminium screen or other flux repelling member is required. However, in practice a light screen may be used in some embodiments as errors and imperfections in the ferrite bars comprising the core 14 can cause small leakage fluxes that should be contained.

Figure 5D shows the result of a simulation of the pad construction in Figures 5A-5C when used to generate a magnetic field. As can be seen, the flux path follows a generally arch shape through the space beyond the front surface of the pad.

Inductive power transfer pads according to the arrangement described immediately above are very easy to use as the leakage flux from them is very small. They can be placed quite close to metallic objects without loss in performance, and they are largely unaffected by connecting wires etc.

Second Embodiment

In a further embodiment it may be noted that the arrangement of the coils in a receiver or pick-up pad mounted horizontally on a vehicle, for example, makes the pick-up pad sensitive to a first direction of the flux which is longitudinally directed (i.e. having a direction parallel to the core 14, and being in the X-axis direction with reference to the drawings) with respect to the flux generator (the horizontally oriented transmitter pad). To improve the magnetic coupling of the receiver with respect to misalignment, a “second” coil can be arranged that is sensitive to a second component of the flux that is preferably vertical with respect to the stationary transmitter.

Figure 6 shows an electrical schematic of a further embodiment of a receiver pad with a “horizontal” flux sensitive coil 22 now positioned in the centre and the outer two coils 17 connected out of phase to produce a further coil sensitive to the vertical component.

For the receiving pad of Figures 5A-5C a further flat coil 22 can also be placed above the flux pipe with one suitable arrangement shown in Figures 7A and 7B, coil 22 being sensitive to the vertical component of the field. As in the original pick-up structure, this additional coil exists only on one side of the core 14 and therefore ideally maintains all of the flux lines on the side of the receiver directed towards the transmitter.

As shown in Figures 7A and 7B, only the receiver is modified with a centre, or quadrature, coil 22. This second coil is particularly sensitive to misalignment in the X-direction (i.e. the horizontal longitudinal direction), but not in the Y-direction (being the horizontal transverse direction perpendicular to the core 14). This complements the original receiver which is sensitive to misalignment in the Y-direction, but which because of its structure is less sensitive to movement in the X-direction. The combined output of both receiver coils enhances the sensitivity of the receiver enabling the receiver to be positioned nominally in the ideal position and still couple the required power. Figures 7A and 7B also show an arrangement of spaced ferrite rods or bars 24 that comprise core 24.

As an example, the flux lines using the pad design as shown in Figures 7A and 7B without any form of compensation are shown in Figure 8B and 8A with and without some misalignment. Here the transmitter pad and receiver pad are identical except for the addition of the second "vertical flux" coil (i.e. coil 22 of Figures 7A and 7B) in the receiver pad. The transmitter and receiver pads both have length 588mm and width 406mm and are separated vertically by 200mm. The current in the coils of the transmitting pad is 23 amps at 20kHz. Notably the majority of the flux exists between the transmitter pad and receiver pad while a very small leakage flux is shown to exist outside this area. In Figure 8A these flux lines couple the first receiver coil, while in Figure 8B the majority of the flux lines couple the second receiving coil (i.e. coil 22 of Figures 7A and 7B) thereby enhancing the output power capability of the pick-up.

In Figures 9 and 10 the VA generated from the output of the receiver pad coils with and without misalignment is also shown. In Figure 9 the total and separate VA contribution of receiver coils from a magnetic simulation of the pads shown in Figures 7A and 7B is shown when the receiver pad is misaligned (relative to its ideal position centred above the transmitter pad) in the X direction. In Figure 9 curve 26 represents the VA contribution of coil 22, curve 28 represents the combined VA contribution of coils 17, and the remaining curve represents the total from coils 17 and 22. As noted the second coil 22 substantially enhances the output so that if a 2KW output were required at 0 X-offset the required electronic tuning must boost the VA output by around 3.2. At 140mm X-offset the required electronic boost (Q) without coil 22 is more than 17 times (which is practically difficult due to the sensitivity of the tuning required) whereas with coil 22 an effective boost of around 4.8 is required and that is easily achieved.

Coil 22 is not expected to be sensitive in the Y direction when the receiver is positioned with 0 offset in the X direction. This is verified in the magnetic simulations shown in Figure 10 where there is shown to be no contribution to the total power from the coil 22. This is however not required as the combined output of coils 17 is naturally sensitive in this direction. At 140mm offset in the Y direction, a 2KW output is possible with an electronic tuning (Q) of around 5.5.

Third Embodiment

Turning now to Figure 11, a winding arrangement for coils 17 is shown diagrammatically. In this embodiment the individual turns in the coils 17 are spread at that end of each winding nearer the centre of the pad relative to the ends of the pad. Thus the coils 17 are each asymmetric, the combination of windings of a coil being wider between the pole areas than at a periphery of the pad. This embodiment allows greater separation of the pole areas 11 and 12 (and thus greater flux extension beyond the front face of the pad). The spacing between the pole areas 11 and 12 may be made larger by using an oval or rectangular cross section litz wire wound on the narrow edge for the pole areas and wound on the flat edge for the central flux pipe region between the pole areas.

Alternatively if the coils are wound with a round wire the spacing between the pole areas 11 and 12 may be made larger using gaps between the windings of the flux pipe section between the pole areas. However, we have found that gaps in the individual windings over the flux pipe section are to be treated with care as they can leave holes that flux can leak through spoiling the efficiency of the flux pipe. We have found that it is preferable to keep the windings evenly spaced and if there are gaps they should be typically less than one half to one wire diameter to keep flux losses to a minimum. In practice we have found that the convenience of the simple round wire makes this the technology of choice.

In yet another embodiment, the shape of the windings 17 may assist in obtaining greater pole area separation. For example, the coils 17 may be wound in an approximately triangular shape with the apex of each triangle facing the centre of the pad.

Referring now to Figure 12, a former or support plate 30 for providing a pad including the winding layout of Figure 11 is shown in plan view. The former 30 may be constructed from any non permeable material, such as a plastic for example. The former 30 includes a first general region 32 for forming and supporting one of coils 17 (not shown) and a second region 38 for forming and supporting the other of the coils 17. Depressions 34 are provided for locating and supporting ferrite bars or other permeable members. Grooves 36 receive the individual wires that comprise the turns of coils 17 and ensure that the turns are correctly spaced. For clarity, an isometric representation of the former 30 is shown in Figure 13.

Figure 14 shows a backing plate 40 adapted for location on a rear surface of the former 30 i.e. for location on a side of the former that is opposite the side on which the coils 17 are located, and adjacent to a rear surface of the pad. Backing plate 40 may be constructed from a flux repelling material, for example aluminium. It is not necessary for prevention of flux exiting the rear face of a pad in use, since the design of the flat coils 17 and their location on the core 14 substantially directs the flux into a space in front of the front surface of the pad. However, plate 40 can provide additional structural support for the pad. Plate 40 can also act to prevent any changes in the magnetic properties of the pad (for example a change in inductance) should the pad be mounted in use adjacent to a magnetically permeable material for example.

The dimensions of the former 30 are approximately 790mm by 600mm by 25mm, and a pad constructed from such a former will have very similar dimensions.

Further Practical Considerations

In practice it is prudent to ensure that the voltage at the terminals of the pad does not reach unsafe levels. Therefore in some embodiments, capacitance may be added in series with the windings inside the pad to lower the inductance seen at the pad terminals and therefore control the voltage at these terminals to be within suitable limits (say 300-400V). Without this the terminal voltage could be several KV which is undesirable and potentially unsafe. Capacitance can be placed in series with the windings at nearly any convenient place with the apparatus. Thus in some embodiments one or more capacitors can be placed in series with the windings at the terminal points inside the pad housing, and in other embodiments capacitors can be distributed along the windings by breaking the winding into suitable sections with series capacitances in case the internal voltages on a single coil are ever too high.

Therefore, the invention provides a low profile device, referred to herein as a pad, which can be used as a magnetic flux generator that can be used to generate useful flux a significant distance from the device. The device can also be used as a receiver of flux to thereby produce electric energy from the received field. The ability of the pad to generate or receive flux over a significant distance is particularly useful for charging or energising an electric vehicle.

The entire disclosures of all applications, patents and publications cited above and below, if any, are herein incorporated by reference.

Reference to any prior art in this specification is not, and should not be taken as, an acknowledgement or any form of suggestion that that prior art forms part of the common general knowledge in the field of endeavour in any country in the world.

Wherein the foregoing description reference has been made to integers or components having known equivalents thereof, those integers are herein incorporated as if individually set forth.

It should be noted that various changes and modifications to the presently preferred embodiments described herein will be apparent to those skilled in the art. Such changes and modifications may be made without departing from the spirit and scope of the invention and without diminishing its attendant advantages. It is therefore intended that such changes and modifications be included within the present invention.

Claims

1. A magnetic flux pad having a front face and a back face for generating or receiving magnetic flux in or from a space beyond the front face, the pad comprising:
   two pole areas for sending or receiving flux,
   a magnetically permeable core,
   two coils magnetically associated with the core,
   whereby the flux enters the pad at one of the pole areas and exits the pad at the other pole area.

2. A magnetic flux pad as claimed in claim 1 wherein the coils are flat coils.

3. A magnetic flux pad as claimed in claim 1 or claim 2 wherein each coil defines one of the pole areas.

4. A magnetic flux pad as claimed in any one of the preceding claims wherein the coils are spiral wound coils.

5. A magnetic flux pad as claimed in any one of the preceding claims wherein the coils are located on a side of the core nearest to the front face of the pad.

6. A magnetic flux pad as claimed in any one of the preceding claims wherein the coils and the core together form a flux path in the pad,

7. A magnetic flux pad as claimed in any one of the preceding claims wherein the core comprises lengths of permeable material.

8. A magnetic flux pad as claimed in claim 7 wherein the permeable material comprises ferrite.

9. A magnetic flux pad as claimed in anyone of the preceding claims wherein the coils are each asymmetric, the combination of windings of a coil being wider between the pole areas than at a periphery of the pad.

10. A magnetic flux pad as claimed in any one of the preceding claims wherein the coils are located immediately adjacent to each other in the region between the pole areas.

11. A magnetic flux pad as claimed in any one of the preceding claims wherein the coils are shaped to provide the poles areas and a flux pipe between the pole areas.

12. A magnetic flux pad having a front face and back face for generating a magnetic flux in a space beyond the front face of the pad, the pad comprising:

    two pole areas for sending or receiving flux,

    a magnetically permeable core,

    two coils associated with the core and being provided on a side of the core adjacent to the front face of the pad,

    whereby the pad produces an arch shaped flux in the space such that essentially each line flux starts on one of the pole areas, arches to the second pole area and joins on itself through the core with essentially no flux present at the back face of the pad.

13. A magnetic flux pad having a front face and back face for receiving a magnetic flux from a space beyond the front face of the pad, the pad comprising:

    two pole areas for sending or receiving flux,

    a magnetically permeable core,

    two coils associated with the core and being provided on a side of the core adjacent to the front face of the pad and adapted to receive a horizontal flux component,

    and a further coil magnetically associated with the core and adapted to receive a vertical flux component.

14. An IPT system including a magnetic flux pad according to any one of the preceding claims.

15. An IPT system including a combination of a magnetic flux pad of claim 12 and magnetic flux pad of claim 13.

16. An IPT system as claimed in claim 14 or claim 15 for supplying power to an electric vehicle.

17. An IPT power transfer magnetic flux transmitter or receiver including:
a magnetic flux carrying member having a relatively high magnetic permeability and having two ends, two windings electromagnetically associated with the flux carrying member, and the flux carrying member having two flux transmission or reception regions, one region being provided adjacent to each end, whereby magnetic flux exists or enters the flux carrying member substantially only at or adjacent to the transmission or reception regions.

18. A magnetic flux pad substantially as herein described with reference to any one of the embodiments shown in the drawings.

19. An IPT system substantially as herein described with reference to the accompanying drawings.

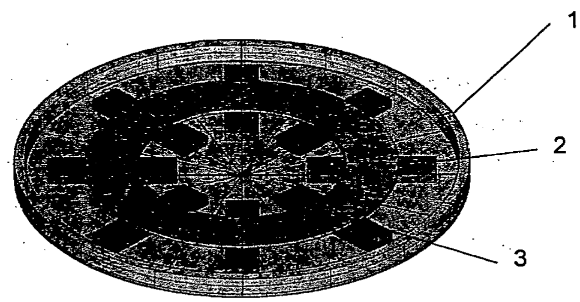


FIGURE 1

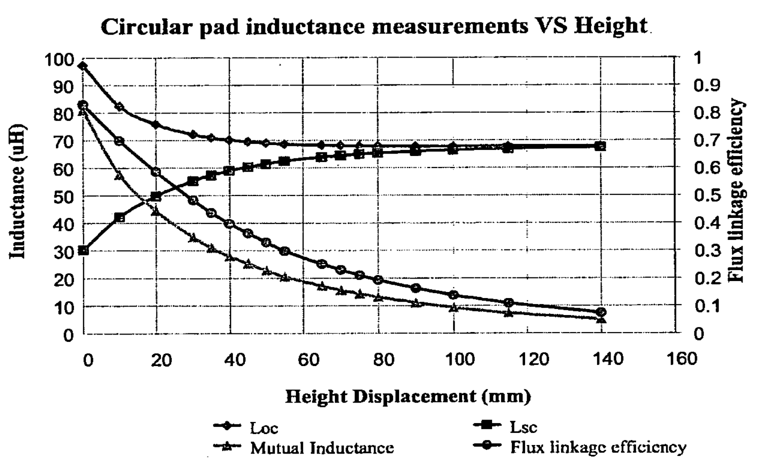


FIGURE 2

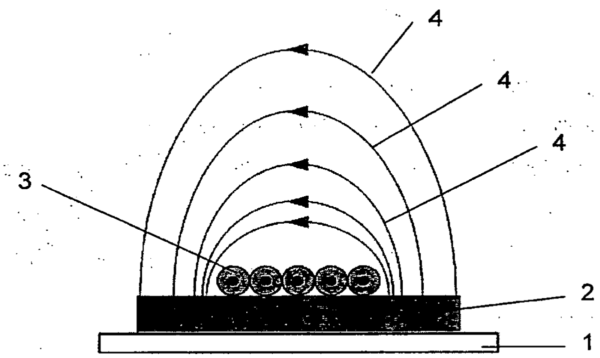


FIGURE 3

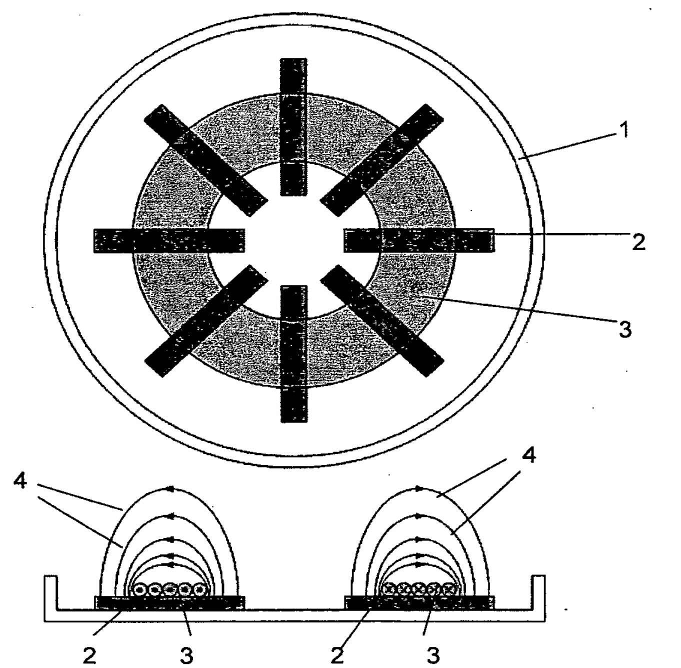


FIGURE 4

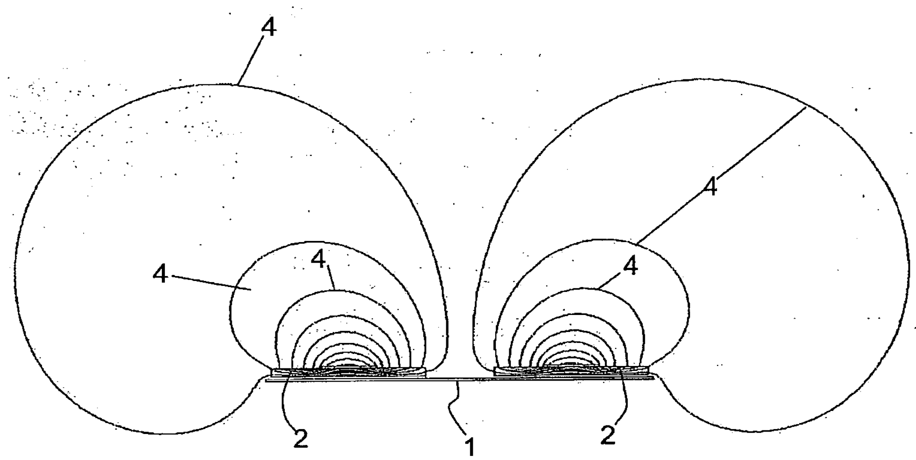


FIGURE 4A

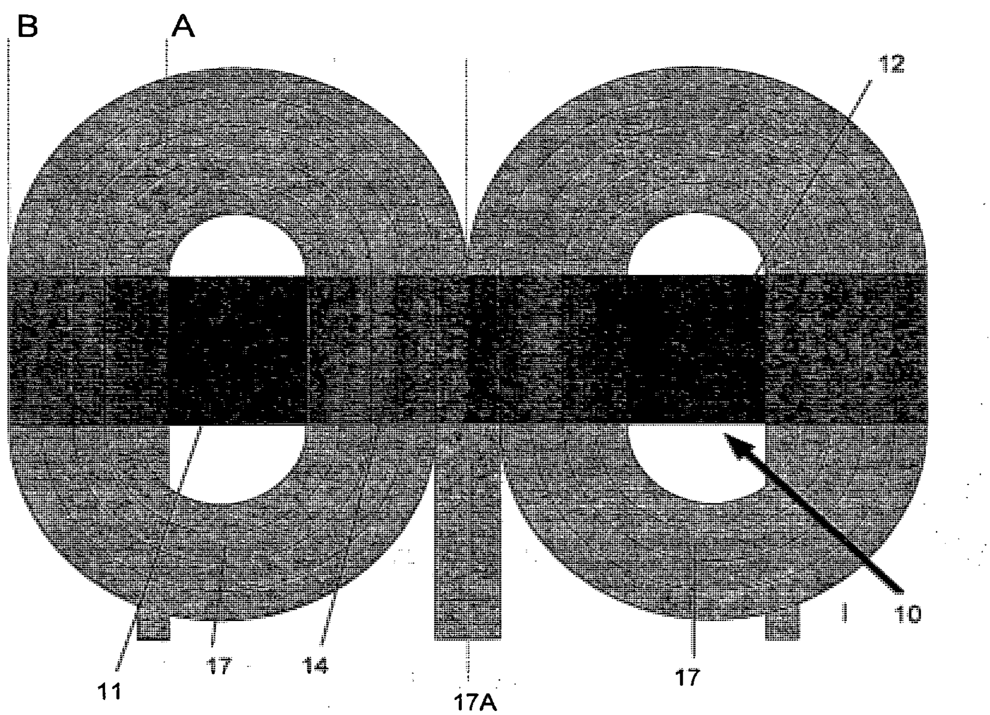


FIGURE 5A

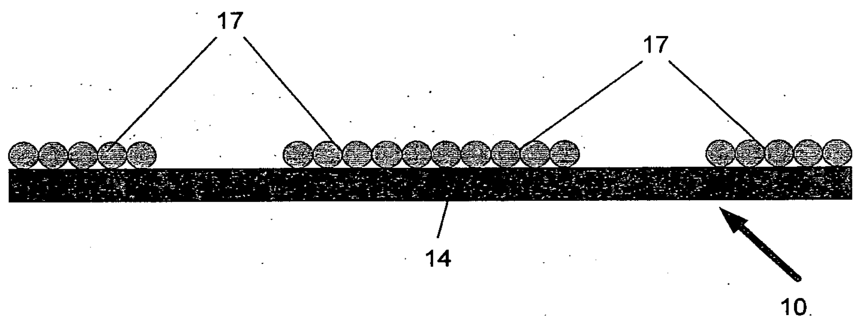


FIGURE 5B

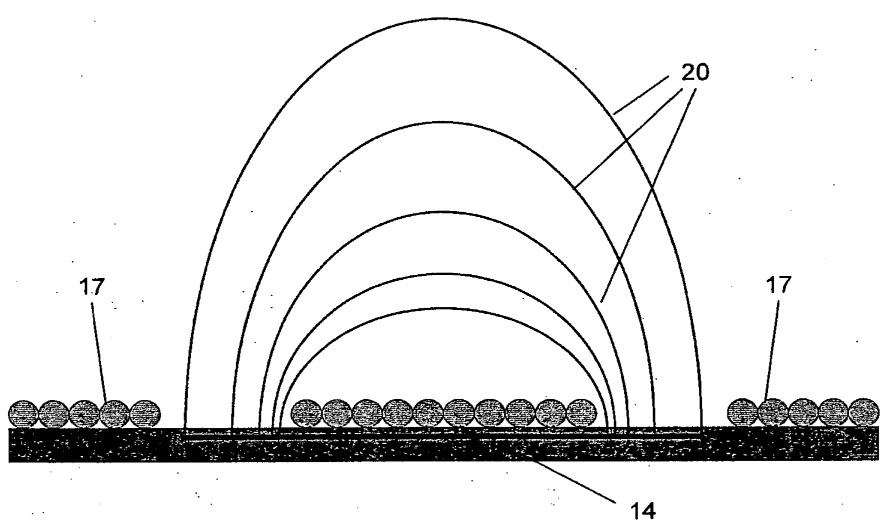


FIGURE 5C

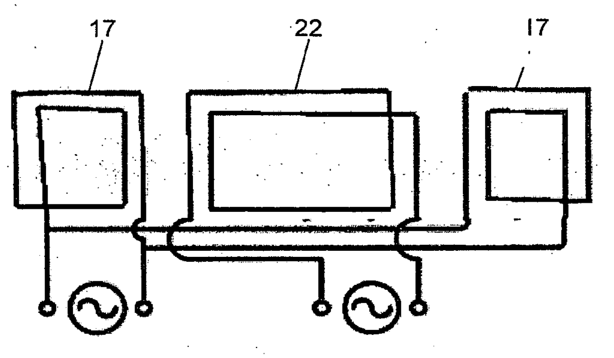


FIGURE 6

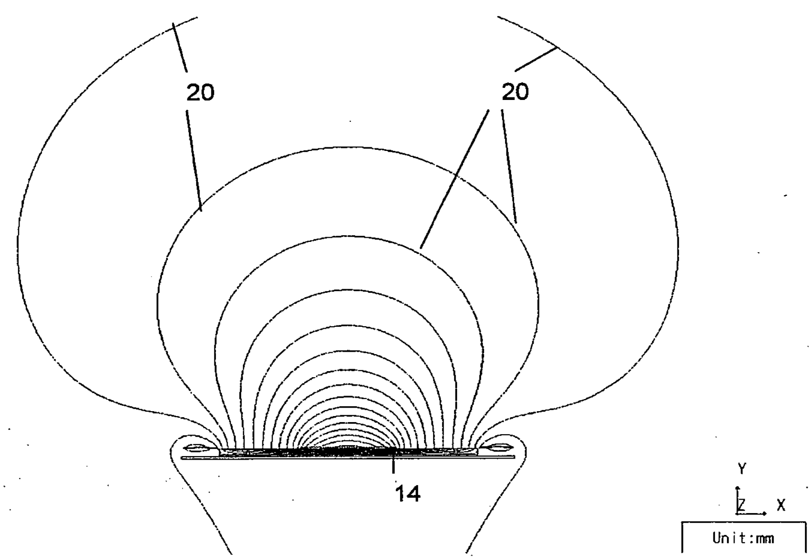


FIGURE 5D

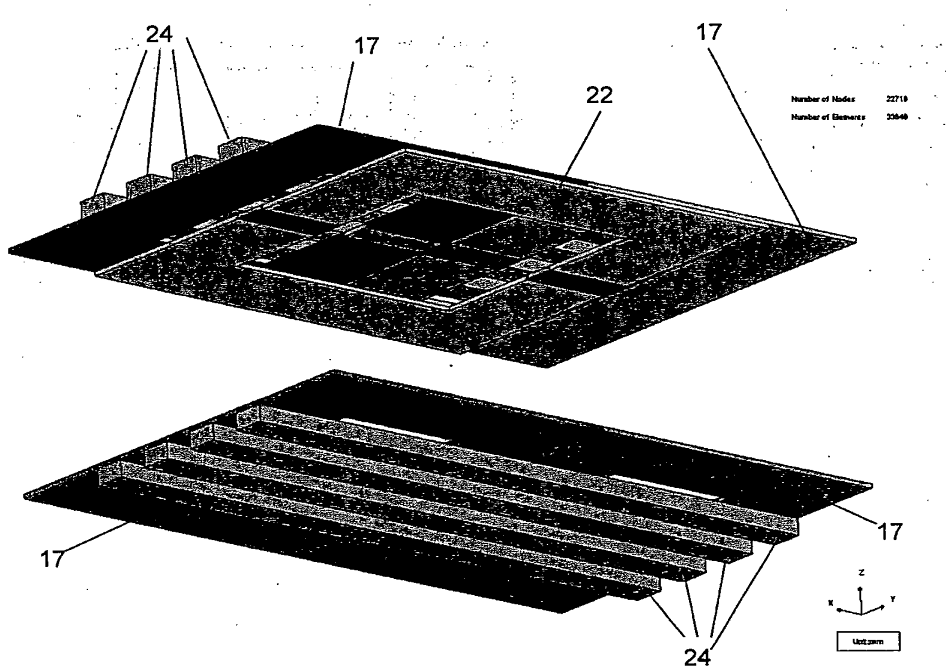


FIGURE 7A

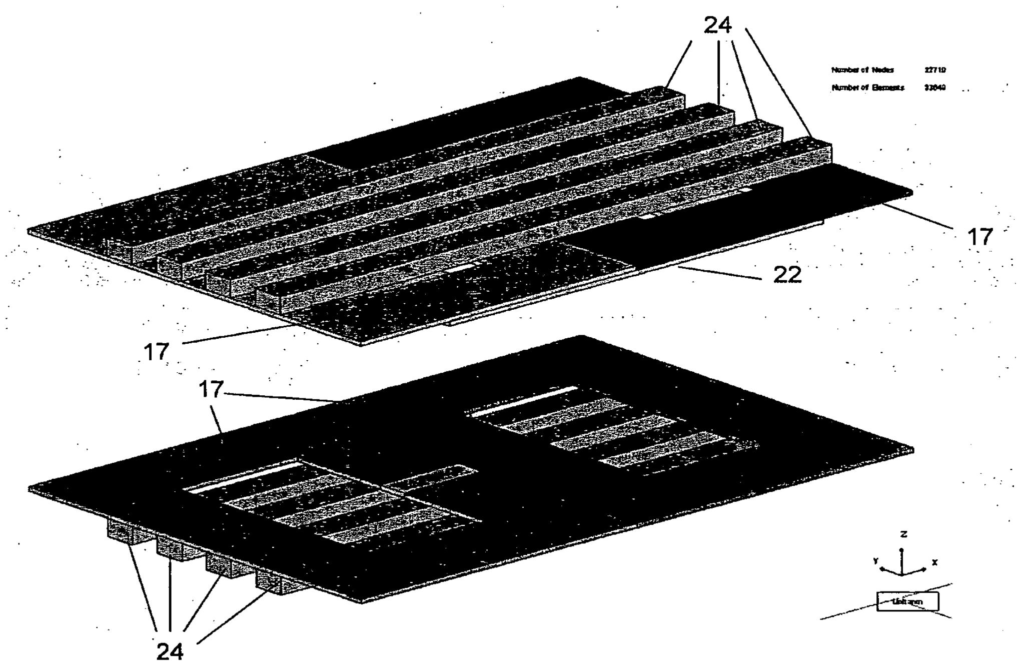


FIGURE 7B

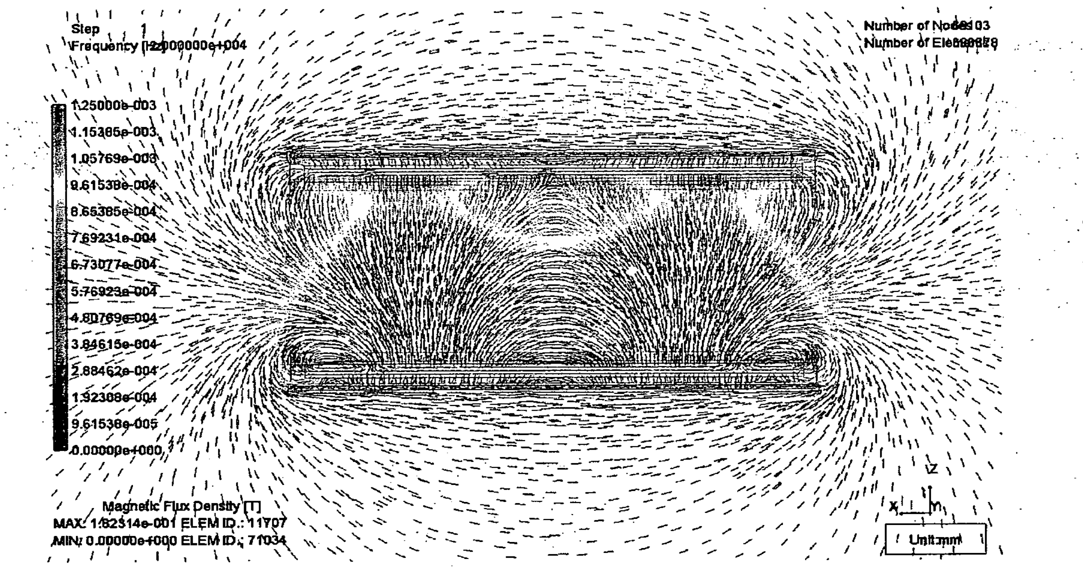


FIGURE 8A

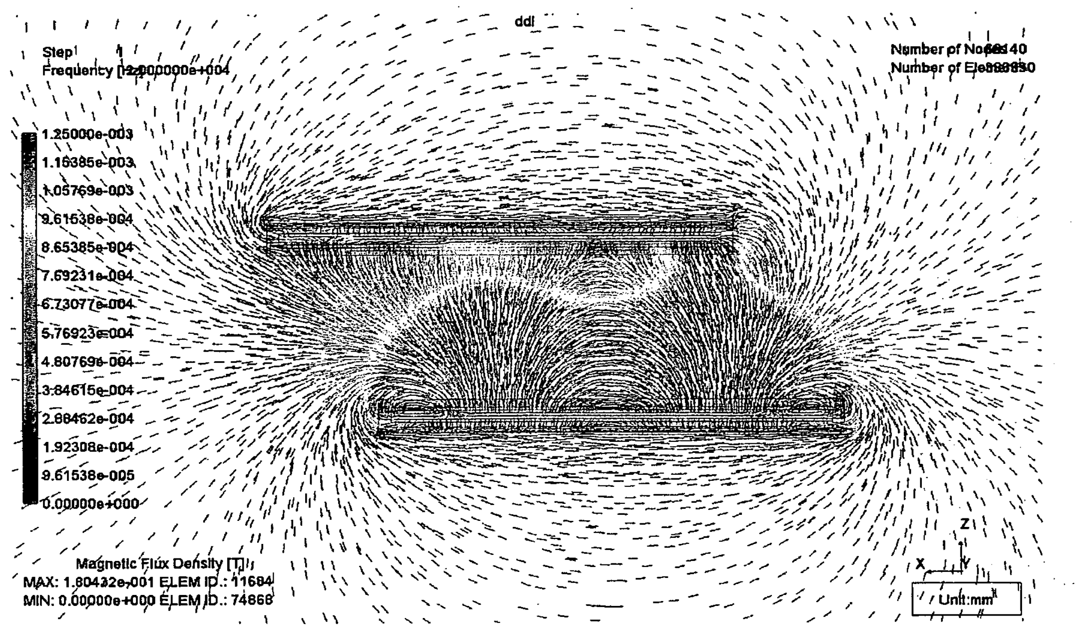


FIGURE 8B

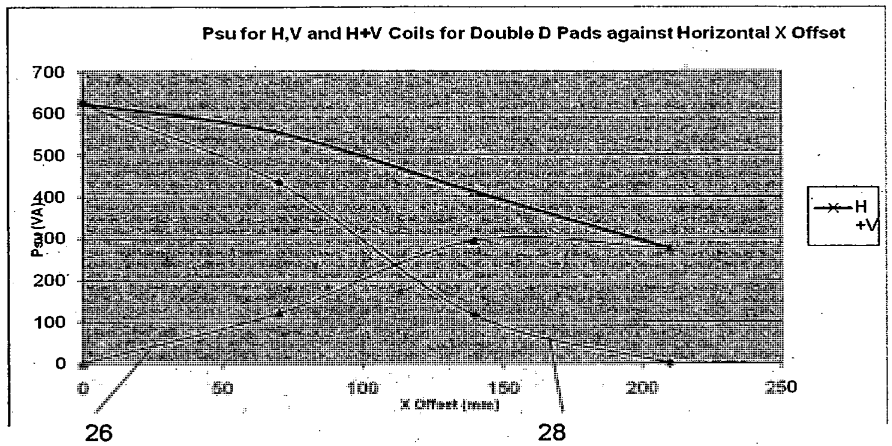


FIGURE 9

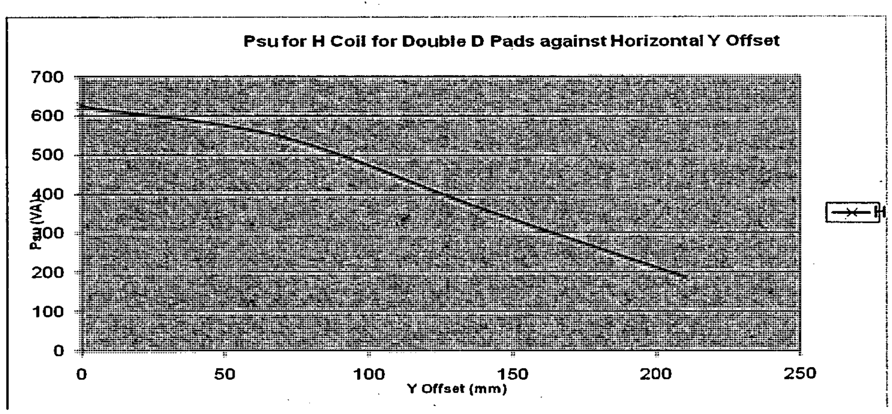


FIGURE 10

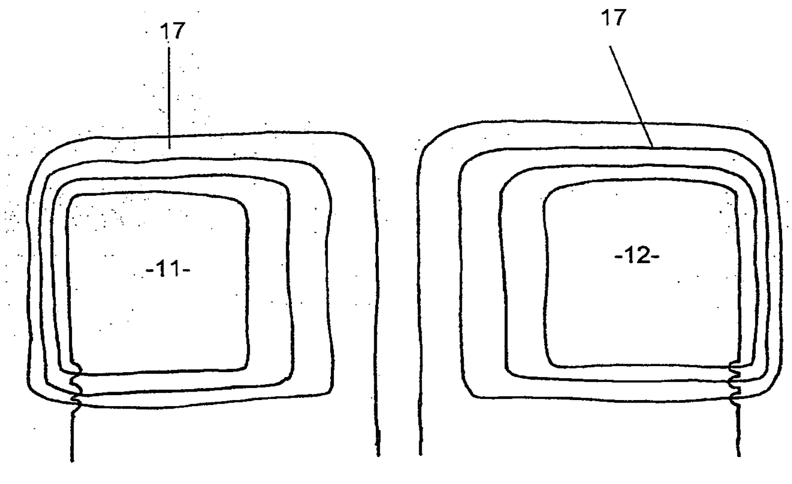


FIGURE 11

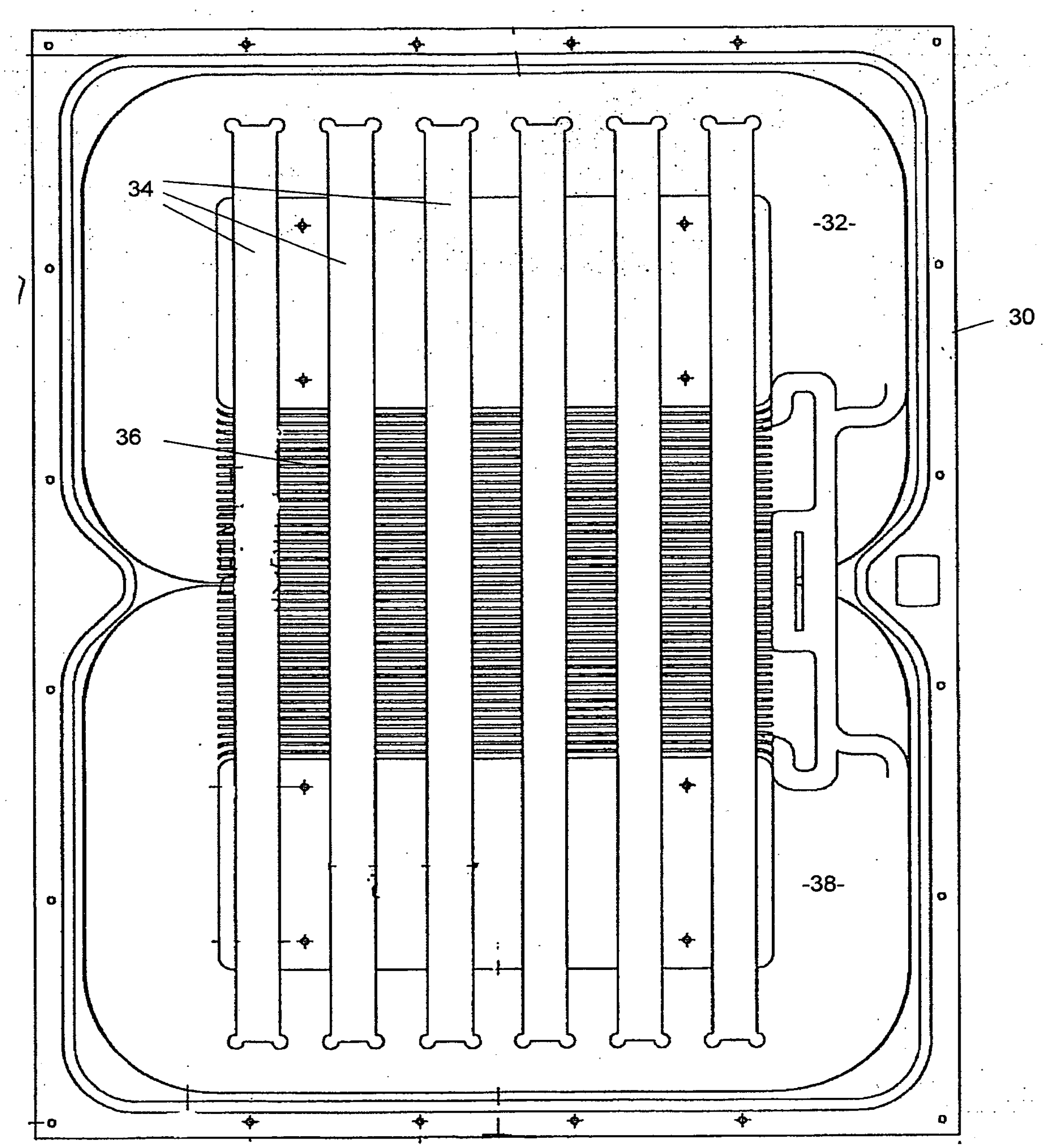


FIGURE 12

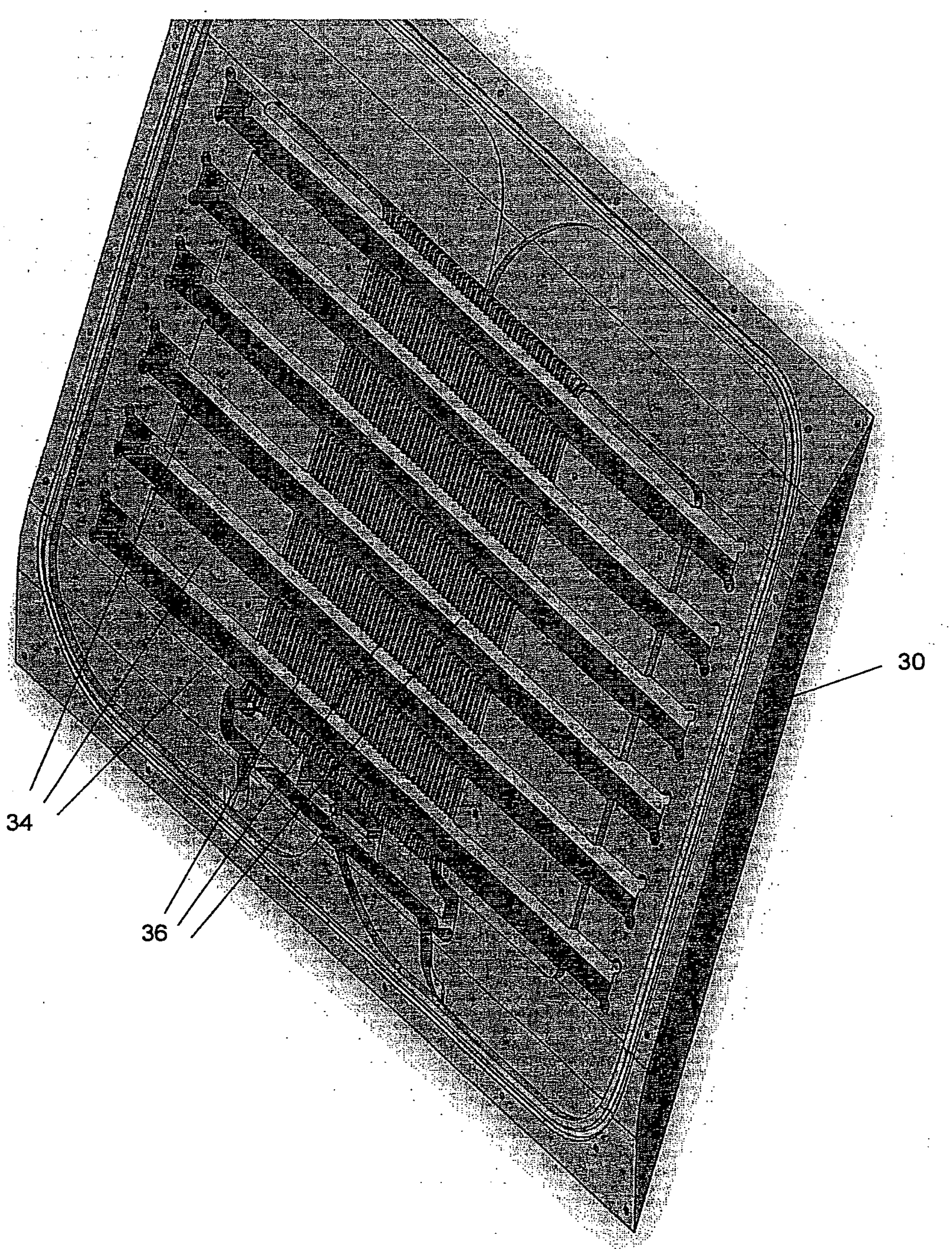


FIGURE 13

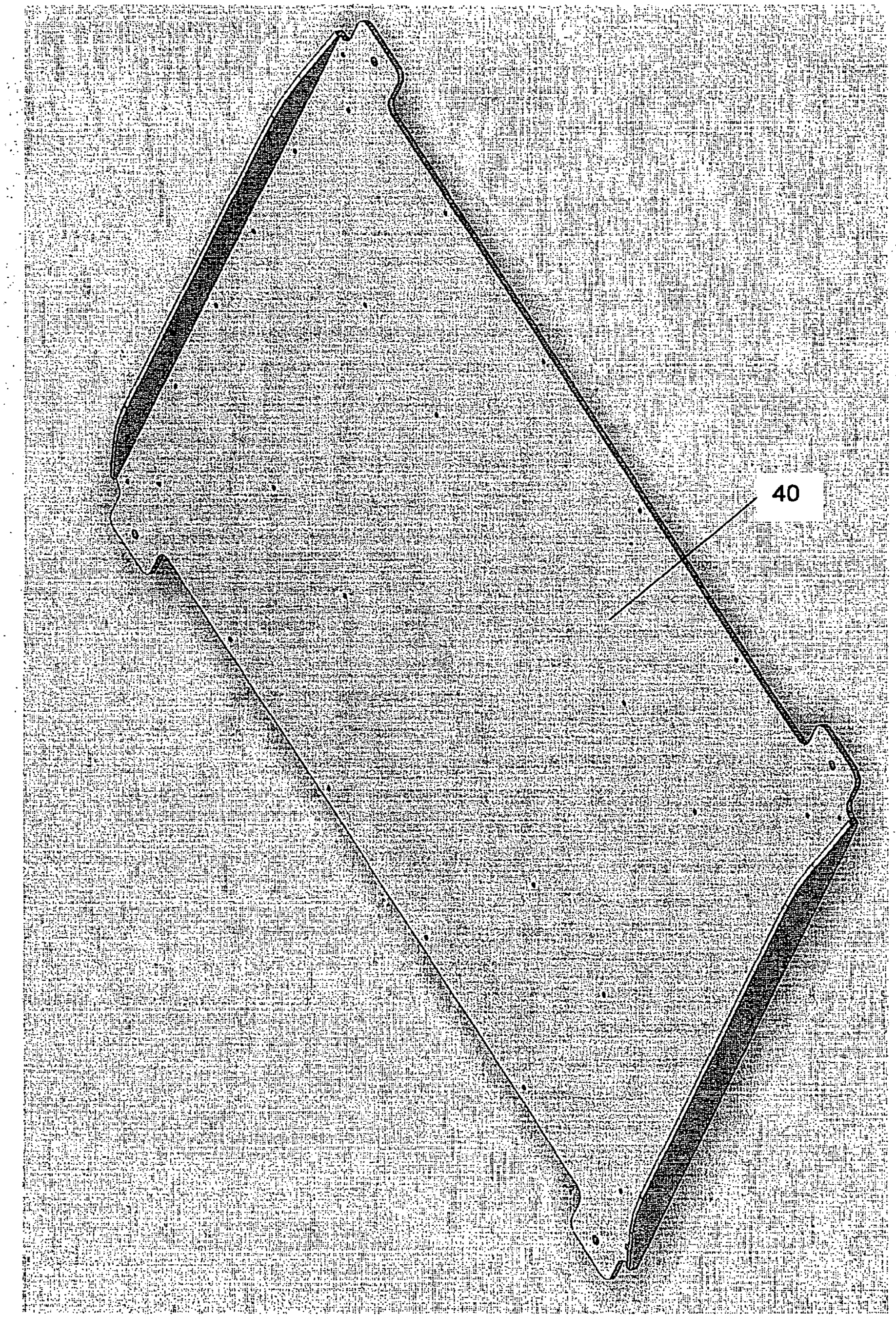


FIGURE 14

INTERNATIONAL SEARCH REPORT

International application No.
PCT/NZ2010/000018

A. CLASSIFICATION OF SUBJECT MATTER

Int. Cl.

H02J 3/00 (2006.01)

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)
WPi: magnet, flux, pad, permeabe, core, power transfer, inductive and similar terms.
Googl Patents & Esp@ce: inductive power transfer amd similar terms

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	WO 2007/126321 A1 (AUKLAND UNISERVICES LIMITED) 8 November 2007 Abstract; figs. 1, 6; page 2, line 18; page 6, lines 5-15	1, 2, 4-8, 10, 13, 14, 17
A	US 7,042,196 B2 (KA-LAI et al.) 9 May 2006 whole document, especially fig. 5; column 19, lines 25-62	
A	WO 2008/140333 A2 (AUKLAND UNISERVICES LIMITED) 20 November 2008 whole document	

☐ Further documents are listed in the continuation of Box C ☒ See patent family annex

* Special categories of cited documents:

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier application or patent but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
31 May 2010	24 JUN 2010

Name and mailing address of the ISA/AU	Authorized officer
AUSTRALIAN PATENT OFFICE PO BOX 200, WODEN ACT 2606, AUSTRALIA E-mail address: pct@ipaustralia.gov.au Facsimile No. +61 2 6283 7999	**JAMES WILLIAMS** AUSTRALIAN PATENT OFFICE (ISO 9001 Quality Certified Service) Telephone No : +61 2 6283 2599

Form PCT/ISA/210 (second sheet) (July 2009)

INTERNATIONAL SEARCH REPORT

International application No.
PCT/NZ2010/000018

Box No. II Observations where certain claims were found unsearchable (Continuation of item 2 of first sheet)

This international search report has not been established in respect of certain claims under Article 17(2)(a) for the following reasons:

1. [X] Claims Nos.: **18,19**

   because they relate to subject matter not required to be searched by this Authority, namely:

   The claims do not comply with Rule 6.2(a) because they rely on references to the description and/or drawings.

2. [ ] Claims Nos.:

   because they relate to parts of the international application that do not comply with the prescribed requirements to such an extent that no meaningful international search can be carried out, specifically:

3. [ ] Claims Nos.:

   because they are dependent claims and are not drafted in accordance with the second and third sentences of Rule 6.4(a)

Box No. III Observations where unity of invention is lacking (Continuation of item 3 of first sheet)

This International Searching Authority found multiple inventions in this international application, as follows:

1. [ ] As all required additional search fees were timely paid by the applicant, this international search report covers all searchable claims.
2. [ ] As all searchable claims could be searched without effort justifying additional fees, this Authority did not invite payment of additional fees.
3. [ ] As only some of the required additional search fees were timely paid by the applicant, this international search report covers only those claims for which fees were paid, specifically claims Nos.:
4. [ ] No required additional search fees were timely paid by the applicant. Consequently, this international search report is restricted to the invention first mentioned in the claims; it is covered by claims Nos.:

Remark on Protest

[ ] The additional search fees were accompanied by the applicant's protest and, where applicable, the payment of a protest fee.

[ ] The additional search fees were accompanied by the applicant's protest but the applicable protest fee was not paid within the time limit specified in the invitation.

[ ] No protest accompanied the payment of additional search fees.

INTERNATIONAL SEARCH REPORT

Information on patent family members

International application No.
PCT/NZ2010/000018

This Annex lists the known "A" publication level patent family members relating to the patent documents cited in the above-mentioned international search report. The Australian Patent Office is in no way liable for these particulars which are merely given for the purpose of information.

Patent Document Cited in Search Report		Patent Family Member					
WO	2007/126321	CN	101461114	EP	2013957	NZ	546955
		US	2009303749				
US	7042196	AU	2003233895	AU	2003240999	AU	2003282214
		AU	2008255158	CN	1653669	CN	101699708
		CN	101699709	CN	101699710	CN	101699711
		EP	1506554	EP	1506605	GB	2398176
		GB	2388715	GB	2388716	GB	2399225
		GB	2399226	GB	2399227	GB	2399228
		GB	2399229	GB	2399230	JP	2009010394
		US	2003210106	US	6906495	US	2005140482
		US	2005135122	US	7239110	US	2006076922
		US	7248017	US	2005116683	US	7525283
		US	2006061323	US	7622891	US	2009189565
		US	7714537	US	2009096414	WO	03096361
		WO	03096512	WO	2004038888	ZA	200408863
WO	2008/140333	AU	2008251143	CA	2687060	CN	101689761
		EP	2156532	KR	20100017582	NZ	555128
		US	2010109604				

Due to data integration issues this family listing may not include 10 digit Australian applications filed since May 2001.

END OF ANNEX

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
19 November 2009 (19.11.2009)

PCT

(10) International Publication Number
WO 2009/140506 A1

(51) International Patent Classification:
H01P 7/00 (2006.01)

(21) International Application Number:
PCT/US2009/043970

(22) International Filing Date:
14 May 2009 (14.05.2009)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
61/127,661 14 May 2008 (14.05.2008) US

(71) Applicant *(for all designated States except US)*: **MASSACHUSETTS INSTITUTE OF TECHNOLOGY** [US/US]; 77 Massachusetts Avenue, Cambridge, Massachusetts 02142-1324 (US).

(72) Inventors; and
(75) Inventors/Applicants *(for US only)*: **KARALIS, Aristeidis** [GR/US]; 151 Tremont Street, Apt. 21F, Boston, Massachusetts 02111 (US). **HAMAM, Rafif, E.** [LB/US]; 550 Memorial Drive, Apt. 9C-2, Cambridge, Massachusetts 02139 (US). **JOANNOPOULOS, John, D.** [US/US]; 64 Douglas Road, Belmont, Massachusetts 02478 (US). **SOLJACIC, Marin** [HR/US]; 44 Westlund Road, Belmont, Massachusetts 02478 (US).

(74) Agent: **WEFERS, Marc, M.**; Fish & Richardson P.C., PO Box 1022, Minneapolis, Minnesota 55440-1022 (US).

(81) Designated States *(unless otherwise indicated, for every kind of national protection available)*: AE, AG, AL, AM, AO, AT, AU, AZ, BA, BB, BG, BH, BR, BW, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DO, DZ, EC, EE, EG, ES, FI, GB, GD, GE, GH, GM, GT, HN, HR, HU, ID, IL, IN, IS, JP, KE, KG, KM, KN, KP, KR, KZ, LA, LC, LK, LR, LS, LT, LU, LY, MA, MD, ME, MG, MK, MN, MW, MX, MY, MZ, NA, NG, NI, NO, NZ, OM, PG, PH, PL, PT, RO, RS, RU, SC, SD, SE, SG, SK, SL, SM, ST, SV, SY, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VC, VN, ZA, ZM, ZW.

(84) Designated States *(unless otherwise indicated, for every kind of regional protection available)*: ARIPO (BW, GH, GM, KE, LS, MW, MZ, NA, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European (AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, HR, HU, IE, IS, IT, LT, LU, LV, MC, MK, MT, NL, NO, PL, PT, RO, SE, SI, SK, TR), OAPI (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Published:
— *with international search report (Art. 21(3))*

(54) Title: WIRELESS ENERGY TRANSFER, INCLUDING INTERFERENCE ENHANCEMENT

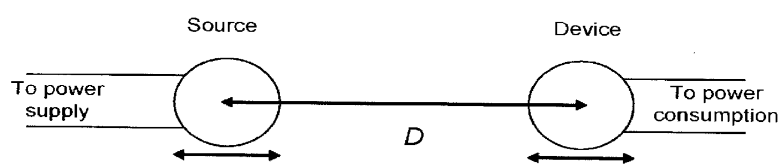


FIG. 1

(57) Abstract: Disclosed is an apparatus for use in wireless energy transfer, which includes a first resonator structure configured for energy transfer with a second resonator structure over a distance D larger than characteristic sizes, [insert formula] and [insert formula], of the first and second resonator structures. A power generator is coupled to the first structure and configured to drive the first resonator structure or the second resonator structure at an angular frequency away from the resonance angular frequencies and shifted towards a frequency corresponding to an odd normal mode for the resonator structures to reduce radiation from the resonator structures by destructive far-field interference.

WO 2009/140506 A1

WIRELESS ENERGY TRANSFER, INCLUDING INTERFERENCE ENHANCEMENT

CROSS REFERENCE TO RELATED APPLICATIONS

Pursuant to U.S.C. § 119(e), this application claims priority to U.S. Provisional Application Serial No. 61/127,661, filed May 14, 2008.

This application is also related by subject matter to the following commonly owned applications: U.S. Utility Patent Application Serial No. 12/055,963, filed March 26, 2008; U.S. Utility Patent Application Serial No. 11/481,077, filed July 5, 2006; U.S. Provisional Application Serial No. 60/698,442, filed July 12, 2005; U.S. Provisional Application Serial No. 60/908,383, filed March 27, 2007; U.S. Provisional Application Serial No. 60/908,666, filed March 28, 2007; and International Application No. PCT/US2007/070892, filed June 11, 2007.

The contents of the prior applications are incorporated herein by reference in their entirety.

BACKGROUND

The disclosure relates to wireless energy transfer. Wireless energy transfer can for example, be useful in such applications as providing power to autonomous electrical or electronic devices.

Radiative modes of omni-directional antennas (which work very well for information transfer) are not suitable for such energy transfer, because a vast majority of energy is wasted into free space. Directed radiation modes, using lasers or highly-directional antennas, can be efficiently used for energy transfer, even for long distances (transfer distance $L_{TRANS} \gg L_{DEV}$, where L_{DEV} is the characteristic size of the device and/or the source), but require existence of an uninterruptible line-of-sight and a complicated tracking system in the case of mobile objects. Some transfer schemes rely on induction, but are typically restricted to very close-range ($L_{TRANS} \ll L_{DEV}$) or low power (~mW) energy transfers.

The rapid development of autonomous electronics of recent years (e.g. laptops, cell-phones, house-hold robots, that all typically rely on chemical energy storage) has led to an increased need for wireless energy transfer.

SUMMARY

Efficient wireless energy-transfer between two resonant objects can be achieved at mid-range distances, provided these resonant objects are designed to operate in the 'strong-coupling' regime. We describe an implementation of a method to increase the efficiency of energy-transfer or to suppress the power radiated, which can be harmful or a cause of interference to other communication systems, by utilizing destructive interference between the radiated far-fields of the resonant coupled objects. 'Strong coupling' is a necessary condition for efficient energy-transfer, in the absence of far-field interference. 'Strong coupling' can be demonstrated in the case of realistic systems: self-resonant conducting coils, capacitively-loaded conducting coils, inductively-loaded conducting rods and dielectric disks, all bearing high-Q electromagnetic resonant modes. Also, an analytical model can be developed to take far-field interference into account for wireless energy-transfer systems. The analytical model can be used to demonstrate the efficiency enhancement and radiation suppression, in the presence of interference. In an example implementation, we describe improved performance based on the above principles in the case of two realistic systems: capacitively-loaded conducting coils and dielectric disks, both bearing high-Q electromagnetic resonant modes and far-field interference.

In an aspect, an apparatus for use in wireless energy transfer includes a first resonator structure configured for energy transfer with a second resonator structure, over a distance D larger than a characteristic size L_1 of said first resonator structure and larger than a characteristic size L_2 of said second resonator structure. The energy transfer has a rate κ and is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure. The resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$ at least larger than 300, and the said resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$ at least larger than 300. The absolute value of the difference of said angular frequencies ω_1 and ω_2 is smaller than the broader of said resonant widths Γ_1 and

Γ_2, and the quantity $\kappa/\sqrt{\Gamma_1\Gamma_2}$ is at least larger than 20. The apparatus also includes a power supply coupled to the first structure and configured to drive the first resonator structure or the second resonator structure at an angular frequency away from the resonance angular frequencies and shifted towards a frequency corresponding to an odd normal mode for the resonator structures to reduce radiation from the resonator structures by destructive far-field interference.

In some examples, the power supply is configured to drive the first resonator structure or the second resonator structure at the angular frequency away from the resonance angular frequencies and shifted towards the frequency corresponding to an odd normal mode for the resonator structures to substantially suppress radiation from the resonator structures by destructive far-field interference.

In an aspect, a method for wireless energy transfer involves a first resonator structure configured for energy transfer with a second resonator structure, over a distance D larger than a characteristic size L_1 of said first resonator structure and larger than a characteristic size L_2 of said second resonator structure, wherein the energy transfer has a rate κ and is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, said resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1/2\Gamma_1$ at least larger than 300, and said resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2/2\Gamma_2$ at least larger than 300, the absolute value of the difference of said angular frequencies ω_1 and ω_2 is smaller than the broader of said resonant widths Γ_1 and Γ_2, and the quantity $\kappa/\sqrt{\Gamma_1\Gamma_2}$ is at least larger than 20. The method includes driving the first resonator structure or the second resonator structure at an angular frequency away from the resonance angular frequencies and shifted towards a frequency corresponding to an odd normal mode for the resonator structures to reduce radiation from the resonator structures by destructive far-field interference.

In some examples, the first resonator structure or the second resonator structure is driven at the angular frequency away from the resonance angular frequencies and shifted towards the frequency corresponding to an odd normal mode for the resonator structures to substantially suppress radiation from the resonator structures by destructive far-field interference.

In an aspect, an apparatus for use in wireless energy transfer includes a first resonator structure configured for energy transfer with a second resonator structure, over a distance *D* larger than a characteristic size L_1 of said first resonator structure and larger than a characteristic size L_2 of said second resonator structure. The energy transfer has a rate κ and is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure. The resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$ at least larger than 300, and the resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$ at least larger than 300. The absolute value of the difference of said angular frequencies ω_1 and ω_2 is smaller than the broader of said resonant widths Γ_1 and Γ_2, and the quantity $\kappa / \sqrt{\Gamma_1 \Gamma_2}$ is at least larger than 20. For a desired range of the distances *D*, the resonance angular frequencies for the resonator structures increase transmission efficiency *T* by accounting for radiative interference, wherein the increase is relative to a transmission efficiency *T* calculated without accounting for the radiative interference.

In some examples, the resonance angular frequencies for the resonator structures are selected by optimizing the transmission efficiency T to account for both a resonance quality factor *U* and an interference factor *V*.

In an aspect, a method involves designing a wireless energy transfer apparatus, the apparatus including a first resonator structure configured for energy transfer with a second resonator structure, over a distance *D* larger than a characteristic size L_1 of said first resonator structure and larger than a characteristic size L_2 of said second resonator structure, wherein the energy transfer has a rate κ and is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, wherein said resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$ at least larger than 300, and said resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$ at least larger than 300, wherein the absolute value of the difference of said angular frequencies ω_1 and ω_2 is smaller than the broader of said resonant widths Γ_1 and Γ_2, and the quantity $\kappa / \sqrt{\Gamma_1 \Gamma_2}$ is at least larger than 20. The method includes selecting the resonance angular frequencies for the resonator structures to substantially optimize the transmission efficiency by accounting for radiative interference between the resonator structures.

In some examples, the resonance angular frequencies for the resonator structures are selected by optimizing the transmission efficiency T to account for both a resonance quality factor *U* and an interference factor *V*.

In an aspect, an apparatus for use in wireless energy transfer includes a first resonator structure configured for energy transfer with a second resonator structure over a distance *D*. The energy transfer is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, with a coupling factor *k*. The resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$, and is radiative in the far field, with an associated radiation quality factor $Q_{1,\text{rad}} \geq Q_1$, and the resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$, and is radiative in the far field, with an associated radiation quality factor $Q_{2,\text{rad}} \geq Q_2$. An absolute value of a difference of said angular frequencies ω_1 and ω_2 is smaller than broader of said resonant widths Γ_1 and Γ_2, and an average resonant angular frequency is defined as $\omega_o = \sqrt{\omega_1 \omega_2}$, corresponding to an average resonant wavelength $\lambda_o = 2\pi c / \omega_o$, where *c* is the speed of light in free space, and a strong-coupling factor being defined as $U = k\sqrt{Q_1 Q_2}$. The apparatus is configured to employ interference between said radiative far fields of the resonant fields of the first and second resonator, with an interference factor V_{rad}, to reduce a total amount of radiation from the apparatus compared to an amount of radiation from the apparatus in the absence of interference, a strong-interference factor being defined as $V = V_{\text{rad}}\sqrt{(Q_1 / Q_{1,\text{rad}})(Q_2 / Q_{2,\text{rad}})}$.

The following are examples within the scope of this aspect.

The apparatus has $Q_1 / Q_{1,\text{rad}} \geq 0.01$ and $Q_2 / Q_{2,\text{rad}} \geq 0.01$. The apparatus has $Q_1 / Q_{1,\text{rad}} \geq 0.1$ and $Q_2 / Q_{2,\text{rad}} \geq 0.1$. The apparatus has D / λ_o larger than 0.001 and the strong-interference factor V is larger than 0.01. The apparatus has D / λ_o larger than 0.001 and the strong-interference factor V is larger than 0.1. The apparatus includes the second resonator structure.

During operation, a power generator is coupled to one of the first and second resonant structure, with a coupling rate κ_g, and is configured to drive the resonator structure, to which it is coupled, at a driving frequency f, corresponding to a driving angular frequency $\omega = 2\pi f$, wherein U_g is defined as κ_g / Γ_1, if the power generator is coupled to the first resonator structure and defined as κ_g / Γ_2, if the power generator is coupled to the second resonator structure. The driving frequency is different from the resonance frequencies of the first and second resonator structures and is closer to a frequency corresponding to an odd normal mode of the system of the two resonator structures, wherein the detuning of the first resonator from the driving frequency is defined as $D_1 = (\omega - \omega_1) / \Gamma_1$ and the detuning of the second resonator structure from the driving frequency is defined as $D_2 = (\omega - \omega_2) / \Gamma_2$.

D_1 is approximately equal to UV_{rad} and D_2 is approximately equal to UV_{rad}. U_g is chosen to maximize the ratio of the energy-transfer efficiency to the radiation efficiency. U_g is approximately equal to $\sqrt{1 + U^2 - V_{\text{rad}}^2 U^2 + V^2 - 2VV_{\text{rad}}}$. f is at least larger than 100 kHz and smaller than 500MHz. f is at least larger than 1MHz and smaller than 50MHz. The apparatus further includes the power generator. During operation, a power load is coupled to the resonant structure to which the power generator is not coupled, with a coupling rate κ_l, and is configured to receive from the resonator structure, to which it is coupled, a usable power, wherein U_l is defined as κ_l / Γ_1, if the power load is coupled to the first resonator structure and defined as κ_l / Γ_2, if the power load is coupled to the second resonator structure. U_l is chosen to maximize the ratio of the energy-transfer efficiency to the radiation efficiency. The driving frequency is different from the resonance frequencies of the first and second resonator structures and is closer to a frequency corresponding to an odd normal mode of the system of the two resonator structures, wherein the detuning of the first resonator from the driving frequency is defined as $D_1 = (\omega - \omega_1) / \Gamma_1$ and is approximately equal to UV_{rad}, and the detuning of the second resonator structure from the driving frequency is defined as $D_2 = (\omega - \omega_2) / \Gamma_2$ and is approximately equal to UV_{rad}, and U_l is approximately equal to $\sqrt{1 + U^2 - V_{\text{rad}}^2 U^2 + V^2 - 2VV_{\text{rad}}}$.

At least one of the first and second resonator structures comprises a capacitively loaded loop or coil of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon. The characteristic size of said loop or coil is less than 30 cm and the width of said conducting wire or Litz wire or ribbon is less than 2cm. The characteristic size of said loop or coil is less than 1m and the width of said conducting wire or Litz wire or ribbon is less than 2cm.

The apparatus further includes a feedback mechanism for maintaining the resonant frequency of one or more of the resonant objects. The feedback mechanism comprises an oscillator with a fixed driving frequency and is configured to adjust the resonant frequency of the one or more resonant objects to be detuned by a fixed amount with respect to the fixed frequency.

In an aspect, an apparatus for use in wireless energy transfer includes a first resonator structure configured for energy transfer with a second resonator structure over a distance *D*. The energy transfer is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, with a coupling factor k. The resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$, and is radiative in the far field, with an associated radiation quality factor $Q_{1,\text{rad}} \geq Q_1$, and the resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$, and is radiative in the far field, with an associated radiation quality factor $Q_{2,\text{rad}} \geq Q_2$. An absolute value of a difference of said angular frequencies ω_1 and ω_2 is smaller than the broader of said resonant widths Γ_1 and Γ_2, and an average resonant angular frequency is defined as $\omega_o = \sqrt{\omega_1\omega_2}$, corresponding to an average resonant wavelength $\lambda_o = 2\pi c / \omega_o$, where c is the speed of light in free space, and a strong-coupling factor is defined as $U = k\sqrt{Q_1Q_2}$. The apparatus is configured to employ interference between said radiative far fields of the resonant fields of the first and second resonator, with an interference factor V_{rad}, to increase efficiency of energy transfer for the apparatus compared to efficiency for the apparatus in the absence of interference, the strong-interference factor being defined as $V = V_{\text{rad}}\sqrt{\left(Q_1 / Q_{1,\text{rad}}\right)\left(Q_2 / Q_{2,\text{rad}}\right)}$.

The following are examples within the scope of this aspect.

The apparatus has $Q_1 / Q_{1,\text{rad}} \geq 0.05$ and $Q_2 / Q_{2,\text{rad}} \geq 0.05$. The apparatus has $Q_1 / Q_{1,\text{rad}} \geq 0.5$ and $Q_2 / Q_{2,\text{rad}} \geq 0.5$. The apparatus has D / λ_o larger than 0.01 and the strong-interference factor V is larger than 0.05. The apparatus has D / λ_o larger than 0.01 and the strong-interference factor V is larger than 0.5. The apparatus further includes the second resonator structure.

During operation, a power generator is coupled to one of the first and second resonant structure, with a coupling rate κ_g, and is configured to drive the resonator structure, to which it is coupled, at a driving frequency f, corresponding to a driving angular frequency $\omega = 2\pi f$, wherein U_g is defined as κ_g / Γ_1, if the power generator is coupled to the first resonator structure and defined as κ_g / Γ_2, if the power generator is coupled to the second resonator structure. The driving frequency is different from the resonance frequencies of the first and second resonator structures and is closer to a frequency corresponding to an odd normal mode of the system of the two resonator structures, wherein the detuning of the first resonator from the driving frequency is defined as $D_1 = (\omega - \omega_1)/\Gamma_1$ and the detuning of the second resonator structure from the driving frequency is defined as $D_2 = (\omega - \omega_2)/\Gamma_2$.

D_1 is approximately equal to UV and D_2 is approximately equal to UV. U_g is chosen to maximize the energy-transfer efficiency. U_g is approximately equal to $\sqrt{(1+U^2)(1-V^2)}$. f is at least larger than 100 kHz and smaller than 500MHz. f is at least larger than 1MHz and smaller than 50MHz. The apparatus further includes the power generator.

During operation, a power load is coupled to the resonant structure to which the power generator is not coupled, with a coupling rate κ_l, and is configured to receive from the resonator structure, to which it is coupled, a usable power, wherein U_l is defined as κ_l / Γ_1, if the power load is coupled to the first resonator structure and defined as κ_l / Γ_2, if the power load is coupled to the second resonator structure. U_l is chosen to maximize the energy-transfer efficiency. The driving frequency is different from the resonance frequencies of the first and second resonator structures and is closer to a frequency corresponding to an odd normal mode of the system of the two resonator structures, wherein the detuning of the first resonator from the driving frequency is defined as $D_1 = (\omega - \omega_1)/\Gamma_1$ and is approximately equal to UV, and the detuning of the second resonator structure from the driving frequency is defined as $D_2 = (\omega - \omega_2)/\Gamma_2$ and is approximately equal to UV, and U_l is approximately equal to $\sqrt{(1+U^2)(1-V^2)}$.

At least one of the first and second resonator structures comprises a capacitively loaded loop or coil of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon. The characteristic size of said loop or coil is less than 30 cm and the width of said conducting wire or Litz wire or ribbon is less than 2cm. The characteristic size of said loop or coil is less than 1m and the width of said conducting wire or Litz wire or ribbon is less than 2cm. The apparatus includes a feedback mechanism for maintaining the resonant frequency of one or more of the resonant objects. The feedback mechanism comprises an oscillator with a fixed driving frequency and is configured to adjust the resonant frequency of the one or more resonant objects to be detuned by a fixed amount with respect to the fixed frequency. The feedback mechanism is configured to monitor an efficiency of the energy transfer, and adjust the resonant frequency of the one or more resonant objects to maximize the efficiency. The resonance angular frequencies for the resonator structures are selected to optimize the energy-transfer efficiency by accounting for both the strong-coupling factor U and the strong-interference interference factor V.

In an aspect, a method for wireless energy transfer includes providing a first resonator structure configured for energy transfer with a second resonator structure over a distance D, wherein the energy transfer is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, with a coupling factor k, wherein said resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$, and is radiative in the far field, with an associated radiation quality factor $Q_{1,\text{rad}} \geq Q_1$, and resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$, and is radiative in the far field, with an associated radiation quality factor $Q_{2,\text{rad}} \geq Q_2$, wherein an absolute value of a difference of said angular frequencies ω_1 and ω_2 is smaller than broader of said resonant widths Γ_1 and Γ_2, and an average resonant angular frequency is defined as $\omega_o = \sqrt{\omega_1\omega_2}$, corresponding to an average resonant wavelength $\lambda_o = 2\pi c / \omega_o$, where c is the speed of light in free space, and the strong-coupling factor is defined as $U = k\sqrt{Q_1 Q_2}$, and employing interference between said radiative far fields of the resonant fields of the first and second resonator, with an interference factor V_{rad}, to reduce a total amount of radiation from the first and second resonator compared to an amount of radiation from the first and second resonator in the absence of interference, a strong-interference factor being defined as

$V = V_{\text{rad}}\sqrt{(Q_1 / Q_{1,\text{rad}})(Q_2 / Q_{2,\text{rad}})}$.

The following are examples within the scope of this aspect.

The method has $Q_1 / Q_{1,\text{rad}} \geq 0.01$ and $Q_2 / Q_{2,\text{rad}} \geq 0.01$. During operation, a power generator is coupled to one of the first and second resonant structure and is configured to drive the resonator structure, to which it is coupled, at a driving frequency *f*, corresponding to a driving angular frequency $\omega = 2\pi f$, wherein the driving frequency is different from the resonance frequencies of the first and second resonator structures and is closer to a frequency corresponding to an odd normal mode of the system of the two resonator structures. During operation, a power load is coupled to the resonant structure to which the power generator is not coupled and is configured to receive from the resonator structure, to which it is coupled, a usable power.

In an aspect, a method for wireless energy transfer includes providing a first resonator structure configured for energy transfer with a second resonator structure over a distance *D*, wherein the energy transfer is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, with a coupling factor *k*, wherein said resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$, and is radiative in the far field, with an associated radiation quality factor $Q_{1,\text{rad}} \geq Q_1$, and said resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$, and is radiative in the far field, with an associated radiation quality factor $Q_{2,\text{rad}} \geq Q_2$, wherein an absolute value of the difference of said angular frequencies ω_1 and ω_2 is smaller than the broader of said resonant widths Γ_1 and Γ_2,and an average resonant angular frequency is defined as $\omega_o = \sqrt{\omega_1\omega_2}$, corresponding to an average resonant wavelength $\lambda_o = 2\pi c / \omega_o$, where *c* is the speed of light in free space, and the strong-coupling factor is defined as $U = k\sqrt{Q_1Q_2}$, and employing interference between said radiative far fields of the resonant fields of the first and second resonator, with an interference factor V_{rad}, to increase efficiency of energy transfer between the first and second resonator compared to efficiency of energy transfer between the first and second resonator in the absence of interference, a strong-interference factor being defined as $V = V_{\mathrm{rad}}\sqrt{(Q_1 / Q_{1,\mathrm{rad}})(Q_2 / Q_{2,\mathrm{rad}})}$.

The following are examples within the scope of this aspect.

The method has $Q_1 / Q_{1,\mathrm{rad}} \geq 0.05$ and $Q_2 / Q_{2,\mathrm{rad}} \geq 0.05$. During operation, a power generator is coupled to one of the first and second resonant structure and is configured to drive the resonator structure, to which it is coupled, at a driving frequency *f*, corresponding to a driving angular frequency $\omega = 2\pi f$, wherein the driving frequency is different from the resonance frequencies of the first and second resonator structures and is closer to a frequency corresponding to an odd normal mode of the system of the two resonator structures. During operation, a power load is coupled to the resonant structure to which the power generator is not coupled and is configured to receive from the resonator structure, to which it is coupled, a usable power. The resonance angular frequencies for the resonator structures are selected to optimize the energy-transfer efficiency by accounting for both the strong-coupling factor *U* and the strong-interference interference factor *V*.

Various examples may include any of the above features, alone or in combination. Other features, objects, and advantages of the disclosure will be apparent from the following detailed description.

BRIEF DESCRIPTION OF THE DRAWINGS

Fig. 1 shows a schematic of an example wireless energy transfer scheme.

Figs. 2(a)-(b) show the efficiency of power transmission η_{P} for (a) $U = 1$ and (b) $U = 3$, as a function of the frequency detuning D_o and for different values of the loading rate U_o.

Fig. 2(c) shows the optimal (for zero detuning and under conditions of impedance matching) efficiency for energy transfer $\eta_{\mathrm{E}*}$ and power transmission $\eta_{\mathrm{P}*}$, as a function of the coupling-to-loss figure-of-merit U.

Fig. 3 shows an example of a self-resonant conducting-wire coil.

Fig. 4 shows an example of a wireless energy transfer scheme featuring two self-resonant conducting-wire coils

Fig. 5 is a schematic of an experimental system demonstrating wireless energy transfer.

Fig. 6 shows a comparison between experimental and theoretical results for the coupling rate of the system shown schematically in Fig. 5.

Fig. 7 shows a comparison between experimental and theoretical results for the strong-coupling factor of the system shown schematically in Fig. 5.

Fig. 8 shows a comparison between experimental and theoretical results for the power-transmission efficiency of the system shown schematically in Fig. 5.

Fig. 9 shows an example of a capacitively loaded conducting-wire coil, and illustrates the surrounding field.

Fig. 10 shows an example wireless energy transfer scheme featuring two capacitively loaded conducting-wire coils, and illustrates the surrounding field.

Fig. 11 illustrates an example circuit model for wireless energy transfer.

Fig. 12 shows the efficiency, total (loaded) device Q, and source and device currents, voltages and radiated powers (normalized to 1Watt of output power to the load) as functions of the resonant frequency, for a particular choice of source and device loop dimensions, wp and N_s and different choices of N_d=1,2,3,4,5,6,10 (red, green, blue, magenta, yellow, cyan, black respectively).

Fig.13 shows the efficiency, total (loaded) device Q, and source and device currents, voltages and radiated powers (normalized to 1Watt of output power to the load) as functions of frequency and wp for a particular choice of source and device loop dimensions, and number of turns N_s and N_d.

Fig. 14 shows an example of an inductively-loaded conducting-wire coil.

Fig. 15 shows (a) an example of a resonant dielectric disk, and illustrates the surrounding field and (b) a wireless energy transfer scheme featuring two resonant dielectric disks, and illustrates the surrounding field.

Figs. 16(a)-(b) show the efficiency of power transmission η_P for (a) U = 1, V = 0.5 and (b) U = 3, V = 0.5, as a function of the frequency detuning Do and for different values of the loading rate U_o. (The dotted lines show, for comparison, the results when there is no interference, as shown in Fig.2(a)-(b).) and Figs. 16(c)-(d) show the optimal (for optimal detuning and under conditions of impedance matching) efficiency for energy transfer (only in (c)) and power transmission, as a function of the strong-coupling factor U and the strong-interference factor V.

Fig. 17 shows coupled-mode theory (CMT) results for (a) the coupling factor k and (b) the strong-coupling factor U as a function of the relative distance D/r between two identical capacitively-loaded conducting single-turn loops, for three different dimensions of the loops. Note that, for conducting material, copper ($\sigma = 5.998 \cdot 10^7\,\mathrm{S/m}$) was used.

Fig. 18 shows AT results for the interference factor V_{rad} as a function of the distance D (normalized to the wavelength λ) between two capacitively-loaded conducting loops.

Fig. 19 shows CMT results for the strong-coupling factor U and AT results for the interference factor V_{rad} and strong-interference factor V as a function of the resonant eigenfrequency of two identical capacitively-loaded conducting single-turn loops with $r = 30\mathrm{cm}$ and $a = 2\mathrm{cm}$, at a relative distance $D/r = 5$ between them. Note that, for conducting material, copper ($\sigma = 5.998 \cdot 10^7\,\mathrm{S/m}$) was used.

Fig. 20 shows the power-transmission efficiency as a function of the resonant eigenfrequency of two identical capacitively-loaded conducting single-turn loops. Results for two different loop dimensions are shown and for two relative distances between the identical loops. For each loops dimension and distance, four different cases are examined: without far-field interference (dotted), with far-field interference but no driving-frequency detuning (dashed) and with driving-frequency detuning to maximize either the efficiency (solid) or the ratio of efficiency over radiation (dash-dotted).

Fig. 21 shows the driving-frequency detunings required in the presence of far-field interference as a function of the resonant eigenfrequency of two identical capacitively-loaded conducting single-turn loops of Fig. 20 to maximize either the efficiency (solid) or the ratio of efficiency over radiation (dash-dotted).

Fig. 22(a) shows the resonant eigenfrequencies f_{U} and f_{η}, where the strong-coupling factor U and the power-transmission efficiency η peak respectively, as a function of the relative distance D/r between two identical loops with $r = 30\text{cm}$ and $a = 2\text{cm}$.

Fig. 22(b) illustrates the strong-coupling factor U and the strong-interference factor V as a curve in the $U-V$ plane, parametrized with the relative distance D/r between the two loops, for the cases with interference and eigenfrequency f_η (solid), with interference and eigenfrequency f_U (dashed), and without interference and eigenfrequency f_U (dotted).

Fig. 22(c) shows the efficiency enhancement ratio of the solid curve in Fig. 22(b) relative to the dashed and dotted curves in Fig. 22(b).

Fig. 23 shows the radiation efficiency as a function of the resonant eigenfrequency of two identical capacitively-loaded conducting single-turn loops. Results for two different loop dimensions are shown and for two relative distances between the identical loops. For each loops dimension and distance, four different cases are examined: without far-field interference (dotted), with far-field interference but no driving-frequency detuning (dashed) and with driving-frequency detuning to maximize either the efficiency (solid) or the ratio of efficiency over radiation (dash-dotted).

Fig. 24 shows CMT results for (a) the coupling factor k and (b) the strong-coupling factor U, for three different m values of subwavelength resonant modes of two same dielectric disks at distance $D/r = 5$ (and also a couple more distances for $m = 2$), when varying their ϵ in the range $250 \geq \epsilon \geq 35$. Note that disk-material loss-tangent $\tan\delta = 6{\cdot}10^{-6}\epsilon - 2{\cdot}10^{-4}$ was used. (c) Relative U error between CMT and numerical FEFD calculations of part (b).

Fig. 25 shows Antenna Theory (AT) results for (a) the normalized interference term $2\Lambda/\sqrt{\omega_1\omega_2}$ and (b) magnitude of the strong-interference factor $|V|$, as a function of frequency, for the exact same parameters as in Fig.24. (c) Relative V error between AT and numerical FEFD calculations of part (b).

Fig. 26 shows results for the overall power transmission as a function of frequency, for the same set of resonant modes and distances as in Figs.24 and 25, based on the predictions including interference (solid lines) and without interference, just from U (dotted lines).

Fig. 27 (a) shows the frequencies f_U and f_η, where the strong-coupling factor U and the power-transmission efficiency η are respectively maximized, as a function of the transfer distance between the $m = 2$ disks of Fig. 15. Fig. 27(b) shows the efficiencies achieved at the frequencies of (a) and, in inset, the enhancement ratio of the optimal (by definition) efficiency for f_η versus the achievable efficiency at f_U. Fig. 27(c) shows the D-parametrized path of the transmission efficiency for the frequency choices of (a) on the $U - V$ efficiency map.

Fig. 28 shows results for the radiation efficiency as a function of the transfer distance at resonant frequency f_U, when the operating frequency is detuned (solid line), when it is not (dashed line), and when there is no interference whatsoever (dotted line). In the inset, we show the corresponding radiation suppression factors.

Figs. 29(a)-(b) show schematics for frequency control mechanisms.

Figs. 30(a)-(c) illustrate a wireless energy transfer scheme using two dielectric disks in the presence of various extraneous objects.

DETAILED DESCRIPTION

1. Efficient energy-transfer by 'strongly coupled' resonances

Fig. 1 shows a schematic that generally describes one example of the invention, in which energy is transferred wirelessly between two resonant objects. Referring to Fig. 1, energy is transferred, over a distance D, between a resonant source object having a characteristic size r_1 and a resonant device object of characteristic size r_2. Both objects are resonant objects. The wireless non-radiative energy transfer is performed using the field (e.g. the electromagnetic field or acoustic field) of the system of two resonant objects.

The characteristic size of an object can be regarded as being equal to the radius of the smallest sphere which can fit around the entire object. The characteristic thickness of an object can be regarded as being, when placed on a flat surface in any arbitrary configuration, the smallest possible height of the highest point of the object above a flat surface. The characteristic width of an object can be regarded as being the radius of the smallest possible circle that the object can pass through while traveling in a straight line. For example, the characteristic width of a cylindrical object is the radius of the cylinder.

It is to be understood that while two resonant objects are shown in the example of Fig. 1, and in many of the examples below, other examples can feature three or more resonant objects. For example, in some examples, a single source object can transfer energy to multiple device objects. In some examples, energy can be transferred from a first resonant object to a second resonant object, and then from the second resonant object to a third resonant object, and so forth.

Initially, we present a theoretical framework for understanding non-radiative wireless energy transfer. Note however that it is to be understood that the scope of the invention is not bound by theory.

Different temporal schemes can be employed, depending on the application, to transfer energy between two resonant objects. Here we will consider two particularly simple but important schemes: a one-time finite-amount energy-transfer scheme and a continuous finite-rate energy-transfer (power) scheme.

1.1 Finite-amount energy-transfer efficiency

Let the source and device objects be 1, 2 respectively and their resonance eigemodes, which we will use for the energy exchange, have angular frequencies $\omega_{1,2}$, frequency-widths due to intrinsic (absorption, radiation etc.) losses $\Gamma_{1,2}$ and (generally) vector fields $\mathbf{F}_{1,2}(\mathbf{r})$, normalized to unity energy. Once the two resonant objects are brought in proximity, they can interact and an appropriate analytical framework for modeling this resonant interaction is that of the well-known coupled-mode theory (CMT). In this picture, the field of the system of the two resonant objects 1, 2 can be approximated by $\mathbf{F}(\mathbf{r},t)=a_1(t)\mathbf{F}_1(\mathbf{r})+a_2(t)\mathbf{F}_2(\mathbf{r})$, where $a_{1,2}(t)$ are the field amplitudes, with $|a_{1,2}(t)|^2$ equal to the energy stored inside the object 1, 2 respectively, due to the normalization. Then, using $e^{-i\omega t}$ time dependence, the field amplitudes can be shown to satisfy, to lowest order:

$$\begin{aligned}\frac{d}{dt}a_1(t)&=-i(\omega_1-i\Gamma_1)a_1(t)+i\kappa_{11}a_1(t)+i\kappa_{12}a_2(t)\\ \frac{d}{dt}a_2(t)&=-i(\omega_2-i\Gamma_2)a_2(t)+i\kappa_{21}a_1(t)+i\kappa_{22}a_2(t)\end{aligned} \tag{1}$$

where $\kappa_{11,22}$ are the shifts in each object's frequency due to the presence of the other, which are a second-order correction and can be absorbed into the eigenfrequencies by setting $\omega_{1,2} \to \omega_{1,2}+\kappa_{11,22}$, and $\kappa_{12,21}$ are the coupling coefficients, which from the reciprocity requirement of the system must satisfy $\kappa_{21}=\kappa_{12}\equiv\kappa$.

The normal modes of the combined system are found, by substituting $[a_1(t), a_2(t)] = [A_1, A_2]e^{-i\bar{\omega}t}$, to have complex frequencies

$$\bar{\omega}_{\pm}=\frac{\omega_1+\omega_2}{2}-i\frac{\Gamma_1+\Gamma_2}{2}\pm\sqrt{\left(\frac{\omega_1-\omega_2}{2}-i\frac{\Gamma_1-\Gamma_2}{2}\right)^2+\kappa^2} \tag{2}$$

whose splitting we denote as $\delta_E \equiv \bar{\omega}_+ - \bar{\omega}_-$. Note that, at exact resonance $\omega_1 = \omega_2$ and for $\Gamma_1 = \Gamma_2$, we get $\delta_E = 2\kappa$.

Assume now that at time $t = 0$ the source object 1 has finite energy $|a_1(0)|^2$, while the device object has $|a_2(0)|^2 = 0$. Since the objects are coupled, energy will be transferred from 1 to 2. With these initial conditions, Eqs.(1) can be solved, predicting the evolution of the device field-amplitude to be

$$\frac{a_2(t)}{|a_1(0)|}=\frac{2\kappa}{\delta_E}\sin\left(\frac{\delta_E t}{2}\right)e^{-\frac{\Gamma_1+\Gamma_2}{2}t}. \tag{3}$$

The energy-transfer efficiency will be $\eta_E \equiv |a_2(t)|^2/|a_1(0)|^2$. Note that, at exact resonance $\omega_1 = \omega_2$ and in the special case $\Gamma_1 = \Gamma_2 \equiv \Gamma_o$, Eq.(3) can be written as

$$\frac{a_2(T)}{|a_1(0)|}=\sin(UT)\cdot e^{-T} \tag{4}$$

where $T \equiv \Gamma_o t$ and $U = \kappa/\Gamma_o$.

In some examples, the system designer can adjust the duration of the coupling t at will. In some examples, the duration t can be adjusted to maximize the device energy (and thus efficiency η_E). Then, in the special case $\Gamma_1 = \Gamma_2 = \Gamma_o$, it can be inferred from Eq.(4) that η_E is maximized for

$$T_* = \frac{\tan^{-1} U}{U} \tag{5}$$

resulting in an optimal energy-transfer efficiency

$$\eta_{E*} \equiv \eta_E(T_*) = \frac{U^2}{1+U^2} \exp\left(-\frac{2\tan^{-1} U}{U}\right) \tag{6}$$

which is only a function of the coupling-to-loss ratio $U = \kappa/\Gamma_o$ and tends to unity when $U \gg 1$, as depicted in Fig.2(c). In general, also for $\Gamma_1 \neq \Gamma_2$, the energy transfer is nearly perfect, when the coupling rate is much faster than all loss rates ($\kappa/\Gamma_{1,2} \gg 1$).

In a real wireless energy-transfer system, the source object can be connected to a power generator (not shown in Fig.1), and the device object can be connected to a power consuming load (e.g. a resistor, a battery, an actual device, not shown in Fig.1). The generator will supply the energy to the source object, the energy will be transferred wirelessly and non-radiatively from the source object to the device object, and the load will consume the energy from the device object. To incorporate such supply and consumption mechanisms into this temporal scheme, in some examples, one can imagine that the generator is very briefly but very strongly coupled to the source at time $t = 0$ to almost instantaneously provide the energy, and the load is similarly very briefly but very strongly coupled to the device at the optimal time $t = t_*$ to almost instantaneously drain the energy. For a constant powering mechanism, at time $t = t_*$ also the generator can again be coupled to the source to feed a new amount of energy, and this process can be repeated periodically with a period t_*.

1.2 Finite-rate energy-transfer (power-transmission) efficiency

Let the generator be continuously supplying energy to the source object 1 at a rate κ_1 and the load continuously draining energy from the device object 2 at a rate κ_2. Field amplitudes $s_{\pm1,2}(t)$ are then defined, so that $|s_{\pm1,2}(t)|^2$ is equal to the power ingoing to (for the + sign) or outgoing from (for the - sign) the object 1, 2 respectively, and the CMT equations are modified to

$$\begin{aligned}
&\frac{d}{dt}a_1(t) = -i(\omega_1 - i\Gamma_1)a_1(t) + i\kappa_{11}a_1(t) + i\kappa_{12}a_2(t) - \kappa_1 a_1(t) + \sqrt{2\kappa_1}s_{+1}(t) \\
&\frac{d}{dt}a_2(t) = -i(\omega_2 - i\Gamma_2)a_2(t) + i\kappa_{21}a_1(t) + i\kappa_{22}a_2(t) - \kappa_2 a_2(t) \\
&s_{-1}(t) = \sqrt{2\kappa_1}a_1(t) - s_{+1}(t) \\
&s_{-2}(t) = \sqrt{2\kappa_2}a_2(t)
\end{aligned} \tag{7}$$

where again we can set $\omega_{1,2} \rightarrow \omega_{1,2} + \kappa_{11,22}$ and $\kappa_{21} = \kappa_{12} \equiv \kappa$.

Assume now that the excitation is at a fixed frequency ω, namely has the form $s_{+1}(t) = S_{+1}e^{-i\omega t}$. Then the response of the linear system will be at the same frequency, namely $a_{1,2}(t) = A_{1,2}e^{-i\omega t}$ and $s_{-1,2}(t) = S_{-1,2}e^{-i\omega t}$. By substituting these into Eqs.(7), using $\delta_{1,2} \equiv \omega - \omega_{1,2}$, and solving the system, we find the field-amplitude transmitted to the load (S_{21} scattering-matrix element)

$$\begin{aligned}
S_{21} \equiv \frac{S_{-2}}{S_{+1}} &= \frac{2i\kappa\sqrt{\kappa_1\kappa_2}}{(\Gamma_1 + \kappa_1 - i\delta_1)(\Gamma_2 + \kappa_2 - i\delta_2) + \kappa^2} \\
&= \frac{2iU\sqrt{U_1U_2}}{(1 + U_1 - iD_1)(1 + U_2 - iD_2) + U^2}
\end{aligned} \tag{8}$$

and the field-amplitude reflected to the generator (S_{11} scattering-matrix element)

$$\begin{aligned}
S_{11} \equiv \frac{S_{-1}}{S_{+1}} &= \frac{(\Gamma_1 - \kappa_1 - i\delta_1)(\Gamma_2 + \kappa_2 - i\delta_2) + \kappa^2}{(\Gamma_1 + \kappa_1 - i\delta_1)(\Gamma_2 + \kappa_2 - i\delta_2) + \kappa^2} \\
&= \frac{(1 - U_1 - iD_1)(1 + U_2 - iD_2) + U^2}{(1 + U_1 - iD_1)(1 + U_2 - iD_2) + U^2}
\end{aligned} \tag{9}$$

where $D_{1,2} \equiv \delta_{1,2}/\Gamma_{1,2}$, $U_{1,2} \equiv \kappa_{1,2}/\Gamma_{1,2}$ and $U \equiv \kappa/\sqrt{\Gamma_1\Gamma_2}$. Similarly, the scattering-matrix elementsS_{12}, S_{22} are given by interchanging 1 ↔ 2 in Eqs.(8),(9) and, as expected from reciprocity, $S_{21} = S_{12}$. The coefficients for power transmission (efficiency) and reflection and loss are respectively $\eta_P \equiv |S_{21}|^2 = |S_{-2}|^2/|S_{+1}|^2$ and $|S_{11}|^2 = |S_{-1}|^2/|S_{+1}|^2$ and $1 - |S_{21}|^2 - |S_{11}|^2 = (2\Gamma_1|\mathrm{A}_1|^2 + 2\Gamma_2|\mathrm{A}_2|^2)/|S_{+1}|^2$.

In practice, in some implementations, the parameters $D_{1,2}$, $U_{1,2}$ can be designed (engineered), since one can adjust the resonant frequencies $\omega_{1,2}$ (compared to the desired operating frequency ω) and the generator/load supply/drain rates $\kappa_{1,2}$. Their choice can target the optimization of some system performance-characteristic of interest:

In some examples, a goal can be to maximize the power transmission (efficiency) $\eta_P \equiv |S_{21}|^2$ of the system, so one would require

$$\eta_P'\left(D_{1,2}\right) = \eta_P'\left(U_{1,2}\right) = 0 \tag{10}$$

Since S_{21} (from Eq.(8)) is symmetric upon interchanging $1 \leftrightarrow 2$, the optimal values for $D_{1,2}$ (determined by Eqs.(10)) will be equal, namely $D_1 = D_2 \equiv D_o$, and similarly $U_1 = U_2 \equiv U_o$. Then,

$$S_{21} = \frac{2iUU_o}{\left(1+U_o - iD_o\right)^2 + U^2} \tag{11}$$

and from the condition $\eta_P'(D_o) = 0$ we get that, for fixed values of U and U_o, the efficiency can be maximized for the following values of the symmetric detuning

$$D_o = \begin{array}{ll} \pm\sqrt{U^2 - \left(1+U_o\right)^2}, & \text{if} \quad U > 1+U_o \\ 0, & \text{if} \quad U \le 1+U_o \end{array}, \tag{12}$$

which, in the case $U > 1 + U_o$, can be rewritten for the two frequencies at which the efficiency peaks as

$$\omega_{\pm} = \frac{\omega_1\Gamma_2 + \omega_2\Gamma_1}{\Gamma_1 + \Gamma_2} \pm \frac{2\sqrt{\Gamma_1\Gamma_2}}{\Gamma_1 + \Gamma_2}\sqrt{\kappa^2 - \left(\Gamma_1 + \kappa_1\right)\left(\Gamma_2 + \kappa_2\right)}, \tag{13}$$

whose splitting we denote as $\delta_P \equiv \bar{\omega}_+ - \bar{\omega}_-$. Note that, at exact resonance $\omega_1 = \omega_2$, and for $\Gamma_1 = \Gamma_2 \equiv \Gamma_o$ and $\kappa_1 = \kappa_2 \equiv \kappa_o$, we get $\delta_P = 2\sqrt{\kappa^2 - (\Gamma_o + \kappa_o)^2} < \delta_E$, namely the transmission-peak splitting is smaller than the normal-mode splitting. Then, by substituting D_o into η_P from Eq.(12), from the condition $\eta_P'(U_o) = 0$ we get that, for fixed value of U, the efficiency can be maximized for

$$U_{o*} = \sqrt{1+U^2} \quad \overset{\text{Eq.(12)}}{\Rightarrow} \quad D_{o*} = 0 \tag{14}$$

which is known as ‘critical coupling’ condition, whereas for $U_o < U_{o*}$ the system is called ‘undercoupled’ and for $U_o > U_{o*}$ it is called ‘overcoupled’. The dependence of the efficiency on the frequency detuning D_o for different values of U_o (including the 'critical-coupling' condition) are shown in Fig. 2(a,b). The overall optimal power efficiency using Eqs.(14) is

$$\eta_{P*} \equiv \eta_P\left(D_{o*}, U_{o*}\right) = \frac{U_{o*}-1}{U_{o*}+1} = \left(\frac{U}{1+\sqrt{1+U^2}}\right)^2, \tag{15}$$

which is again only a function of the coupling-to-loss ratio $U = \kappa/\sqrt{\Gamma_1\Gamma_2}$ and tends to unity when $U \gg 1$, as depicted in Fig. 2(c).

In some examples, a goal can be to minimize the power reflection at the side of the generator $|S_{11}|^2$ and the load $|S_{22}|^2$, so one would then need

$$S_{11,22} = 0 \Rightarrow \left(1 \mp U_1 - iD_1\right)\left(1 \pm U_2 - iD_2\right) + U^2 = 0, \tag{16}$$

The equations above present 'impedance matching' conditions. Again, the set of these conditions is symmetric upon interchanging $1 \leftrightarrow 2$, so, by substituting $D_1 = D_2 \equiv D_o$ and $U_1 = U_2 \equiv U_o$ into Eqs.(16), we get

$$\left(1 - iD_o\right)^2 - U_o^2 + U^2 = 0, \tag{17}$$

from which we easily find that the values of D_o and U_o that cancel all reflections are again exactly those in Eqs.(14).

It can be seen that, for this particular problem, the two goals and their associated sets of conditions (Eqs.(10) and Eqs.(16)) result in the same optimized values of the intra-source and intra-device parameters $D_{1,2}$, $U_{1,2}$. Note that for a lossless system this would be an immediate consequence of power conservation (Hermiticity of the scattering matrix), but this is not apparent for a lossy system.

Accordingly, for any temporal energy-transfer scheme, once the parameters specific only to the source or to the device (such as their resonant frequencies and their excitation or loading rates respectively) have been optimally designed, the efficiency monotonically increases with the ratio of the source-device coupling-rate to their loss rates. Using the definition of a resonance quality factor $Q = \omega/2\Gamma$ and defining by analogy the coupling factor $k \equiv 1/Q_\kappa \equiv 2\kappa/\sqrt{\omega_1\omega_2}$, it is therefore exactly this ratio

$$U = \frac{\kappa}{\sqrt{\Gamma_1 \Gamma_2}} = k\sqrt{Q_1 Q_2} \tag{18}$$

that has been set as a figure-of-merit for any system under consideration for wireless energy-transfer, along with the distance over which this ratio can be achieved (clearly, U will be a decreasing function of distance). The desired optimal regime $U > 1$ is called 'strong-coupling' regime and it is a necessary and sufficient condition for efficient energy-transfer. In particular, for $U > 1$ we get, from Eq.(15), $\eta_{P*} > 17\%$, large enough for practical applications. The figure-of-merit U is called the strong-coupling factor. We will further show how to design systems with a large strong-coupling factor.

To achieve a large strong-coupling factor U, in some examples, the energy-transfer application preferably uses resonant modes of high quality factors Q, corresponding to low (i.e. slow) intrinsic-loss rates Γ. This condition can be satisfied by designing resonant modes where all loss mechanisms, typically radiation and absorption, are sufficiently suppressed.

This suggests that the coupling be implemented using, not the lossy radiative far-field, which should rather be suppressed, but the evanescent (non-lossy) stationary near-field. To implement an energy-transfer scheme, usually more appropriate are finite objects, namely ones that are topologically surrounded everywhere by air, into where the near field extends to achieve the coupling. Objects of finite extent do not generally support electromagnetic states that are exponentially decaying in *all* directions in air away from the objects, since Maxwell's Equations in free space imply that $\mathbf{k}^2 = \omega^2/c^2$, where $\mathbf{k}$ is the wave vector, ω the angular frequency, and c the speed of light, because of which one can show that such finite objects cannot support states of infinite Q, rather there always is some amount of radiation. However, very long-lived (so-called "high-Q") states can be found, whose tails display the needed exponential or exponential-like decay away from the resonant object over long enough distances before they turn oscillatory (radiative). The limiting surface, where this change in the field behavior happens, is called the "radiation caustic", and, for the wireless energy-transfer scheme to be based on the near field rather than the far/radiation field, the distance between the coupled objects must be such that one lies within the radiation caustic of the other. One typical way of achieving a high radiation-Q (Q_{rad}) is to design subwavelength resonant objects. When the size of an object is much smaller than the wavelength of radiation in free space, its electromagnetic field couples to radiation very weakly. Since the extent of the near-field into the area surrounding a finite-sized resonant object is set typically by the wavelength, in some examples, resonant objects of subwavelength size have significantly longer evanescent field-tails. In other words, the radiation caustic is pushed far away from the object, so the electromagnetic mode enters the radiative regime only with a small amplitude.

Moreover, most realistic materials exhibit some nonzero amount of absorption, which can be frequency dependent, and thus cannot support states of infinite Q, rather there always is some amount of absorption. However, very long-lived ("high-Q") states can be found, where electromagnetic modal energy is only weakly dissipated. Some typical ways of achieving a high absorption-Q (Q_{abs}) is to use materials which exhibit very small absorption at the resonant frequency and/or to shape the field to be localized more inside the least lossy materials.

Furthermore, to achieve a large strong-coupling factor U, in some examples, the energy-transfer application preferably uses systems that achieve a high coupling factor k, corresponding to strong (i.e. fast) coupling rate κ, over distances larger than the characteristic sizes of the objects.

Since finite-sized subwavelength resonant objects can often be accompanied with a high Q, as was discussed above and will be seen in examples later on, such an object will typically be the appropriate choice for the possibly-mobile resonant device-object. In these cases, the electromagnetic field is, in some examples, of quasi-static nature and the distance, up to which sufficient coupling can be achieved, is dictated by the decay-law of this quasi-static field.

Note, though, that in some examples, the resonant source-object will be immobile and thus less restricted in its allowed geometry and size. It can be therefore chosen large enough that the near-field extent is not limited by the wavelength, and can thus have nearly infinite radiation-Q. Some objects of nearly infinite extent, such as dielectric waveguides, can support guided modes, whose evanescent tails are decaying exponentially in the direction away from the object, slowly if tuned close to cutoff, therefore a good coupling can also be achieved over distances quite a few times larger than a characteristic size of the source- and/or device-object.

2 'Strongly-coupled' resonances at mid-range distances for realistic systems

In the following, examples of systems suitable for energy transfer of the type described above are described. We will demonstrate how to compute the CMT parameters $\omega_{1,2}$, $Q_{1,2}$ and k described above and how to choose or design these parameters for particular examples in order to produce a desirable figure-of-merit $U = \kappa/\sqrt{\Gamma_1\Gamma_2} = k\sqrt{Q_1 Q_2}$ at a desired distance D. In some examples, this figure-of-merit is maximized when $\omega_{1,2}$ are tuned close to a particular angular frequency ω_U.

2.1 Self-resonant conducting coils

In some examples, one or more of the resonant objects are self-resonant conducting coils. Referring to Fig. 3, a conducting wire of length l and cross-sectional radius a is wound into a helical coil of radius r and height h (namely with $N = \sqrt{l^2 - h^2}\,/2\pi r$ number of turns), surrounded by air. As described below, the wire has distributed inductance and distributed capacitance, and therefore it supports a resonant mode of angular frequency ω. The nature of the resonance lies in the periodic exchange of energy from the electric field within the capacitance of the coil, due to the charge distribution $\rho(\mathbf{x})$ across it, to the magnetic field in free space, due to the current distribution $\mathbf{j}(\mathbf{x})$ in the wire. In particular, the charge conservation equation $\nabla \cdot \mathbf{j} = i\omega\rho$ implies that: (i) this periodic exchange is accompanied by a $\pi/2$ phase-shift between the current and the charge density profiles, namely the energy W contained in the coil is at certain points in time completely due to the current and at other points in time completely due to the charge, and (ii) if $\rho_l(x)$ and $I(x)$ are respectively the linear charge and current densities in the wire, where x runs along the wire, $q_o = \frac{1}{2}\int dx\left|\rho_l(x)\right|$ is the maximum amount of positive charge accumulated in one side of the coil (where an equal amount of negative charge always also accumulates in the other side to make the system neutral) and $I_o = \max\{\left|I(x)\right|\}$ is the maximum positive value of the linear current distribution, then

$I_o = \omega q_o$. Then, one can define an effective total inductance L and an effective total capacitance C of the coil through the amount of energy W inside its resonant mode:

$$W \equiv \frac{1}{2} I_o^2 L \Rightarrow L = \frac{\mu_o}{4\pi I_o^2} \iint d\mathbf{x} d\mathbf{x}' \frac{\mathbf{j}(\mathbf{x}) \cdot \mathbf{j}(\mathbf{x}')}{|\mathbf{x} - \mathbf{x}'|}, \tag{19}$$

$$W \equiv \frac{1}{2} q_o^2 \frac{1}{C} \Rightarrow \frac{1}{C} = \frac{1}{4\pi\varepsilon_o q_o^2} \iint d\mathbf{x} d\mathbf{x}' \frac{\rho(\mathbf{x}) \rho(\mathbf{x}')}{|\mathbf{x} - \mathbf{x}'|}, \tag{20}$$

where μ_o and ε_o are the magnetic permeability and electric permittivity of free space.

With these definitions, the resonant angular frequency and the effective impedance can be given by the formulas $\omega = 1/\sqrt{LC}$ and $Z = \sqrt{L/C}$ respectively.

Losses in this resonant system consist of ohmic (material absorption) loss inside the wire and radiative loss into free space. One can again define a total absorption resistance R_{abs} from the amount of power absorbed inside the wire and a total radiation resistance R_{rad} from the amount of power radiated due to electric- and magnetic-dipole radiation:

$$P_{abs} \equiv \tfrac{1}{2} I_o^2 R_{abs} \Rightarrow R_{abs} \approx \zeta_c \frac{l}{2\pi a} \cdot \frac{I_{rms}^2}{I_o^2} \tag{21}$$

$$P_{rad} \equiv \tfrac{1}{2} I_o^2 R_{rad} \Rightarrow R_{rad} \approx \frac{\zeta_o}{6\pi} \left[\left(\frac{\omega |\mathbf{p}|}{c} \right)^2 + \left(\frac{\omega \sqrt{|\mathbf{m}|}}{c} \right)^4 \right], \tag{22}$$

where $c = 1/\sqrt{\mu_o \varepsilon_o}$ and $\zeta_o = \sqrt{\mu_o / \varepsilon_o}$ are the light velocity and light impedance in free space, the impedance ζ_c is $\zeta_c = 1/\sigma\delta = \sqrt{\mu_o \omega / 2\sigma}$ with σ the conductivity of the conductor and δ the skin depth at the frequency ω, $I_{rms}^2 = \frac{1}{l} \int dx |I(x)|^2$, $\mathbf{p} = \int dx\, \mathbf{r} \rho_l(x)$ is the electric-dipole moment of the coil and $\mathbf{m} = \frac{1}{2} \int dx\, \mathbf{r} \times \mathbf{j}(x)$ is the magnetic-dipole moment of the coil. For the radiation resistance formula Eq.(22), the assumption of operation in the quasi-static regime ($h, r \ll \lambda = 2\pi c / \omega$) has been used, which is the desired regime of a subwavelength resonance. With these definitions, the absorption and radiation quality factors of the resonance are given by $Q_{abs} = Z / R_{abs}$ and $Q_{rad} = Z / R_{rad}$ respectively.

From Eq.(19)-(22) it follows that to determine the resonance parameters one simply needs to know the current distribution $\mathbf{j}$ in the resonant coil. Solving Maxwell's equations to rigorously find the current distribution of the resonant electromagnetic eigenmode of a conducting-wire coil is more involved than, for example, of a standard LC circuit, and we can find no exact solutions in the literature for coils of finite length, making an exact solution difficult. One could in principle write down an elaborate transmission-line-like model, and solve it by brute force. We instead present a model that is (as described below) in good agreement (~5%) with experiment. Observing that the finite extent of the conductor forming each coil imposes the boundary condition that the current has to be zero at the ends of the coil, since no current can leave the wire, we assume that the resonant mode of each coil is well approximated by a sinusoidal current profile along the length of the conducting wire. We shall be interested in the lowest mode, so if we denote by x the coordinate along the conductor, such that it runs from $-l/2$ to $+l/2$, then the current amplitude profile would have the form $I(x) = I_o \cos(\pi x / l)$, where we have assumed that the current does not vary significantly along the wire circumference for a particular x, a valid assumption provided $a \ll r$. It immediately follows from the continuity equation for charge that the linear charge density profile should be of the form $\rho_l(x) = \rho_o \sin(\pi x / l)$, and thus $q_o = \int_0^{l/2} dx \rho_o \left|\sin(\pi x / l)\right| = \rho_o l / \pi$. Using these sinusoidal profiles we find the so-called "self-inductance" L_s and "self-capacitance" C_s of the coil by computing numerically the integrals Eq.(19) and (20); the associated frequency and effective impedance are ω_s and Z_s respectively. The "self-resistances" R_s are given analytically by Eq.(21) and (22) using $I_{rms}^2 = \frac{1}{l}\int_{-l/2}^{l/2} dx \left|I_o \cos(\pi x / l)\right|^2 = \frac{1}{2} I_o^2$, $|\mathbf{p}| = q_o \sqrt{\left(\frac{2}{\pi} h\right)^2 + \left(\frac{4N\cos(\pi N)}{(4N^2-1)\pi} r\right)^2}$ and

$|\mathbf{m}| = I_o \sqrt{\left(\frac{2}{\pi} N\pi r^2\right)^2 + \left(\frac{\cos(\pi N)(12N^2-1)-\sin(\pi N)\pi N(4N^2-1)}{(16N^4-8N^2+1)\pi} hr\right)^2}$, and therefore the associated Q_s factors can be calculated.

The results for two examples of resonant coils with subwavelength modes of $\lambda_s / r \geq 70$ (i.e. those highly suitable for near-field coupling and well within the quasi-static limit) are presented in Table 1. Numerical results are shown for the wavelength and absorption, radiation and total loss rates, for the two different cases of subwavelength-coil resonant modes. Note that, for conducting material, copper (σ=5.998•10^-7 S/m) was used. It can be seen that expected quality factors at microwave frequencies are $Q_{s,abs} \geq 1000$ and $Q_{s,rad} \geq 5000$.

Table 1

single coil	λ_s / r	f (MHz)	$Q_{s,rad}$	$Q_{s,abs}$	Q_s
r=30cm, h=20cm, a=1cm, N=4	74.7	13.39	4164	8170	2758
r=10cm, h=3cm, a=2mm, N=6	140	21.38	43919	3968	3639

Referring to Fig. 4, in some examples, energy is transferred between two self-resonant conducting-wire coils. The electric and magnetic fields are used to couple the different resonant conducting-wire coils at a distance D between their centers. Usually, the electric coupling highly dominates over the magnetic coupling in the system under consideration for coils with $h \gg 2r$ and, oppositely, the magnetic coupling highly dominates over the electric coupling for coils with $h \ll 2r$. Defining the charge and current distributions of two coils 1,2 respectively as $\rho_{1,2}(\mathbf{x})$ and $\mathbf{j}_{1,2}(\mathbf{x})$, total charges and peak currents respectively as $q_{1,2}$ and $I_{1,2}$, and capacitances and inductances respectively as $C_{1,2}$ and $L_{1,2}$, which are the analogs of $\rho(\mathbf{x})$, $\mathbf{j}(\mathbf{x})$, q_o, I_o, C and L for the single-coil case and are therefore well defined, we can *define* their mutual capacitance and inductance through the total energy:

$$W \equiv W_1 + W_2 + \frac{1}{2}\left(q_1^* q_2 + q_2^* q_1\right) / M_C + \frac{1}{2}\left(I_1^* I_2 + I_2^* I_1\right) M_L$$

$$\Rightarrow 1/M_C = \frac{1}{4\pi\varepsilon_o q_1 q_2} \iint d\mathbf{x} d\mathbf{x}' \frac{\rho_1(\mathbf{x})\rho_2(\mathbf{x}')}{|\mathbf{x}-\mathbf{x}'|} u, \quad M_L = \frac{\mu_o}{4\pi I_1 I_2} \iint d\mathbf{x} d\mathbf{x}' \frac{\mathbf{j}_1(\mathbf{x}) \cdot \mathbf{j}_2(\mathbf{x}')}{|\mathbf{x}-\mathbf{x}'|} u, \ (23)$$

where $W_1 = \frac{1}{2} q_1^2 / C_1 = \frac{1}{2} I_1^2 L_1$, $W_2 = \frac{1}{2} q_2^2 / C_2 = \frac{1}{2} I_2^2 L_2$ and the retardation factor of $u = \exp\left(i\omega|\mathbf{x}-\mathbf{x}'|/c\right)$ inside the integral can been ignored in the quasi-static regime $D \ll \lambda$ of interest, where each coil is within the near field of the other. With this definition, the coupling factor is given by $k = \sqrt{C_1 C_2} / M_C + M_L / \sqrt{L_1 L_2}$.

Therefore, to calculate the coupling rate between two self-resonant coils, again the current profiles are needed and, by using again the assumed sinusoidal current profiles, we compute numerically from Eq.(23) the mutual capacitance $M_{C,s}$ and inductance $M_{L,s}$ between two self-resonant coils at a distance D between their centers, and thus $k = 1/Q_\kappa$ is also determined.

Table 2

pair of coils	D/r	Q	$Q_\kappa = 1/k$	U
r=30cm, h=20cm, a=1cm, N=4, $\lambda/r \approx 75$, $Q_s^{abs} \approx 8170$, $Q_s^{rad} \approx 4164$	3	2758	38.9	70.9
	5	2758	139.4	19.8
	7	2758	333.0	8.3
	10	2758	818.9	3.4
r=10cm, h=3cm, a=2mm, N=6, $\lambda/r \approx 140$, $Q_s^{abs} \approx 3968$, $Q_s^{rad} \approx 43919$	3	3639	61.4	59.3
	5	3639	232.5	15.7
	7	3639	587.5	6.2
	10	3639	1580	2.3

Referring to Table 2, relevant parameters are shown for exemplary examples featuring pairs or identical self resonant coils. Numerical results are presented for the average wavelength and loss rates of the two normal modes (individual values not shown), and also the coupling rate and figure-of-merit as a function of the coupling distance *D*, for the two cases of modes presented in Table1. It can be seen that for medium distances $D/r = 10-3$ the expected coupling-to-loss ratios are in the range $U \sim 2-70$.

2.1.1 Experimental Results

An experimental realization of an example of the above described system for wireless energy transfer consists of two self-resonant coils of the type described above, one of which (the source coil) is coupled inductively to an oscillating circuit, and the second (the device coil) is coupled inductively to a resistive load, as shown schematically in Fig. 5. Referring to Fig. 5, A is a single copper loop of radius 25cm that is part of the driving circuit, which outputs a sine wave with frequency 9.9MHz. s and d are respectively the source and device coils referred to in the text. B is a loop of wire attached to the load ("light-bulb"). The various κ's represent direct couplings between the objects. The angle between coil d and the loop A is adjusted so that their direct coupling is zero, while coils s and d are aligned coaxially. The direct coupling between B and A and between B and s is negligible.

The parameters for the two identical helical coils built for the experimental validation of the power transfer scheme were $h = 20$ cm, $a = 3$ mm, $r = 30$ cm and $N = 5.25$. Both coils are made of copper. Due to imperfections in the construction, the spacing between loops of the helix is not uniform, and we have encapsulated the uncertainty about their uniformity by attributing a 10% (2 cm) uncertainty to h. The expected resonant frequency given these dimensions is $f_0 = 10.56 \pm 0.3$ MHz, which is about 5% off from the measured resonance at around 9.90 MHz.

The theoretical Q for the loops is estimated to be ~ 2500 (assuming perfect copper of resistivity $\rho = 1/\sigma = 1.7 \times 10^{-8}\,\Omega\,\text{m}$) but the measured value is 950 ± 50. We believe the discrepancy is mostly due to the effect of the layer of poorly conducting copper oxide on the surface of the copper wire, to which the current is confined by the short skin depth ($\sim 20\mu$m) at this frequency. We have therefore used the experimentally observed Q (and $\Gamma_1 = \Gamma_2 = \Gamma = \omega/(2Q)$ derived from it) in all subsequent computations.

The coupling coefficient κ can be found experimentally by placing the two self-resonant coils (fine-tuned, by slightly adjusting h, to the same resonant frequency when isolated) a distance D apart and measuring the splitting in the frequencies of the two resonant modes in the transmission spectrum. According to Eq.(13) derived by coupled-mode theory, the splitting in the transmission spectrum should be $\delta_{\mathrm{P}} = 2\sqrt{\kappa^2 - \Gamma^2}$, when $\kappa_{\mathrm{A,B}}$ are kept very small by keeping A and B at a relatively large distance. The comparison between experimental and theoretical results as a function of distance when the two the coils are aligned coaxially is shown in Fig. 6.

Fig. 7 shows a comparison of experimental and theoretical values for the strong-coupling factor $U = \kappa / \Gamma$ as a function of the separation between the two coils. The theory values are obtained by using the theoretically obtained κ and the experimentally measured Γ. The shaded area represents the spread in the theoretical U due to the $\sim 5\%$ uncertainty in Q. As noted above, the maximum theoretical efficiency depends only on the parameter U, which is plotted as a function of distance in Fig. 7. U is greater than 1 even for $D = 2.4\,\mathrm{m}$ (eight times the radius of the coils), thus the sytem is in the strongly-coupled regime throughout the entire range of distances probed.

The power-generator circuit was a standard Colpitts oscillator coupled inductively to the source coil by means of a single loop of copper wire 25cm in radius (see Fig. 5). The load consisted of a previously calibrated light-bulb, and was attached to its own loop of insulated wire, which was in turn placed in proximity of the device coil and inductively coupled to it. Thus, by varying the distance between the light-bulb and the device coil, the parameter $U_B = \kappa_{\mathrm{B}} / \Gamma$ was adjusted so that it matched its optimal value, given theoretically by Eq.(14) as $U_{\mathrm{B}*} = \sqrt{1 + U^2}$. Because of its inductive nature, the loop connected to the light-bulb added a small reactive component to κ_{B} which was compensated for by slightly retuning the coil. The work extracted was determined by adjusting the power going into the Colpitts oscillator until the light-bulb at the load was at its full nominal brightness.

In order to isolate the efficiency of the transfer taking place specifically between the source coil and the load, we measured the current at the mid-point of each of the self-resonant coils with a current-probe (which was not found to lower the Q of the coils noticeably.) This gave a measurement of the current parameters I_1 and I_2 defined above. The power dissipated in each coil was then computed from $P_{1,2} = \Gamma L \, | I_{1,2} |^2$, and the efficiency was directly obtained from $\eta = P_B / (P_1 + P_2 + P_B)$. To ensure that the experimental setup was well described by a two-object coupled-mode theory model, we positioned the device coil such that its direct coupling to the copper loop attached to the Colpitts oscillator was zero. The experimental results are shown in Fig. 8, along with the theoretical prediction for maximum efficiency, given by Eq.(15).

Using this example, we were able to transmit significant amounts of power using this setup from the source coil to the device coil, fully lighting up a 60W light-bulb from distances more than 2m away, for example. As an additional test, we also measured the total power going into the driving circuit. The efficiency of the wireless power-transmission itself was hard to estimate in this way, however, as the efficiency of the Colpitts oscillator itself is not precisely known, although it is expected to be far from 100%. Nevertheless, this gave an overly conservative lower bound on the efficiency. When transmitting 60W to the load over a distance of 2m, for example, the power flowing into the driving circuit was 400W. This yields an overall wall-to-load efficiency of $\sim 15\%$, which is reasonable given the expected $\sim 40\%$ efficiency for the wireless power transmission at that distance and the low efficiency of the driving circuit.

From the theoretical treatment above, we see that in typical examples it is important that the coils be on resonance for the power transmission to be practical. We found experimentally that the power transmitted to the load dropped sharply as one of the coils was detuned from resonance. For a fractional detuning $\Delta f/f_0$ of a few times the inverse loaded Q, the induced current in the device coil was indistinguishable from noise.

The power transmission was not found to be visibly affected as humans and various everyday objects, such as metallic and wooden furniture, as well as electronic devices large and small, were placed between the two coils, even when they drastically obstructed the line of sight between source and device. External objects were found to have an effect only when they were closer than 10cm from either one of the coils. While some materials (such as aluminum foil, styrofoam and humans) mostly just shifted the resonant frequency, which could in principle be easily corrected with a feedback circuit of the type described earlier, others (cardboard, wood, and PVC) lowered Q when placed closer than a few centimeters from the coil, thereby lowering the efficiency of the transfer.

This method of power transmission is believed safe for humans. When transmitting 60W (more than enough to power a laptop computer) across 2m, we estimated that the magnitude of the magnetic field generated is much weaker than the Earth's magnetic field for all distances except for less than about 1cm away from the wires in the coil, an indication of the safety of the scheme even after long-term use. The power radiated for these parameters was ~ 5 W, which is roughly an order of magnitude higher than cell phones but could be drastically reduced, as discussed below.

Although the two coils are currently of identical dimensions, it is possible to make the device coil small enough to fit into portable devices without decreasing the efficiency. One could, for instance, maintain the product of the characteristic sizes of the source and device coils constant.

These experiments demonstrated experimentally a system for power transmission over medium range distances, and found that the experimental results match theory well in multiple independent and mutually consistent tests.

The efficiency of the scheme and the distances covered can be appreciably improved by silver-plating the coils, which should increase their Q, or by working with more elaborate geometries for the resonant objects. Nevertheless, the performance characteristics of the system presented here are already at levels where they could be useful in practical applications.

2.2 Capacitively-loaded conducting loops or coils

In some examples, one or more of the resonant objects are capacitively-loaded conducting loops or coils . Referring to Fig. 9 a helical coil with N turns of conducting wire, as described above, is connected to a pair of conducting parallel plates of area A spaced by distance d via a dielectric material of relative permittivity ε, and everything is surrounded by air (as shown, $N=1$ and $h=0$). The plates have a capacitance $C_p = \varepsilon_o \varepsilon A / d$, which is added to the distributed capacitance of the coil and thus modifies its resonance. Note however, that the presence of the loading capacitor modifies significantly the current distribution inside the wire and therefore the total effective inductance L and total effective capacitance C of the coil are different respectively from L_s and C_s, which are calculated for a self-resonant coil of the same geometry using a sinusoidal current profile. Since some charge is accumulated at the plates of the external loading capacitor, the charge distribution ρ inside the wire is reduced, so $C < C_s$, and thus, from the charge conservation equation, the current distribution $\mathbf{j}$ flattens out, so $L > L_s$. The resonant frequency for this system is $\omega = 1/\sqrt{L(C+C_p)} < \omega_s = 1/\sqrt{L_s C_s}$, and $I(x) \to I_o \cos(\pi x/l) \Rightarrow C \to C_s \Rightarrow \omega \to \omega_s$, as $C_p \to 0$.

In general, the desired CMT parameters can be found for this system, but again a very complicated solution of Maxwell's Equations is required. Instead, we will analyze only a special case, where a reasonable guess for the current distribution can be made. When $C_p \gg C_s > C$, then $\omega \approx 1/\sqrt{LC_p} \ll \omega_s$ and $Z \approx \sqrt{L/C_p} \ll Z_s$, while all the charge is on the plates of the loading capacitor and thus the current distribution is constant along the wire. This allows us now to compute numerically L from Eq.(19). In the case $h = 0$ and N integer, the integral in Eq.(19) can actually be computed analytically, giving the formula $L = \mu_o r\left[\ln(8r/a) - 2\right]N^2$. Explicit analytical formulas are again available for R from Eq.(21) and (22), since $I_{rms} = I_o$, $|\mathbf{p}| \approx 0$ and $|\mathbf{m}| = I_o N \pi r^2$ (namely only the magnetic-dipole term is contributing to radiation), so we can determine also $Q_{abs} = \omega L / R_{abs}$ and $Q_{rad} = \omega L / R_{rad}$. At the end of the calculations, the validity of the assumption of constant current profile is confirmed by checking that indeed the condition $C_p \gg C_s \Leftrightarrow \omega \ll \omega_s$ is satisfied. To satisfy this condition, one could use a large external capacitance, however, this would usually shift the operational frequency lower than the optimal frequency, which we will determine shortly; instead, in typical examples, one often prefers coils with very small self-capacitance C_s to begin with, which usually holds, for the types of coils under consideration, when $N = 1$, so that the self-capacitance comes from the charge distribution across the single turn, which is almost always very small, or when $N > 1$ and

$h \gg 2Na$, so that the dominant self-capacitance comes from the charge distribution across adjacent turns, which is small if the separation between adjacent turns is large.

The external loading capacitance C_p provides the freedom to tune the resonant frequency (for example by tuning A or d). Then, for the particular simple case $h = 0$, for which we have analytical formulas, the total $Q = \omega L / \left(R_{abs} + R_{rad}\right)$ becomes highest at the optimal frequency

$$\omega_Q = \left[\frac{c^4}{\pi}\sqrt{\frac{\varepsilon_o}{2\sigma}} \cdot \frac{1}{aNr^3}\right]^{2/7}, \tag{24}$$

reaching the value

$$Q_{\max} = \frac{6}{7\pi}\left(2\pi^2 \eta_o \frac{\sigma a^2 N^2}{r}\right)^{3/7} \cdot \left[\ln\left(\frac{8r}{a}\right) - 2\right]. \tag{25}$$

At lower frequencies it is dominated by ohmic loss and at higher frequencies by radiation. Note, however, that the formulas above are accurate as long as $\omega_Q \ll \omega_s$ and, as explained above, this holds almost always when $N = 1$, and is usually less accurate when $N > 1$, since $h = 0$ usually implies a large self-capacitance. A coil with large h can be used, if the self-capacitance needs to be reduced compared to the external capacitance, but then the formulas for L and ω_Q, $Q_{\max}$ are again less accurate. Similar qualitative behavior is expected, but a more complicated theoretical model is needed for making quantitative predictions in that case.

The results of the above analysis for two examples of subwavelength modes of $\lambda / r \geq 70$ (namely highly suitable for near-field coupling and well within the quasi-static limit) of coils with $N = 1$ and $h = 0$ at the optimal frequency Eq.(24) are presented in Table 3. To confirm the validity of constant-current assumption and the resulting analytical formulas, mode-solving calculations were also performed using another completely independent method: computational 3D finite-element frequency-domain (FEFD) simulations (which solve Maxwell's Equations in frequency domain exactly apart for spatial discretization) were conducted, in which the boundaries of the conductor were modeled using a complex impedance $\zeta_c = \sqrt{\mu_o \omega / 2\sigma}$ boundary condition, valid as long as $\zeta_c / \zeta_o \ll 1$ (<10^{-5} for copper in the microwave). Table 3 shows Numerical FEFD (and in parentheses analytical) results for the wavelength and absorption, radiation and total loss rates, for two different cases of subwavelength-loop resonant modes. Note that for conducting material copper ($\sigma=5.998\cdot10^7 S/m$) was used. Specific parameters of the plot in Fig. 4 are highlighted in bold in the table. The two methods (analytical and computational) are in good agreement and show that, in some examples, the optimal frequency is in the low-MHz microwave range and the expected quality factors are $Q_{abs} \geq 1000$ and $Q_{rad} \geq 10000$.

Table 3

single coil	λ / r	f	Q_{rad}	Q_{abs}	Q
r=30cm, a=2cm ε=10, A=138cm^2, d=4mm	**111.4 (112.4)**	**8.976 (8.897)**	**29546 (30512)**	**4886 (5117)**	**4193 (4381)**
r=10cm, a=2mm ε=10, A=3.14cm^2, d=1mm	69.7 (70.4)	43.04 (42.61)	10702 (10727)	1545 (1604)	1350 (1395)

Referring to Fig. 10, in some examples, energy is transferred between two capacitively-loaded coils. For the rate of energy transfer between two capacitively-loaded coils 1 and 2 at distance D between their centers, the mutual inductance M_L can be evaluated numerically from Eq.(23) by using constant current distributions in the case $\omega \ll \omega_s$. In the case $h=0$, the coupling is only magnetic and again we have an analytical formula, which, in the quasi-static limit $r \ll D \ll \lambda$ and for the relative orientation shown in Fig. 10, is $M_L \approx \pi\mu_o / 2 \cdot (r_1 r_2)^2 N_1 N_2 / D^3$, which means that $k \propto \left(\sqrt{r_1 r_2} / D\right)^3$ is independent of the frequency ω and the number of turns N_1, N_2. Consequently, the resultant coupling figure-of-merit of interest is

$$U = k\sqrt{Q_1 Q_2} \approx \left(\frac{\sqrt{r_1 r_2}}{D}\right)^3 \cdot \frac{\pi^2 \eta_o \frac{\sqrt{r_1 r_2}}{\lambda} \cdot N_1 N_2}{\prod_{j=1,2}\left(\sqrt{\frac{\pi \eta_o}{\lambda \sigma}} \cdot \frac{r_j}{a_j} N_j + \frac{8}{3}\pi^5 \eta_o \left(\frac{r_j}{\lambda}\right)^4 N_j^2\right)^{1/2}}, \tag{26}$$

which again is more accurate for $N_1 = N_2 = 1$.

From Eq.(26) it can be seen that the optimal frequency ω_U, where the figure-of-merit is maximized to the value $U_{\max}$, is close to the frequency $\omega_{Q_1Q_2}$ at which Q_1Q_2 is maximized, since k does not depend much on frequency (at least for the distances $D \ll \lambda$ of interest for which the quasi-static approximation is still valid). Therefore, the optimal frequency $\omega_U \approx \omega_{Q_1Q_2}$ is mostly independent of the distance D between the two coils and lies between the two frequencies ω_{Q_1} and ω_{Q_2} at which the single-coil Q_1 and Q_2 respectively peak. For same coils, this optimal frequency is given by Eq.(24) and then the strong-coupling factor from Eq.(26) becomes

$$U_{\max} = kQ_{\max} \approx \left(\frac{r}{D}\right)^3 \cdot \frac{3}{7}\left(2\pi^2\eta_o \frac{\sigma a^2 N^2}{r}\right)^{3/7}. \qquad (27)$$

In some examples, one can tune the capacitively-loaded conducting loops or coils, so that their angular eigenfrequencies are close to ω_U within Γ_U, which is half the angular frequency width for which $U > U_{\max}/2$.

Referring to Table 4, numerical FEFD and, in parentheses, analytical results based on the above are shown for two systems each composed of a matched pair of the loaded coils described in Table 3. The average wavelength and loss rates are shown along with the coupling rate and coupling to loss ratio figure-of-merit $U = \kappa / \Gamma$ as a function of the coupling distance D, for the two cases. Note that the average numerical Γ_{rad} shown are slightly different from the single-loop value of Figure 3, analytical results for Γ_{rad} are not shown but the single-loop value is used. (The specific parameters corresponding to the plot in Fig. 10 are highlighted with bold in the table.) Again we chose $N = 1$ to make the constant-current assumption a good one and computed M_L numerically from Eq.(23). Indeed the accuracy can be confirmed by their agreement with the computational FEFD mode-solver simulations, which give κ through the frequency splitting of the two normal modes of the combined system ($\delta_E = 2\kappa$ from Eq.(4)). The results show that for medium distances $D/r = 10-3$ the expected coupling-to-loss ratios are in the range $U \sim 0.5-50$.

Table 4

pair of coils	D/r	Q^{rad}	$Q=\omega/2\Gamma$	$Q_\kappa=\omega/2\kappa$	κ/Γ
r=30cm, a=2cm **ε=10, A=138cm², d=4mm** $\lambda/r\approx 112$ $Q^{abs}\approx 4886$	3	30729	4216	62.6 (63.7)	67.4 (68.7)
	5	29577	4194	235 (248)	17.8 (17.6)
	7	**29128**	**4185**	**589 (646)**	**7.1 (6.8)**
	10	28833	4177	1539 (1828)	2.7 (2.4)
r=10cm, a=2mm ε=10, A=3.14cm², d=1mm $\lambda/r\approx 70$ $Q^{abs}\approx 1546$	3	10955	1355	85.4 (91.3)	15.9 (15.3)
	5	10740	1351	313 (356)	4.32 (3.92)
	7	10759	1351	754 (925)	1.79 (1.51)
	10	10756	1351	1895 (2617)	0.71 (0.53)

2.2.1 Derivation of optimal power-transmission efficiency

Referring to Fig. 11, to rederive and express Eq.(15) in terms of the parameters which are more directly accessible from particular resonant objects, such as the capacitively-loaded conducting loops, one can consider the following circuit-model of the system, where the inductances L_s, L_d represent the source and device loops respectively, R_s, R_d their respective losses, and C_s, C_d are the required corresponding capacitances to achieve for both resonance at frequency ω. A voltage generator V_g is considered to be connected to the source and a load resistance R_l to the device. The mutual inductance is denoted by M.

Then from the source circuit at resonance ($\omega L_s = 1/\omega C_s$):

$$V_g = I_s R_s - j\omega M I_d \Rightarrow \frac{1}{2} V_g^* I_s = \frac{1}{2}\left|I_s\right|^2 R_s + \frac{1}{2} j\omega M I_d^* I_s, \tag{28}$$

and from the device circuit at resonance ($\omega L_d = 1/\omega C_d$):

$$0 = I_d\left(R_d + R_l\right) - j\omega M I_s \Rightarrow j\omega M I_s = I_d\left(R_d + R_l\right) \tag{29}$$

So by substituting Eq.(29) to Eq.(28) and taking the real part (for time-averaged power) we get:

$$P_g = \mathrm{Re}\left\{\frac{1}{2} V_g^* I_s\right\} = \frac{1}{2}\left|I_s\right|^2 R_s + \frac{1}{2}\left|I_d\right|^2\left(R_d + R_l\right) = P_s + P_d + P_l, \tag{30}$$

where we identified the power delivered by the generator $P_g = \mathrm{Re}\left\{V_g^* I_s / 2\right\}$,the power lost inside the source $P_s = \left|I_s\right|^2 R_s / 2$, the power lost inside the device $P_d = \left|I_d\right|^2 R_d / 2$ and the power delivered to the load $P_l = \left|I_d\right|^2 R_l / 2$. Then, the power transmission efficiency is:

$$\eta_{\mathrm{P}} \equiv \frac{P_l}{P_g} = \frac{R_l}{\left|\frac{I_s}{I_d}\right|^2 R_s + \left(R_d + R_l\right)} \overset{(29)}{=} \frac{R_l}{\frac{\left(R_d + R_l\right)^2}{\left(\omega M\right)^2} R_s + \left(R_d + R_l\right)}. \tag{31}$$

If we now choose the load impedance R_l to optimize the efficiency by $\eta_{\mathrm{P}}'\left(R_l\right) = 0$, we get the optimal load impedance

$$\frac{R_{l^*}}{R_d} = \sqrt{1 + \frac{\left(\omega M\right)^2}{R_s R_d}} \tag{32}$$

and the maximum possible efficiency

$$\eta_{\mathrm{P}^*} = \frac{R_{l^*} / R_d - 1}{R_{l^*} / R_d + 1} = \left[\frac{\omega M / \sqrt{R_s R_d}}{1 + \sqrt{1 + \left(\omega M / \sqrt{R_s R_d}\right)^2}}\right]^2. \tag{33}$$

To check now the correspondence with the CMT model, note that $\kappa_l = R_l / 2L_d$, $\Gamma_d = R_d / 2L_d$, $\Gamma_s = R_s / 2L_s$, and $\kappa = \omega M / 2\sqrt{L_s L_d}$, so then $U_l = \kappa_l / \Gamma_d = R_l / R_d$ and $U = \kappa / \sqrt{\Gamma_s \Gamma_d} = \omega M / \sqrt{R_s R_d}$. Therefore, the condition Eq.(32) is identical to the condition Eq.(14) and the optimal efficiency Eq.(33) is identical to the general Eq.(15). Indeed, as the CMT analysis predicted, to get a large efficiency, we need to design a system that has a large strong-coupling factor U.

2.2.2 Optimization of U

The results above can be used to increase or optimize the performance of a wireless energy transfer system, which employs capacitively-loaded coils. For example, from the scaling of Eq.(27) with the different system parameters, one sees that to maximize the system figure-of-merit U, in some examples, one can:

-- Decrease the resistivity of the conducting material. This can be achieved, for example, by using good conductors (such as copper or silver) and/or lowering the temperature. At very low temperatures one could use also superconducting materials to achieve extremely good performance.

-- Increase the wire radius a. In typical examples, this action can be limited by physical size considerations. The purpose of this action is mainly to reduce the resistive losses in the wire by increasing the cross-sectional area through which the electric current is flowing, so one could alternatively use also a Litz wire or a ribbon instead of a circular wire.

-- For fixed desired distance D of energy transfer, increase the radius of the loop r. In typical examples, this action can be limited by physical size considerations, typically especially for the device.

-- For fixed desired distance vs. loop-size ratio D/r, decrease the radius of the loop r. In typical examples, this action can be limited by physical size considerations.

-- Increase the number of turns N. (Even though Eq.(27) is expected to be less accurate for $N > 1$, qualitatively it still provides a good indication that we expect an improvement in the coupling-to-loss ratio with increased N.) In typical examples, this action can be limited by physical size and possible voltage considerations, as will be discussed in following paragraphs.

-- Adjust the alignment and orientation between the two coils. The figure-of-merit is optimized when both cylindrical coils have exactly the same axis of cylindrical symmetry (namely they are "facing" each other). In some examples, particular mutual coil angles and orientations that lead to zero mutual inductance (such as the orientation where the axes of the two coils are perpendicular and the centers of the two coils are on one of the two axes) should be avoided.

-- Finally, note that the height of the coil h is another available design parameter, which can have an impact to the performance similar to that of its radius r, and thus the design rules can be similar.

The above analysis technique can be used to design systems with desired parameters. For example, as listed below, the above described techniques can be used to determine the cross sectional radius a of the wire which one should use when designing as system two same single-turn loops with a given radius in order to achieve a specific performance in terms of $U = \kappa / \Gamma$ at a given D / r between them, when the material is copper (σ=*5.998·10^{7}S/m*):

$$D / r = 5,\ U \geq 10,\ r = 30cm \Rightarrow a \geq 9mm$$
$$D / r = 5,\ U \geq 10,\ r = 5cm \Rightarrow a \geq 3.7mm$$
$$D / r = 5,\ U \geq 20,\ r = 30cm \Rightarrow a \geq 20mm$$
$$D / r = 5,\ U \geq 20,\ r = 5cm \Rightarrow a \geq 8.3mm$$
$$D / r = 10,\ U \geq 1,\ r = 30cm \Rightarrow a \geq 7mm$$
$$D / r = 10,\ U \geq 1,\ r = 5cm \Rightarrow a \geq 2.8mm$$
$$D / r = 10,\ U \geq 3,\ r = 30cm \Rightarrow a \geq 25mm$$
$$D / r = 10,\ U \geq 3,\ r = 5cm \Rightarrow a \geq 10mm$$

Similar analysis can be done for the case of two dissimilar loops. For example, in some examples, the device under consideration is very specific (e.g. a laptop or a cell phone), so the dimensions of the device object (r_d, h_d, a_d, N_d) are very restricted. However, in some such examples, the restrictions on the source object (r_s, h_s, a_s, N_s) are much less, since the source can, for example, be placed under the floor or on the ceiling. In such cases, the desired distance is often well defined, based on the application (e.g. $D \sim 1m$ for charging a laptop on a table wirelessly from the floor). Listed below are examples (simplified to the case $N_s = N_d = 1$ and $h_s = h_d = 0$) of how one can vary the dimensions of the source object to achieve the desired system performance in terms of $U_{sd} = \kappa / \sqrt{\Gamma_s \Gamma_d}$, when the material is again copper (σ=*5.998·10^{7}S/m*):

$$D = 1.5m,\ U_{sd} \geq 15,\ r_d = 30cm,\ a_d = 6mm \Rightarrow r_s = 1.158m,\ a_s \geq 5mm$$
$$D = 1.5m,\ U_{sd} \geq 30,\ r_d = 30cm,\ a_d = 6mm \Rightarrow r_s = 1.15m,\ a_s \geq 33mm$$
$$D = 1.5m, U_{sd} \geq 1,\ r_d = 5cm,\ a_d = 4mm \Rightarrow r_s = 1.119m,\ a_s \geq 7mm$$
$$D = 1.5m,\ U_{sd} \geq 2,\ r_d = 5cm,\ a_d = 4mm \Rightarrow r_s = 1.119m,\ a_s \geq 52mm$$
$$D = 2m,\ U_{sd} \geq 10,\ r_d = 30cm,\ a_d = 6mm \Rightarrow r_s = 1.518m,\ a_s \geq 7mm$$
$$D = 2m,\ U_{sd} \geq 20,\ r_d = 30cm,\ a_d = 6mm \Rightarrow r_s = 1.514m,\ a_s \geq 50mm$$
$$D = 2m,\ U_{sd} \geq 0.5,\ r_d = 5cm,\ a_d = 4mm \Rightarrow r_s = 1.491m,\ a_s \geq 5mm$$
$$D = 2m,\ U_{sd} \geq 1,\ r_d = 5cm,\ a_d = 4mm \Rightarrow r_s = 1.491m,\ a_s \geq 36mm$$

2.2.3 Optimization of k

As described below, in some examples, the quality factor Q of the resonant objects is limited from external perturbations and thus varying the coil parameters cannot lead to improvement in Q. In such cases, one can opt to increase the strong-coupling factor U by increasing the coupling factor k. The coupling does not depend on the frequency and the number of turns. Therefore, in some examples, one can:

-- Increase the wire radii a_1 and a_2. In typical examples, this action can be limited by physical size considerations.

-- For fixed desired distance D of energy transfer, increase the radii of the coils r_1 and r_2. In typical examples, this action can be limited by physical size considerations, typically especially for the device.

-- For fixed desired distance vs. coil-sizes ratio $D/\sqrt{r_1 r_2}$, only the weak (logarithmic) dependence of the inductance remains, which suggests that one should decrease the radii of the coils r_1 and r_2. In typical examples, this action can be limited by physical size considerations.

-- Adjust the alignment and orientation between the two coils. In typical examples, the coupling is optimized when both cylindrical coils have exactly the same axis of cylindrical symmetry (namely they are "facing" each other). Particular mutual coil angles and orientations that lead to zero mutual inductance (such as the orientation where the axes of the two coils are perpendicular and the centers of the two coils are on one of the two axes) should obviously be avoided.

-- Finally, note that the heights of the coils h_1 and h_2 are other available design parameters, which can have an impact to the coupling similar to that of their radii r_1 and r_2, and thus the design rules can be similar.

Further practical considerations apart from efficiency, e.g. physical size limitations, will be discussed in detail below.

2.2.4 Optimization of overall system performance

In many cases, the dimensions of the resonant objects will be set by the particular application at hand. For example, when this application is powering a laptop or a cell-phone, the device resonant object cannot have dimensions larger than those of the laptop or cell-phone respectively. In particular, for a system of two loops of specified dimensions, in terms of loop radii $r_{s,d}$ and wire radii $a_{s,d}$, the independent parameters left to adjust for the system optimization are: the number of turns $N_{s,d}$, the frequency *f*, the power-load consumption rate $\kappa_l = R_l / 2L_d$ and the power-generator feeding rate $\kappa_g = R_g / 2L_s$, where R_g is the internal (characteristic) impedance of the generator.

In general, in various examples, the primary dependent variable that one wants to increase or optimize is the overall efficiency η. However, other important variables need to be taken into consideration upon system design. For example, in examples featuring capacitively-loaded coils, the design can be constrained by, for example, the currents flowing inside the wires $I_{s,d}$ and the voltages across the capacitors $V_{s,d}$. These limitations can be important because for ~*Watt* power applications the values for these parameters can be too large for the wires or the capacitors respectively to handle. Furthermore, the total loaded (by the load) quality factor of the device $Q_{d[l]} = \omega / 2(\Gamma_d + \Gamma_l) = \omega L_d / (R_d + R_l)$ and the total loaded (by the generator) quality factor of the source $Q_{s[g]} = \omega / 2(\Gamma_s + \Gamma_g) = \omega L_s / (R_s + R_g)$ are quantities that should be preferably small, because to match the source and device resonant frequencies to within their Q's, when those are very large, can be challenging experimentally and more sensitive to slight variations. Lastly, the radiated powers $P_{s,rad}$ and $P_{d,rad}$ should be minimized for concerns about far-field interference and safety, even though, in general, for a magnetic, non-radiative scheme they are already typically small. In the following, we examine then the effects of each one of the independent variables on the dependent ones.

We define a new variable wp to express the power-load consumption rate for some particular value of U through $U_l = \kappa_l / \Gamma_d = \sqrt{1 + wp \cdot U^2}$. Then, in some examples, values which impact the choice of this rate are: $U_l = 1 \Leftrightarrow wp = 0$ to minimize the required energy stored in the source (and therefore I_s and V_s), $U_l = \sqrt{1 + U^2} > 1 \Leftrightarrow wp = 1$ to maximize the efficiency, as seen earlier, or $U_l \gg 1 \Leftrightarrow wp \gg 1$ to decrease the required energy stored in the device (and therefore I_d and V_d) and to decrease or minimize $Q_{d[l]}$. Similar is the impact of the choice of the power-generator feeding rate $U_g = \kappa_g / \Gamma_s$, with the roles of the source/device and generator/load reversed.

In some examples, increasing N_s and N_d increases Q_s and Q_d, and thus U and the efficiency significantly, as seen before. It also decreases the currents I_s and I_d, because the inductance of the loops increases, and thus the energy $W_{s,d} = L_{s,d} \left|I_{s,d}\right|^2 / 2$ required for given output power P_l can be achieved with smaller currents. However, in some examples, increasing N_d and thus Q_d can increase $Q_{d[l]}$, $P_{d,rad}$ and the voltage across the device capacitance V_d. Similar can be the impact of increasing N_s on $Q_{s[g]}$, $P_{s,rad}$ and V_s. As a conclusion, in some examples, the number of turns N_s and N_d should be chosen large enough (for high efficiency) but such that they allow for reasonable voltages, loaded Q's and/or powers radiated.

With respect to the resonant frequency, again, there is an optimal one for efficiency. Close to that optimal frequency $Q_{d[l]}$ and/or $Q_{s[g]}$ can be approximately maximum. In some examples, for lower frequencies the currents typically get larger but the voltages and radiated powers get smaller. In some examples, one should pick either the frequency that maximizes the efficiency or somewhat lower.

One way to decide on an operating regime for the system is based on a graphical method. Consider two loops of $r_s = 25cm,\ r_d = 15cm,\ h_s = h_d = 0,\ a_s = a_d = 3mm$ and distance $D = 2m$ between them. In Fig. 12, we plot some of the above dependent variables (currents, voltages and radiated powers normalized to *1Watt* of output power) in terms of frequency f and N_d, given some choice for wp and N_s. Fig. 12 depicts the trend of system performance explained above. In Fig. 13, we make a contour plot of the dependent variables as functions of both frequency and *wp* but for both N_s and N_d fixed. For example, a reasonable choice of parameters for the system of two loops with the dimensions given above are: $N_s = 2,\ N_d = 6$, f*=10MHz* and *wp=10*, which gives the following performance characteristics: $\eta = 20.6\%$, $Q_{d[l]} = 1264$, $I_s = 7.2A, I_d = 1.4A$,

$V_s = 2.55kV$, $V_d = 2.30kV$, $P_{s,rad} = 0.15W$, $P_{d,rad} = 0.006W$. Note that the results in Figs. 12, 13 and the just above calculated performance characteristics are made using the analytical formulas provided above, so they are expected to be less accurate for large values of N_s, N_d, but still they give a good estimate of the scalings and the orders of magnitude.

Finally, one could additionally optimize for the source dimensions, since usually only the device dimensions are limited, as discussed earlier. Namely, one can add r_s and a_s in the set of independent variables and optimize with respect to these too for all the dependent variables of the problem (we saw how to do this only for efficiency earlier). Such an optimization would lead to improved results.

In this description, we propose that, if one ensures operation in the strongly-coupled regime at mid-range distances, at least medium-power transmission (~W) at mid-range distances with high efficiency is possible.

2.3 Inductively-loaded conducting rods

A straight conducting rod of length $2h$ and cross-sectional radius a has distributed capacitance and distributed inductance, and therefore it supports a resonant mode of angular frequency ω. Using the same procedure as in the case of self-resonant coils, one can define an effective total inductance L and an effective total capacitance C of the rod through formulas Eqs.(19) and (20). With these definitions, the resonant angular frequency and the effective impedance are given again by the common formulas $\omega = 1/\sqrt{LC}$ and $Z = \sqrt{L/C}$ respectively. To calculate the total inductance and capacitance, one can assume again a sinusoidal current profile along the length of the conducting wire. When interested in the lowest mode, if we denote by x the coordinate along the conductor, such that it runs from $-h$ to $+h$, then the current amplitude profile would have the form $I(x) = I_o \cos(\pi x/2h)$, since it has to be zero at the open ends of the rod. This is the well-known half-wavelength electric dipole resonant mode.

In some examples, one or more of the resonant objects are inductively-loaded conducting rods. Referring to Fig. 14, a straight conducting rod of length $2h$ and cross-sectional radius a, as in the previous paragraph, is cut into two equal pieces of length h, which are connected via a coil wrapped around a magnetic material of relative permeability μ, and everything is surrounded by air. The coil has an inductance L_C, which is added to the distributed inductance of the rod and thus modifies its resonance. Note however, that the presence of the center-loading inductor modifies significantly the current distribution inside the wire and therefore the total effective inductance L and total effective capacitance C of the rod are different respectively from L_s and C_s, which are calculated for a self-resonant rod of the same total length using a sinusoidal current profile, as in the previous paragraph. Since some current is running inside the coil of the external loading inductor, the current distribution $\mathbf{j}$ inside the rod is reduced, so $L < L_s$, and thus, from the charge conservation equation, the linear charge distribution ρ_l flattens out towards the center (being positive in one side of the rod and negative in the other side of the rod, changing abruptly through the inductor), so $C > C_s$. The resonant frequency for this system is $\omega = 1/\sqrt{(L+L_c)C} < \omega_s = 1/\sqrt{L_s C_s}$, and $I(x) \to I_o \cos(\pi x/2h) \Rightarrow L \to L_s \Rightarrow \omega \to \omega_s$, as $L_c \to 0$.

In general, the desired CMT parameters can be found for this system, but again a very complicated solution of Maxwell's Equations is generally required. In a special case, a reasonable estimate for the current distribution can be made. When $L_c \gg L_s > L$, then $\omega \approx 1/\sqrt{L_c C} \ll \omega_s$ and $Z \approx \sqrt{L_c / C} \gg Z_s$, while the current distribution is triangular along the rod (with maximum at the center-loading inductor and zero at the ends) and thus the charge distribution is positive constant on one half of the rod and equally negative constant on the other side of the rod. This allows us to compute numerically C from Eq.(20). In this case, the integral in Eq.(20) can actually be computed analytically, giving the formula $1/C = 1/(\pi\varepsilon_o h)\left[\ln(h/a) - 1\right]$. Explicit analytical formulas are again available for R from Eq.(21) and (22), since $I_{rms} = I_o$, $|\mathbf{p}| = q_o h$ and $|\mathbf{m}| = 0$ (namely only the electric-dipole term is contributing to radiation), so we can determine also $Q_{abs} = 1/\omega C R_{abs}$ and $Q_{rad} = 1/\omega C R_{rad}$. At the end of the calculations, the validity of the assumption of triangular current profile is confirmed by checking that indeed the condition $L_c \gg L_s \Leftrightarrow \omega \ll \omega_s$ is satisfied. This condition is relatively easily satisfied, since typically a conducting rod has very small self-inductance L_s to begin with.

Another important loss factor in this case is the resistive loss inside the coil of the external loading inductor L_c and it depends on the particular design of the inductor. In some examples, the inductor is made of a Brooks coil, which is the coil geometry which, for fixed wire length, demonstrates the highest inductance and thus quality factor. The Brooks coil geometry has N_{Bc} turns of conducting wire of cross-sectional radius a_{Bc} wrapped around a cylindrically symmetric coil former, which forms a coil with a square cross-section of side r_{Bc}, where the inner side of the square is also at radius r_{Bc} (and thus the outer side of the square is at radius $2r_{Bc}$), therefore $N_{Bc} \approx \left(r_{Bc}/2a_{Bc}\right)^2$. The inductance of the coil is then $L_C = 2.0285\mu_o r_{Bc} N_{Bc}^2 \approx 2.0285\mu_o r_{Bc}^5 / 8a_{Bc}^4$ and its resistance $R_C \approx \frac{1}{\sigma}\frac{l_{Bc}}{\pi a_{Bc}^2}\sqrt{1+\frac{\mu_o\omega\sigma}{2}\left(\frac{a_{Bc}}{2}\right)^2}$, where the total wire length is $l_{Bc} \approx 2\pi\left(3r_{Bc}/2\right)N_{Bc} \approx 3\pi r_{Bc}^3/4a_{Bc}^2$ and we have used an approximate square-root law for the transition of the resistance from the dc to the ac limit as the skin depth varies with frequency.

The external loading inductance L_c provides the freedom to tune the resonant frequency. For example, for a Brooks coil with a fixed size r_{Bc}, the resonant frequency can be reduced by increasing the number of turns N_{Bc} by decreasing the wire cross-sectional radius a_{Bc}. Then the desired resonant angular frequency $\omega = 1/\sqrt{L_c C}$ is achieved for $a_{Bc} \approx \left(2.0285\mu_o r_{Bc}^5 \omega^2 C\right)^{1/4}$ and the resulting coil quality factor is $Q_c \approx 0.169\mu_o\sigma r_{Bc}^2\omega / \sqrt{1+\omega^2\mu_o\sigma\sqrt{2.0285\mu_o\left(r_{Bc}/4\right)^5 C}}$. Then, for the particular simple case $L_c \gg L_s$, for which we have analytical formulas, the total $Q = 1/\omega C\left(R_c + R_{abs} + R_{rad}\right)$ becomes highest at some optimal frequency ω_Q, reaching the value $Q_{\max}$, both determined by the loading-inductor specific design. For example, for the Brooks-coil procedure described above, at the optimal frequency $Q_{\max} \approx Q_c \approx 0.8\left(\mu_o \sigma^2 r_{Bc}^3 / C\right)^{1/4}$. At lower frequencies it is dominated by ohmic loss inside the inductor coil and at higher frequencies by radiation. Note, again, that the above formulas are accurate as long as $\omega_Q \ll \omega_s$ and, as explained above, this is easy to design for by using a large inductance.

The results of the above analysis for two examples, using Brooks coils, of subwavelength modes of $\lambda / h \geq 200$ (namely highly suitable for near-field coupling and well within the quasi-static limit) at the optimal frequency ω_Q are presented in Table 5.

Table 5 shows in parentheses (for similarity to previous tables) analytical results for the wavelength and absorption, radiation and total loss rates, for two different cases of subwavelength-loop resonant modes. Note that for conducting material copper ($\sigma=5.998 \cdot 10^7 S/m$) was used. The results show that, in some examples, the optimal frequency is in the low-MHz microwave range and the expected quality factors are $Q_{abs} \geq 1000$ and $Q_{rad} \geq 100000$.

Table 5

single rod	λ / h	f (MHz)	Q_{rad}	Q_{abs}	Q
h=30cm, a=2cm μ=1, r_{Bc}=2cm, a_{Bc}=0.88mm, N_{Bc}=129	(403.8)	(2.477)	($2.72*10^6$)	(7400)	(7380)
h=10cm, a=2mm μ=1, r_{Bc}=5mm, a_{Bc}=0.25mm, N_{Bc}=103	(214.2)	(14.010)	($6.92*10^5$)	(3908)	(3886)

In some examples, energy is transferred between two inductively-loaded rods. For the rate of energy transfer between two inductively-loaded rods 1 and 2 at distance D between their centers, the mutual capacitance M_C can be evaluated numerically from Eq.(23) by using triangular current distributions in the case $\omega \ll \omega_s$. In this case, the coupling is only electric and again we have an analytical formula, which, in the quasi-static limit $h \ll D \ll \lambda$ and for the relative orientation such that the two rods are aligned on the same axis, is $1/M_C \approx 1/2\pi\varepsilon_o \cdot (h_1 h_2)^2 / D^3$, which means that $k \propto \left(\sqrt{h_1 h_2} / D\right)^3$ is independent of the frequency ω. One can then get the resultant strong-coupling factor U.

It can be seen that the optimal frequency ω_U, where the figure-of-merit is maximized to the value $U_{\max}$, is close to the frequency $\omega_{Q_1Q_2}$, where Q_1Q_2 is maximized, since k does not depend much on frequency (at least for the distances $D \ll \lambda$ of interest for which the quasi-static approximation is still valid). Therefore, the optimal frequency $\omega_U \approx \omega_{Q_1Q_2}$ is mostly independent of the distance D between the two rods and lies between the two frequencies ω_{Q_1} and ω_{Q_2} at which the single-rod Q_1 and Q_2 respectively peak. In some typical examples, one can tune the inductively-loaded conducting rods, so that their angular eigenfrequencies are close to ω_U within Γ_U, which is half the angular frequency width for which $U > U_{\max}/2$.

Referring to Table 6, in parentheses (for similarity to previous tables) analytical results based on the above are shown for two systems each composed of a matched pair of the loaded rods described in Table 5. The average wavelength and loss rates are shown along with the coupling rate and coupling to loss ratio figure-of-merit $U = \kappa / \Gamma$ as a function of the coupling distance D, for the two cases. Note that for Γ_{rad} the single-rod value is used. Again we chose $L_c \gg L_s$ to make the triangular-current assumption a good one and computed M_C numerically from Eq.(23). The results show that for medium distances $D/h = 10-3$ the expected coupling-to-loss ratios are in the range $U \sim 0.5-100$.

Table 6

pair of rods	D/h	$Q_\kappa = 1/k$	U
h=30cm, a=2cm μ=1, r_{Bc}=2cm, a_{Bc}=0.88mm, N_{Bc}=129 $\lambda/h \approx 404$ $Q \approx 7380$	3	(70.3)	(105.0)
	5	(389)	(19.0)
	7	(1115)	(6.62)
	10	(3321)	(2.22)
h=10cm, a=2mm μ=1, r_{Bc}=5mm, a_{Bc}=0.25mm, N_{Bc}=103 $\lambda/h \approx 214$ $Q \approx 3886$	3	(120)	(32.4)
	5	(664)	(5.85)
	7	(1900)	(2.05)
	10	(5656)	(0.69)

2.4 Dielectric disks

In some examples, one or more of the resonant objects are dielectric objects, such as disks. Consider a two dimensional dielectric disk object, as shown in Fig. 15(a), of radius r and relative permittivity ε surrounded by air that supports high-Q "whispering-gallery" resonant modes. The loss mechanisms for the energy stored inside such a resonant system are radiation into free space and absorption inside the disk material. High-Q_{rad} and long-tailed subwavelength resonances can be achieved when the dielectric permittivity ε is large and the azimuthal field variations are slow (namely of small principal number m). Material absorption is related to the material loss tangent: $Q_{abs} \sim \mathrm{Re}\{\varepsilon\}/\mathrm{Im}\{\varepsilon\}$. Mode-solving calculations for this type of disk resonances were performed using two independent methods: numerically, 2D finite-difference frequency-domain (FDFD) simulations (which solve Maxwell's Equations in frequency domain exactly apart for spatial discretization) were conducted with a resolution of *30pts/r*; analytically, standard separation of variables (SV) in polar coordinates was used.

Table 7

single disk	λ / r	Q^{abs}	Q^{rad}	Q
Re{ε}=147.7, m=2	**20.01 (20.00)**	**10103 (10075)**	**1988 (1992)**	**1661 (1663)**
Re{ε}=65.6, m=3	9.952 (9.950)	10098 (10087)	9078 (9168)	4780 (4802)

The results for two TE-polarized dielectric-disk subwavelength modes of $\lambda / r \geq 10$ are presented in Table 7. Table 7 shows numerical FDFD (and in parentheses analytical SV) results for the wavelength and absorption, radiation and total loss rates, for two different cases of subwavelength-disk resonant modes. Note that disk-material loss-tangent $\mathrm{Im}\{\varepsilon\}/\mathrm{Re}\{\varepsilon\}=10^{-4}$ was used. (The specific parameters corresponding to the plot in Fig. 15(a) are highlighted with bold in the table.) The two methods have excellent agreement and imply that for a properly designed resonant low-loss-dielectric object values of $Q_{rad} \geq 2000$ and $Q_{abs} \sim 10000$ are achievable. Note that for the 3D case the computational complexity would be immensely increased, while the physics would not be significantly different. For example, a spherical object of $\varepsilon=147.7$ has a whispering gallery mode with $m=2$, $Q_{rad}=13962$, and $\lambda/r=17$.

The required values of ε, shown in Table 7, might at first seem unrealistically large. However, not only are there in the microwave regime (appropriate for approximately meter-range coupling applications) many materials that have both reasonably high enough dielectric constants and low losses (e.g. Titania, Barium tetratitanate, Lithium tantalite etc.), but also ε could signify instead the effective index of other known subwavelength surface-wave systems, such as surface modes on surfaces of metallic materials or plasmonic (metal-like, negative-ε) materials or metallo-dielectric photonic crystals or plasmono-dielectric photonic crystals.

To calculate now the achievable rate of energy transfer between two disks 1 and 2, as shown in Fig. 15(b) we place them at distance D between their centers. Numerically, the FDFD mode-solver simulations give κ through the frequency splitting of the normal modes of the combined system ($\delta_E = 2\kappa$ from Eq.(4)), which are even and odd superpositions of the initial single-disk modes; analytically, using the expressions for the separation-of-variables eigenfields $\mathbf{E}_{1,2}(\mathbf{r})$ CMT gives κ through

$$\kappa = \omega_1 / 2 \cdot \int d^3 r \varepsilon_2(r) E_2^*(r) E_1(r) / \int d^3 r \varepsilon(r) \left| E_1(r) \right|^2,$$

where $\varepsilon_j(\mathbf{r})$ and $\varepsilon(\mathbf{r})$ are the dielectric functions that describe only the disk j (minus the constant ε_o background) and the whole space respectively. Then, for medium distances $D/r = 10-3$ and for non-radiative coupling such that $D < 2r_C$, where $r_C = m\lambda / 2\pi$ is the radius of the radiation caustic, the two methods agree very well, and we finally find , as shown in Table 8, strong-coupling factors in the range $U \sim 1-50$. Thus, for the analyzed examples, the achieved figure-of-merit values are large enough to be useful for typical applications, as discussed below.

Table 8

two disks	D/r	Q^{rad}	$Q=\omega/2\Gamma$	$\omega/2\kappa$	κ/Γ
Re{ε}=147.7, m=2	3	2478	1989	46.9 (47.5)	42.4 (35.0)
$\lambda/r\approx 20$	5	2411	1946	298.0 (298.0)	6.5 (5.6)
$Q^{abs}\approx 10093$	7	2196	1804	769.7 (770.2)	2.3 (2.2)
	10	**2017**	**1681**	**1714 (1601)**	**0.98 (1.04)**
Re{ε}=65.6, m=3	3	7972	4455	144 (140)	30.9 (34.3)
$\lambda/r\approx 10$	5	9240	4824	2242 (2083)	2.2 (2.3)
$Q^{abs}\approx 10096$	7	9187	4810	7485 (7417)	0.64 (0.65)

Note that even though particular examples are presented and analyzed above as examples of systems that use resonant electromagnetic coupling for wireless energy transfer, those of self-resonant conducting coils, capacitively-loaded resonant conducting coils, inductively-loaded resonant conducting rods and resonant dielectric disks, any system that supports an electromagnetic mode with its electromagnetic energy extending much further than its size can be used for transferring energy. For example, there can be many abstract geometries with distributed capacitances and inductances that support the desired kind of resonances. In some examples, the resonant structure can be a dielectric sphere. In any one of these geometries, one can choose certain parameters to increase and/or optimize U or, if the Q's are limited by external factors, to increase and/or optimize for k or, if other system performance parameters are of importance, to optimize those.

3 Coupled-Mode Theory for prediction of far-field radiation interference

The two objects in an energy-transfer system generate radiation, which can sometimes be a significant part of the intrinsic losses, and can interfere in the far field. In the previous Sections, we analyzed systems, where this interference phenomenon was not in effect. In this description, we will repeat the analysis, including the interference effects and will show how it can be used to further enhance the power transmission efficiency and/or the radiated power.

The coupled-mode equations of Eqs.(1) fail to predict such an interference phenomenon. In fact, the inability to predict interference phenomena has often been considered inherent to coupled-mode theory (CMT). However, we show here that making only a simple extension to this model, it can actually very successfully predict such interference. The root of the problem stems from the fact that the coupling coefficients were tacitly assumed to be real. This is usually the case when dealing with proper (real) eigenmodes of a Hermitian (lossless) operator. However, this assumption fails when losses are included, as is for example the current case dealing with generally non-proper (leaky, radiative) eigenmodes of a non-Hermitian (lossy) operator. In this case, the coupling-matrix elements will generally be complex and their imaginary parts will be shown to be directly related to far-field radiation interference.

Imagine a system of many resonators in proximity to each other. When their resonances have close enough frequencies compared to their coupling rates, the CMT assumption is that the total-system field ψ is approximately determined only by these resonances as the superposition $\psi(t) = \sum_n a_n(t)\psi_n$, where ψ_n is the eigenfield of the resonance n normalized to unity energy, and a_n is the field amplitude inside it, which corresponds, due to the normalization, to $|a_n|^2$ stored energy. The fundamental Coupled-Mode Equations (CME) of CMT are then those of the evolution of the vector $\boldsymbol{a} = \{a_n\}$

$$\frac{d}{dt}a = -i\overline{\boldsymbol{\Omega}} \cdot a + i\overline{\mathbf{K}} \cdot a \tag{34}$$

where the frequency matrix $\overline{\boldsymbol{\Omega}}$ and the coupling matrix $\overline{\mathbf{K}}$ are found usually using a Perturbation Theory (PT) approach.

We restate here one of the many perturbative formulations of CMT in a system of ElectroMagnetic (EM) resonators: Let $\mu = \mu_o$ and $\epsilon = \epsilon_o + \sum_n \epsilon_n$ be the magnetic-permeability and dielectric-permittivity functions of space that describe the whole system, where ϵ_n is the permittivity of only the dielectric, reciprocal and generally anisotropic object n of volume V_n, in excess to the constant μ_o, ϵ_o background space. Each resonator n, when alone in the background space, supports a resonant eigenmode of complex frequency $\Omega_n = \omega_n - i\Gamma_n$ and field profiles $\psi_n = [\mathbf{E}_n, \mathbf{H}_n]$ normalized to unity energy, satisfying the equations $\nabla \times \mathbf{E}_n = i\Omega_n\mu_o\mathbf{H}_n$ and $\nabla \times \mathbf{H}_n = -i\Omega_n(\epsilon_o + \epsilon_n)\mathbf{E}_n$, and the boundary condition $\hat{\mathbf{n}} \times \mathbf{E}_n = 0$ on the potential metallic surface S_n of object n. The whole system fields $\psi = [\mathbf{E}, \mathbf{H}]$ satisfy the equations $\nabla \times \mathbf{E} = -\mu \frac{\partial}{\partial t}\mathbf{H}$ and $\nabla \times \mathbf{H} = \epsilon \frac{\partial}{\partial t}\mathbf{E}$, and the boundary condition $\hat{\mathbf{n}} \times \mathbf{E} = 0$ on $S = \sum_n S_n$. Then, start by expanding $\nabla \cdot (\mathbf{E} \times \mathbf{H}_n^- - \mathbf{E}_n^- \times \mathbf{H})$ and integrating over all space, apply the CMT superposition assumption, and finally use the PT argument that, when the coupling-rates between the resonators are small compared to their frequencies, within a sum only terms of lowest order in this small perturbation need to be kept. The result is the CME of Eq.(34), with $\overline{\mathbf{\Omega}} = \overline{\mathbf{W}}^{-1} \cdot \overline{\mathbf{\Omega}}_o \cdot \overline{\mathbf{W}}$, $\overline{\mathbf{K}} = \overline{\mathbf{W}}^{-1} \cdot \mathbf{K}_o$, where $\overline{\mathbf{\Omega}}_o = \mathrm{diag}\{\Omega_n\}$,

$$K_{o,nm} = \frac{\Omega_n}{4} \int_{V_m} dv \left(\mathbf{E}_n^- \cdot \varepsilon_m \cdot \mathbf{E}_m \right) + \frac{i}{4} \oint_{S_m} da \hat{\mathrm{n}} \cdot \left(\mathbf{E}_n^- \times \mathbf{H}_m \right) \quad (35)$$

$$W_{nm} = \frac{1}{4} \int_{V_m} dv \left(\mathbf{E}_n^- \cdot \varepsilon \cdot \mathbf{E}_m + \mathbf{H}_n^- \cdot \mu \cdot \mathbf{H}_m \right) \quad (36)$$

and where $\psi_n = [\mathbf{E}_n^-, \mathbf{H}_n^-]$ satisfy the time-reversed equations (where $\Omega_n \to -\Omega_n$). The choice of these fields in the analysis rather than $\psi_n^* = [\mathbf{E}_n^*, \mathbf{H}_n^*]$ allows to treat also lossy (due to absorption and/or radiation) but reciprocal systems (so $\overline{\mathbf{K}}$ is complex symmetric but non-Hermitian). In the limit, though, of weak loss (high-Q resonances), these two sets of fields can be approximately equal. Therefore, again to lowest order, $\overline{\mathbf{W}} \approx \overline{\mathbf{I}}$, due to the unity-energy normalization, so $\overline{\mathbf{\Omega}} \approx \overline{\mathbf{\Omega}}_o$ and for $\overline{\mathbf{K}}$ the off-diagonal terms

$$K_{nm} \approx K_{o,mn} \approx \frac{i}{4} \int_{V_m} dv \left(\mathbf{E}_n^* \cdot \mathbf{J}_m \right); \quad n \neq m \quad (37)$$

where $\mathbf{J}_m$ includes both the volume-polarization currents $\mathbf{J}_{p,m} = -i\Omega_m \epsilon_m \mathbf{E}_m$ in V_m and the surface electric currents $\mathbf{J}_{s,m} = \hat{\mathbf{n}} \times \mathbf{H}_m$ on S_m, while the diagonal terms K_{nn} are higher-order small and can often lead to anomalous coupling-induced frequency shifts. The term of Eq.(37) can generally be complex $K_{\mathrm{nm}} = \kappa_{\mathrm{nm}} + i\Lambda_{\mathrm{nm}}$ and, even though the physical interpretation of its real part is well understood, as describing the coupling between the resonators, it is not so the case for its imaginary part

$$\Lambda_{nm} = \frac{1}{4}\mathrm{Re}\left\{\int_{V_m} dv\left[i\omega\mathbf{A}_n - \nabla\phi_n\right]^* \cdot \mathbf{J}_m\right\}$$
$$= \frac{1}{4}\mathrm{Re}\left\{\int_{V_m} dv\left[\int_{V_n} dv \frac{e^{ik|\mathbf{r_m}-\mathbf{r_n}|}}{4\pi|\mathbf{r_m}-\mathbf{r_n}|}\left(i\omega\mu_0\mathbf{J}_n + \frac{\rho_n}{\varepsilon_0}\nabla\right)\right]^* \cdot \mathbf{J}_m \qquad (38)$$
$$= \frac{\omega}{16\pi}\int_{V_m} dv \int_{V_n} dv\, \mathrm{Re}\left\{\left(\frac{\rho_n^*\rho_m}{\varepsilon_0} - \mu_0\mathbf{J}_n^* \cdot \mathbf{J}_m\right)\frac{e^{ik|\mathbf{r}_m-\mathbf{r}_n|}}{4\pi|\mathbf{r}_m-\mathbf{r}_n|}\right\}$$

where integration by parts was used for the $\nabla\phi_n$ term and the continuity equation $\nabla \cdot \mathbf{J} = i\omega\rho$, with ρ being the volume charge density.

Towards understanding this term, let us consider two resonators 1, 2 and evaluate from Eqs.(34) the total power lost from the system

$$P_{loss} = -\frac{d}{dt}\left(|a_1|^2 + |a_2|^2\right) = 2\Gamma_1|a_1|^2 + 2\Gamma_2|a_2|^2 + 4\Lambda_{12}\,\mathrm{Re}\left\{a_1^* a_2\right\} \qquad (39)$$

Clearly, the term involving an interaction between the two objects should not relate to material absorption, since this is a very localized process inside each object. We therefore split this lost power into absorbed and radiated in the following way

$$P_{abs} = 2\Gamma_{1,abs}|a_1|^2 + 2\Gamma_{2,abs}|a_2|^2 \qquad (40)$$

$$P_{rad} = 2\Gamma_{1,rad}|a_1|^2 + 2\Gamma_{2,rad}|a_2|^2 + 4\Lambda_{12}\,\mathrm{Re}\left\{a_1^* a_2\right\} \qquad (41)$$

so Λ_{12} is associated with the radiation from the two-object system. However, we have a tool to compute this radiated power separately: Antenna Theory (AT).

Let $\zeta_o = \sqrt{\mu_o/\epsilon_o}$ and $c_o = 1/\sqrt{\mu_o\epsilon_o}$ be the background impedance and light-velocity, and $f = (g, \mathbf{f}) = \int_V dv' J^\nu(\mathbf{r}')e^{-i\mathbf{k}\cdot\mathbf{r}'}$ the moment of the current-distribution 4-vector $J^\nu = (c_o\rho, \mathbf{J})$ of an electromagnetic resonator, where unity-energy normalization is again assumed for J^{ν} and $g = \hat{\mathbf{k}} \cdot \mathbf{f}$, as can be shown using the continuity equation and integration by parts. The power radiated from one EM resonator is:

$$P_{rad} = 2\Gamma_{rad}\left|a\right|^2 = \frac{\varsigma_o k^2}{32\pi^2}\left(\oint d\Omega\left|f\right|^2\right)\left|a\right|^2 \tag{42}$$

where $|f|^2 = f^* \cdot f \equiv |\mathbf{f}|^2 - |g|^2$. The power radiated from an 'array' of two resonators 1 and 2, at vector-distance $\mathbf{D}$ between their centers, is given by:

$$\begin{aligned} P_{rad} &= \frac{\varsigma_o k^2}{32\pi^2}\oint d\Omega\left|a_1 f_1 + a_2 f_2 e^{-i\mathbf{k}\cdot\mathbf{D}}\right|^2 \\ &= \frac{\varsigma_o k^2}{32\pi^2}\left[\begin{array}{l}\left(\oint d\Omega\left|f_1\right|^2\right)\left|a_1\right|^2 + \left(\oint d\Omega\left|f_2\right|^2\right)\left|a_2\right|^2 \\ +2\,\mathrm{Re}\left\{\oint d\Omega f_1^* \cdot f_2 e^{-i\mathbf{k}\cdot\mathbf{D}} a_1^* a_2\right\}\end{array}\right] \end{aligned} \tag{43}$$

where $f_1^* \cdot f_2 \equiv \mathbf{f}_1^* \cdot \mathbf{f}_2 - g_1^* \cdot g_2$. Thus, by comparing Eqs.(41) and (43), using Eq.(42),

$$\Lambda_{12} = \frac{\varsigma_o k^2}{64\pi^2}\frac{\mathrm{Re}\left\{\oint d\Omega f_1^* \cdot f_2 e^{-i\mathbf{k}\cdot\mathbf{D}} a_1^* a_2\right\}}{\mathrm{Re}\left\{a_1^* a_2\right\}} \tag{44}$$

namely Λ_{12} is exactly the interference term in AT. By substituting for the 4-vector current-moments and making the change of variables $\mathbf{r}_1 = \mathbf{r}_1'$, $\mathbf{r}_2 = \mathbf{r}_2' + \mathbf{D}$,

$$\begin{aligned} \Lambda_{12} &= \frac{\varsigma_o k^2}{64\pi^2}\frac{\mathrm{Re}\left\{\int_{V_1} dv \int_{V_2} dv J_1^* \cdot J_2 \oint d\Omega e^{-i\mathbf{k}\cdot\left(\mathbf{r_2}-\mathbf{r_1}\right)} a_1^* a_2\right\}}{\mathrm{Re}\left\{a_1^* a_2\right\}} \\ &= \frac{\varsigma_o k}{16\pi}\frac{\mathrm{Re}\left\{\int_{V_1} dv \int_{V_2} dv J_1^* \cdot J_2 \dfrac{\sin\left(k\left|\mathbf{r_2}-\mathbf{r_1}\right|\right)}{\left|\mathbf{r_2}-\mathbf{r_1}\right|} a_1^* a_2\right\}}{\mathrm{Re}\left\{a_1^* a_2\right\}} \end{aligned} \tag{45}$$

where we evaluated the integral over all angles of $\mathbf{k}$ with $\mathbf{r}_2 - \mathbf{r}_1$.

Note now that Eqs.(38) and (45) will become identical, if we can take the currents $J_{1,2}^{v}$ to be real. This is indeed the case for eigenmodes, where the field solution in bounded regions (such as those where the currents are flowing) is always stationary (in contrast to the leaky part of the eigenmode, which is radiative) and for high enough Q it can be chosen so that it is approximately real in the entire bounded region. Therefore, from either Eq.(38) or (45) we can write

$$\Lambda_{12} = \frac{\zeta_o k}{16\pi} \int_{V_1} dv \int_{V_2} dv J_1 \cdot J_2 \frac{\sin\left(k\left|\mathbf{r}_2 - \mathbf{r}_1\right|\right)}{\left|\mathbf{r}_2 - \mathbf{r}_1\right|} \tag{46}$$

and from Eq.(44), using Eq.(42), we can define the interference factor

$$V_{\text{rad},12} \equiv \frac{\Lambda_{12}}{\sqrt{\Gamma_{1,\text{rad}}\Gamma_{2,\text{rad}}}} = \frac{\oint d\Omega f_1^* \cdot f_2 e^{-i\mathbf{k}\cdot\mathbf{D}}}{\sqrt{\oint d\Omega \left|f_1\right|^2 \oint d\Omega \left|f_2\right|^2}} \tag{47}$$

We have shown that, in the high-Q limit, both PT and AT give the same expression for the imaginary part Λ_{nm} of the coupling coefficient, which thus physically describes within CMT the effects of far-field radiation interference. Again, this phenomenon was so far not considered to be predictable from CMT.

4 Efficiency enhancement and radiation suppression by far-field destructive interference

Physically, one can expect that far-field radiation interference can in principle be engineered to be destructive, resulting in reduced overall radiation losses for the two-object system and thus in enhanced system efficiency. In this section, we show that, indeed, in the presence of far-field interference, energy transfer can be more efficient and with less radiated power than what our previous model predicts.

Once more, we will treat the same temporal energy-transfer schemes as before (finite-amount and finite-rate), so that a direct comparison can be made.

4.1 Finite-amount energy-transfer efficiency

Considering again the source and device objects 1,2 to include the interference effects, the same CMT equations as in Eq.(1) can be used, but with the substitutions $\kappa_{\text{nm}} \rightarrow K_{\text{nm}} = \kappa_{\text{nm}} + i\Lambda_{\text{nm}};\ \text{n}, \text{m} = 1,2$. The real parts $\kappa_{11,22}$ can describe, as before, the shift in each object's resonance frequency due to the presence of the other; the imaginary parts $\Lambda_{11,22}$ can describe the change in the losses of each object due to the presence of the other (due to absorption in it or scattering from it, in which latter case losses could be either increased or decreased); both of these are second-order effects and, for the purposes of our mathematical analysis, can again be absorbed into the complex eigenfrequencies by setting $\omega_{1,2} \to \omega_{1,2} + \kappa_{11,22}$ and $\Gamma_{1,2} \to \Gamma_{1,2} - \Lambda_{11,22}$. The real parts $\kappa_{12,21}$ can denote, as before, the coupling coefficients; the imaginary parts $\Lambda_{12,21}$ can describe the far-field interference, as was shown in Section 3; again, from reciprocity $K_{12} = K_{21} \equiv K \equiv \kappa + i\Lambda$ (note that for a Hermitian problem, the additional requirement $K_{12} = K_{21}^*$ would impose K to be real, which makes sense, since without losses there cannot be any radiation interference).

Substituting $\kappa \to \kappa + i\Lambda$ into Eq.(2), we can find the normal modes of the system including interference effects. Note that, when the two objects are at exact resonance $\omega_1 = \omega_2 \equiv \omega_o$ and $\Gamma_1 = \Gamma_2 \equiv \Gamma_o$, the normal modes are found to be

$$\Omega_+ = (\omega_0 + \kappa) - i(\Gamma_0 - \Lambda) \text{ and } \Omega_- = (\omega_0 - \kappa) - i(\Gamma_0 + \Lambda), \tag{48}$$

which is exactly the typical case for respectively the odd and even normal modes of a system of two coupled objects, where for the even mode the objects' field-amplitudes have the same sign and thus the frequency is lowered and the radiative far-fields interfere constructively so loss is increased, while for the odd mode the situation is the opposite. This is another confirmation for the fact that the coefficient Λ can describe the far-field interference phenomenon under examination.

To treat now again the problem of energy transfer to object 2 from 1, but in the presence of radiative interference, again simply substitute $\kappa \to \kappa + i\Lambda$ into Eq.(3). Note that, at exact resonance $\omega_1 = \omega_2$ and, in the special case $\Gamma_1 = \Gamma_2 \equiv \Gamma_o$, we can just substitute into Eq.(4) $U \to U + iV$, where $U \equiv \kappa/\Gamma_o$ and $V \equiv \Lambda/\Gamma_o$, and then, with $T \equiv \Gamma_o t$, the evolution of the device field-amplitude becomes

$$\frac{a_2(T)}{|a_1(0)|} = \sin[(U + iV)T] \cdot e^{-T} \tag{49}$$

Now the efficiency $\eta_E \equiv |a_2(t)|^2/|a_1(0)|^2$ can be optimized for the normalized time T_* which is the solution of the transcendental equation

$$Re\{(U+iV)\cdot\cot[(U+iV)T_*]\}=1 \tag{50}$$

and the resulting optimal energy-transfer efficiency depends only on U, V and is depicted in Fig. 16(c), evidently increasing with V for a fixed U.

4.2 Finite-rate energy-transfer (power-transmission) efficiency

Similarly, to treat the problem of continuous powering of object 2 by 1, in the presence of radiative interference, simply substitute $U \rightarrow U + iV$ into the equations of Section 1.2, where $V \equiv \Lambda/\sqrt{\Gamma_1\Gamma_2}$ we call the strong-interference factor and quantifies the degree of far-field interference that the system experiences compared to loss. In practice, the parameters $D_{1,2}$, $U_{1,2}$ can be designed (engineered), since one can adjust the resonant frequencies $\omega_{1,2}$ (compared to the desired operating frequency ω) and the generator/load supply/drain rates $\kappa_{1,2}$. Their choice can target the optimization of some system performance-characteristic of interest.

In some examples, a goal can be to maximize the power transmission (efficiency) $\eta_P \equiv |S_{21}|^2$ of the system. The symmetry upon interchanging $1 \leftrightarrow 2$ is then preserved and, using Eq.(11), the field-amplitude transmission coefficient becomes

$$S_{21}=\frac{2i(U+iV)U_0}{(1+U_0-iD_0)^2+(U+iV)^2} \tag{51}$$

and from $\eta_P'(D_o)=0$ we get that, for fixed U, V and U_o, the efficiency can be maximized for the symmetric detuning

$$D_0=\begin{cases}2\sqrt{\alpha}\cos\left(\dfrac{\theta+2\nu\pi}{3}\right);\nu=0,1, & if\quad U^{2/3}-V^{2/3}>(1+U_0)^{2/3}\\ \sqrt[3]{\beta+\sqrt{\beta^2-\alpha^3}}+\sqrt[3]{\beta-\sqrt{\beta^2-\alpha^3}} & if\quad U^{2/3}-V^{2/3}\le(1+U_0)^{2/3}\end{cases} \tag{52}$$

where $\alpha \equiv [U^2-V^2-(1+U_o)^2]/3$, $\beta \equiv UV(1+U_o)$, $\theta \equiv \tan^{-1}\sqrt{\alpha^3/\beta^2-1}$ and $U^{2/3}-V^{2/3}>(1+U_o)^{2/3} \Leftrightarrow \alpha^3-\beta^2>0 \Leftrightarrow \alpha>0$. Note that, in the first case, the two peaks of the transmission curve are not equal for $V>0$, but the one at higher frequencies ($\nu=0 \Rightarrow$ positive detuning) corresponding to the odd system normal mode is higher, as should be expected, since the odd mode is the one that radiates less. Finally, by substituting D_o into η_P from Eq.(52), then from $\eta'_P(U_o) = 0$ we get that, for fixed U and V, the efficiency can be maximized for

$$U_{0*} = \sqrt{(1+U^2)(1-V^2)} \text{ and } D_{0*} = UV. \tag{53}$$

The dependence of the efficiency on D_o for different U_o (including the new 'critical-coupling' condition) are shown in Figs. 16(a,b). The overall optimal power efficiency using Eqs.(53) is

$$\eta_{P*} \equiv \eta_P(D_{0*}, U_{0*}) = \frac{U^2 + V^2}{(U_{0*} + 1)^2 + U^2 V^2} \tag{54}$$

which depends only on U, $|V|$ and is depicted in Figs. 16 (c,d), increasing with $|V|$ for a fixed U, and actually $\eta_P \to 1$ as $|V| \to 1$ for all values of U.

In some examples, a goal can be to minimize the power reflection at the side of the generator $|S_{11}|^2$ and the load $|S_{22}|^2$. The symmetry upon interchanging $1 \leftrightarrow 2$ is again preserved and, using then Eq.(17), one would require the 'impedance matching' condition

$$(1 - iD_0)^2 - U_0^2 + (U + iV)^2 = 0 \tag{55}$$

from which again we easily find that the values of D_o and U_o that cancel all reflections are exactly those in Eqs.(53).

In some examples, it can be of interest to minimize the power radiated from the system, since e.g. it can be a cause of interference to other communication systems, while still maintaining good efficiency. In some examples, the two objects can be the same, and then, using Eq.(41), we find

$$\eta_{rad} \equiv \frac{P_{rad}}{|S_{+1}|^2} = \frac{4U_0(|1 + U_0 - iD_0|^2 + |U + iV|^2)\frac{Q}{Q_{rad}} - 2V(V + VU_0 + UD_0)}{|(1 + U_0 - iD_0)^2 + (U + iV)^2|^2} \tag{56}$$

Then, to achieve our goal, we maximize η_P/η_{rad} and find that this can be achieved for

$$U_{0**} = \sqrt{1 + U^2 - V_{rad}^2 U^2 + V^2 - 2VV_{rad}} \text{ and } D_{0**} = UV_{rad},$$

(57)

where $V_{\text{rad}} \equiv \Lambda/\sqrt{\Gamma_{1,\text{rad}}\Gamma_{2,\text{rad}}}$, as defined in Eq.(47), we call the interference factor and quantifies the degree of far-field interference that the system experiences compared to the radiative loss, thus $V_{\text{rad}} = V\sqrt{\frac{Q_{1,\text{rad}}}{Q_1}\frac{Q_{2,\text{rad}}}{Q_2}} \geq V$, and $V = V_{\text{rad}}$ when all loss is radiative, in which case Eq.(57) reduces to Eq.(53).

In this description, we suggest that, for any temporal energy-transfer scheme and given some achieved coupling-to-loss ratio, the efficiency can be enhanced and the radiation can be suppressed by shifting the operational frequency away from exact resonance with each object's eigenfrequency and closer to the frequency of the odd normal-mode, which suffers less radiation due to destructive far-field interference. It is the parameters

$$U = \frac{\kappa}{\sqrt{\Gamma_1\Gamma_2}} = k\sqrt{Q_1Q_2} \qquad \text{and} \qquad V = \frac{\Lambda}{\sqrt{\Gamma_1\Gamma_2}} = V_{\text{rad}}\sqrt{\frac{Q_1}{Q_{1,\text{rad}}}\frac{Q_2}{Q_{2,\text{rad}}}} \tag{58}$$

that are the figures-of-merit for any system under consideration for wireless energy-transfer, along with the distance over which large$U$, $|V|$ can be achieved. Clearly, also $|V|$ can be a decreasing function of distance, since two sources of radiation distant by more than a few wavelengths do not interfere substantially. It is important also to keep in mind that the magnitude of V depends on the degree to which radiation dominates the objects' losses, since it is only these radiative losses that can contribute to interference, as expressed from $V_{\text{rad}} \geq V$.

To achieve a large strong-interference factor V, in some examples, the energy-transfer application preferably uses again subwavelength resonances, because, for a given source-device distance, the interference factor V_{rad} will increase as frequency decreases, since naturally the odd mode of two coupled objects, distant much closer than a wavelength, will not radiate at all.

To achieve a large strong-interference factor V, in some examples, the energy-transfer application preferably uses resonant modes of high factors Q/Q_{rad}. This condition can be satisfied by designing resonant modes where the dominant loss mechanism is radiation. As frequency decreases, radiation losses always decrease and typically systems are limited by absorption losses, as discussed earlier, so Q/Q_{rad} decreases; thus, the advantage of interference can be insignificant at some point compared to the deterioration of absorption-Q.

Therefore, $|V|$ will be maximized at some frequency ω_V, dependent on the source-device distance, and this optimal frequency will typically be different than ω_U, the optimal frequency for U. As seen above, the problem of maximizing the energy-transfer efficiency can require a modified treatment in the presence of interference. The choice of eigenfrequency for the source and device objects as ω_U, where U is maximum, can not be a good one anymore, but also V needs to be considered. The optimization of efficiency occurs then at a frequency ω_η between ω_U and ω_V and is a combined problem, which will be demonstrated below for few examples of electromagnetic systems.

Moreover, note that, at some fixed distance between the source and device objects, the figures U, V can not be maximized for the same set of system parameters; in that case, these parameters could be chosen so that the efficiency of Eq.(54) is maximized.

In the following section, we calculate a magnitude of efficiency improvement and radiation reduction for realistic systems at mid-range distances between two objects, by employing this frequency detuning and by doing a joint optimization for U, V.

5 Far-field interference at mid-range distances for realistic systems

In the case of two objects 1, 2 supporting radiative electromagnetic resonant modes of the same eigenfrequency $\omega_1 = \omega_2 \equiv \omega_o$ and placed at distance D between their arbitrarily-chosen centers, so that they couple in the near field and interfere in the far field, the interference factor V_{rad} is predicted from antenna theory (AT) to be that in Eq.(47).

We have also seen above how to compute the resonance quality factors Q and Q_{rad}, for some example structures, and thus we can compute the factor Q/Q_{rad}.

We will demonstrate the efficiency enhancement and the radiation suppression due to interference for the two examples of capacitively-loaded conducting loops and dielectric disks. The degree of improvement will be shown to depend on the nature of the system.

5.1 Capacitively-loaded conducting loops

Consider two loops 1, 2 of radius r with N turns of conducting wire with circular cross-section of radius a at distance D, as shown in Fig. 10. It was shown in Section 2.2 how to calculate the quality, coupling and strong-coupling factors for such a system.

Their coupling factor is shown in Fig. 17(a) as a function of the relative distance D/r, for three different dimensions of single-turn ($N = 1$) loops. Their strong-coupling factor at the eigenfrequency $\omega_{Q_1Q_2}$ is shown in Fig. 17(b). The approximate scaling $k, U \propto (r/D)^3$, indicated by Eqs.(26) and (27), is apparent.

We compute the interference parameter between two coupled loops at distance D, using the AT analysis Eq.(47), leading to

Consider two loops 1, 2 of radius r with N turns of conducting wire with circular cross-section of radius a at distance D, as shown in Fig. 10. It was shown in Section 2.2 how to calculate the quality, coupling and strong-coupling factors for such a system. Their coupling factor is shown in Fig. 17(a) as a function of the relative distance D/r, for three different dimensions of single-turn ($N = 1$) loops. Their strong-coupling factor at the eigenfrequency $\omega_{Q_1Q_2}$ is shown in Fig. 17(b). The approximate scaling $k, U \propto (r/D)^3$, indicated by Eqs.(26) and (27), is apparent. We compute the interference parameter between two coupled loops at distance D, using the AT analysis Eq.(47), leading to

$$V_{\text{rad}} = \frac{3}{(kD)^3}\left[\sin(kD) - (kD)\cos(kD)\right], \tag{59}$$

for the orientation of optimal coupling, where one loop is above the other. Their interference factor is shown in Fig. 18 as a function of the normalized distance D/λ, where it can be seen that this factor has nulls only upon reaching the radiative regime. Since the resonant loops are highly subwavelength (in many examples $\lambda/r \geq 50$), at mid-range distances ($D/r \leq 10$), we expect $D/\lambda \leq 0.2$ and thus the interference factor to be very large ($V_{\text{rad}} \geq 0.8$).

At a fixed resonant frequency, in some examples, the factor Q / Q_{rad} can be increased by increasing the radii r of the loops. In some examples, the factor Q / Q_{rad} can be increased by increasing the number N of turns of the loops. In some examples, the factor Q / Q_{rad} can be increased by increasing the radius a of the conducting wire of the loops or by using Litz wire or a ribbon to reduce the absorption losses and thus make radiation more dominant loss mechanism.

We also plot in Fig. 19, for the example $r = 30cm$ and $a = 2cm$, the strong-coupling factor U, the interference factor V_{rad} and the strong-interference factor V as a function of the resonant eigenfrequency of the loops, for a fixed distance $D = 5r$. Indeed, for this example, V_{rad} decreases monotonically with frequency in this subwavelength regime and is always great than 0.8, but V exhibits a maximum, since the term Q/Q_{rad} is increasing towards 1 with frequency, as losses become more and more radiation dominated. It can be seen that the resonant eigenfrequencies f_U and f_V, at which U and V become maximum respectively, are different. This implies that the efficiency will now not necessarily peak at the eigenfrequency f_U, at which U is maximized, as would be the assumption based on prior knowledge, but at a different one f_η between f_U and f_V. This is shown below.

In Fig. 20 the efficiency η_P is plotted as a function of the resonant eigenfrequency of the loops for two different examples of loop dimensions $r = 30cm$, $a = 2cm$ and $r = 1m$, $a = 2cm$, at two different loop distances $D = 5r$ and $D = 10r$, and for the cases:

(i) (solid lines) including interference effects and detuning the driving frequency from the resonant frequency by $D_o = UV$ from Eq.(53) to maximize the power-transmission efficiency and similarly using U_o from Eq.(53), which thus implies optimal efficiency as in Eq.(54).

(ii) (dash-dotted lines) including interference effects and detuning the driving frequency from the resonant frequency by $D_o = UV_{\text{rad}}$ from Eq.(57) to maximize the ratio of power transmitted over power radiated and similarly using U_o from Eq.(57).

(iii) (dashed lines) including interference effects but not detuning the driving frequency from the resonant frequency and using U_o from Eq.(14), as one would do to maximize efficiency in the absence of interference.

(iv) (dotted lines) truly in the absence of interference effects and thus maximizing efficiency by not detuning the driving frequency from the resonant frequency and using U_o from Eq.(14), which thus implies efficiency as in Eq.(15).

In Fig. 21 we show the amount of driving-frequency detuning that is used in the presence of interference either to maximize efficiency (case (i) (solid lines) of Fig. 20 - $D_o = UV$) or to maximize the ratio of power transmitted over power radiated (case (ii) (dash-dotted lines) of Fig. 20 - $D_o = UV_{\text{rad}}$). Clearly, this driving-frequency detuning can be a non-trivial amount.

It can be seen from Fig. 20 that, for all frequencies, the efficiency of case (i) (solid lines) is larger than the efficiency of case (iii) (dashed lines) which is in turn larger than the efficiency of case (iv) (dotted lines). Therefore, in this description, we suggest that employing far-field interference improves on the power-transmission efficiency (improvement from (iv) (dotted) to (iii) (dashed)) and, furthermore, employing destructive far-field interference, by detuning the driving frequency towards the low-radiation-loss odd normal mode, improves on the power-transmission efficiency even more (improvement from (iii) (dashed) to (i) (solid)).

If f_η is the eigenfrequency, at which the efficiency of case (i) (solid) is optimized, then, in some examples, the resonant eigenfrequency can be designed to be larger than f_η, namely in a regime where the system is more radiation dominated. In this description, we suggest that at such eigenfrequencies, there can be a significant improvement in efficiency by utilizing the destructive far-field interference effects and driving the system at a frequency close to the odd normal mode. This can be seen again from Fig. 20 by comparing the solid lines to the corresponding dashed lines and the dotted lines.

In general, one would tend to design a system resonant at the frequency f_U where the strong-coupling factor U is maximal. However, as suggested above, in the presence of interference, Fig. 20 shows that the maximum of η_P is at an eigenfrequency f_η different than f_U. In some examples, $f_\eta > f_U$. This is because at higher eigenfrequencies, losses are determined more by radiation than absorption, therefore destructive radiation interference can play a more significant role in reducing overall losses and thus $f_V > f_U$ and the efficiency in increased at $f_\eta > f_U$. In this description, we suggest that, in some examples, the resonant eigenfrequency can be designed to be close to the frequency f_η that optimizes the efficiency rather than the different f_U. In particular, in Fig. 22(a) are plotted these two frequencies f_η (solid line) and f_U (dashed line) as a function of relative distance D/r of two $r = 30cm$ loops. In Fig. 22(b) we show a graded plot of the optimal efficiency from Eq.(54) in the $U - V$ plane. Then, we superimpose the $U - V$ curve of case (i) (solid), parametrized with distance D, for two $r = 30cm$ loops resonant at the optimal frequency f_η for each D. From the path of this curve onto the graded plot the efficiency as a function of distance can be extracted for case (i) (solid). We then also superimpose in Fig. 22(b) the $U - V$ curve of case (iii) (dashed) , parametrized with distance D, for two $r = 30cm$ loops resonant at f_U, and the U range of case (iv) (dotted), parametrized with distance D, for two $r = 30cm$ loops resonant at f_U (note that in this last case there is no interference and thus $V = 0$). In Fig. 22(c) we then show the efficiency enhancement factor achieved by the solid curve of Fig. 22(b), as a function of distance D/r, compared to best that can be achieved without driving-frequency detuning (dashed) and without interference whatsoever (dotted). The improvement by employing interference can reach a factor of 2 at large separation between the loops.

In Fig. 23 we plot the radiation efficiency η_{rad}, using Eq.(39), as a function of the eigenfrequency of the loops for the two different loop dimensions, the two different distances and the four different cases examined in Fig. 20. It can be seen from Fig. 23 that, for all frequencies, η_{rad} of case (ii) (dash-dotted lines) is smaller than η_{rad} of case (i) (solid lines) which is in turn smaller than η_{rad} of case (iii) (dashed lines) and this smaller than η_{rad} of case (iv) (dotted lines). Therefore, in this description, we suggest that employing far-field interference suppresses radiation (improvement from (iv) (dotted) to (iii) (dashed)) and, furthermore, employing destructive far-field interference, by detuning the driving frequency towards the low-radiation-loss odd normal mode, suppress radiation efficiency even more (improvement from (iii) (dashed) to (i) (solid) and (ii) (dash-dotted)), more so in case (ii), specifically optimized for this purpose.

In some examples, the resonant eigenfrequency can be designed to be larger than f_η, namely in a regime where the system is more radiation dominated. In this description, we suggest that at such eigenfrequencies, there can be a significant suppression in radiation by utilizing the destructive far-field interference effects and driving the system at a frequency close to the odd normal mode. The case (ii)=(dash-dotted) accomplishes the greatest suppression in radiation and, as can be seen in Fig. 20, there is a range of eigenfrequencies (close to f_V), for which the efficiency that this configuration can achieve is only little compromised compared to the maximum possible of configuration (i).

In one example, two single-turn loops of $r = 30cm$ and $a = 2cm$ are at a distance $D/r = 5$ in the orientation shown in Fig. 10 and they are designed to resonate at 30MHz. In the absence of interference, the power-transmission efficiency is 59% and the radiation efficiency is 38%. In the presence of interference and without detuning the driving frequency from 30MHz, the power-transmission efficiency is 62% and the radiation efficiency is 32%. In the presence of interference and detuning the driving frequency from 30MHz to 31.3MHz to maximize efficiency, the power-transmission efficiency is increased to 75% and the radiation efficiency is suppressed to 18%.

In another example, two single-turn loops of $r = 30cm$ and $a = 2cm$ are at a distance $D/r = 5$ in the orientation shown in Fig. 10 and they are designed to resonate at 10MHz. In the absence of interference or in the presence of interference and without detuning the driving frequency from 10MHz, the power-transmission efficiency is approximately 81% and the radiation efficiency is approximately 4%. In the presence of interference and detuning the driving frequency from 10MHz to 10.22MHz to maximize transmission over radiation, the power-transmission efficiency is 42%, reduced by less than a factor of 2, while the radiation efficiency is 0.4%, suppressed by an order of magnitude.

In another example, two single-turn loops of $r = 1m$ and $a = 2cm$ are at a distance $D/r = 5$ in the orientation shown in Fig. 10 and they are designed to resonate at 10MHz. In the absence of interference, the power-transmission efficiency is 48% and the radiation efficiency is 47%. In the presence of interference and without detuning the driving frequency from 10MHz, the power-transmission efficiency is 54% and the radiation efficiency is 37%. In the presence of interference and detuning the driving frequency from 10MHz to 14.8MHz to maximize efficiency, the power-transmission efficiency is increased to 66% and the radiation efficiency is suppressed to 24%.

In another example, two single-turn loops of $r = 1m$ and $a = 2cm$ are at a distance $D/r = 5$ in the orientation shown in Fig. 10 and they are designed to resonate at 4MHz. In the absence of interference or in the presence of interference and without detuning the driving frequency from 4MHz, the power-transmission efficiency is approximately 71% and the radiation efficiency is approximately 8%. In the presence of interference and detuning the driving frequency from 4MHz to 5.06MHz to maximize transmission over radiation, the power-transmission efficiency is 40%, reduced by less than a factor of 2, while the radiation efficiency is approximately 1%, suppressed by almost an order of magnitude.

5.2 Dielectric disks

Consider two dielectric disks 1 and 2 of radius *r* and dielectric permittivity ϵ placed at distance *D* between their centers, as shown in Fig. 15(b). Their coupling as a function of distance was calculated in Section 2.4, using analytical and finite-element-frequency-domain (FEFD) methods, and is shown in Fig. 24.

To compute the interference factor between two coupled disks at distance *D*, we again use two independent methods to confirm the validity of our results: numerically, the FEFD calculations again give Λ (and thus V) by the splitting of the loss-rates of the two normal modes; analytically, calculation of the AT prediction of Eq.(47) gives

$$
\begin{aligned}
m=1: &\quad V_{\text{rad}} = \frac{2}{(kD)} J_1(kD) \\
m=2: &\quad V_{\text{rad}} = \frac{8}{(kD)^3}\left\{3(kD)J_0(kD)+\left[(kD)^2-6\right]J_1(kD)\right\} \\
m=3: &\quad V_{\text{rad}} = \frac{6}{(kD)^5}\left\{\left[24(kD)^3-320(kD)\right]J_0(kD)+\left[3(kD)^4-128(kD)^2+640\right]J_1(kD)\right\}
\end{aligned}
\tag{60}
$$

The results for the interference of two same disks, for exactly the same parameters for which the coupling was calculated in Fig. 24, are presented in Fig. 25, as a function of frequency (due to varying ϵ) at fixed distances. It can be seen that also the strong-interference factor [illegible] can have nulls, which can occur even before the system enters the radiative-coupling regime, namely at smaller frequencies than those of U at the same distances, and it decreases with frequency, since then the objects become more and more absorption dominated, so the benefit from radiative interference is suppressed. Both the above effects result into the fact that, for most distances, U (from Fig. 24(b)) and V (from Fig. 25(b)) can be maximized at different values of the frequency (f_U and f_V respectively), and thus different can also be the optimal frequency f_η for the final energy-transfer efficiency of Eq.(54), which is shown in Fig. 26 again for the same set of parameters. From this plot, it can be seen that interference can significantly improve the transfer efficiency, compared to what Eq.(15) would predict from the calculated values of the coupling figure-of-merit U.

Furthermore, not only does a given energy-transfer system perform better than what a prediction which ignores interference would predict, but also our optimization design will typically lead to different optimal set of parameters in the presence of interference. For example, for the particular distance $D/r = 5$, it turns out from Fig. 26 that the $m = 1$ resonant modes can achieve better efficiency than the $m = 2$ modes within the available range of ϵ, by making use of strong interference which counteracts their weaker U, as viewed in Fig. 24, from which one would have concluded the opposite performance. Moreover, even within the same m-branch, one would naively design the system to operate at the frequencyf_U, at which U is maximum. However, the optimization design changes in the presence of interference, since the system should be designed to operate at the different frequency f_η, where the overall efficiency η peaks. In Fig. 27(a), we first calculate those different frequencies where the strong-coupling factor U and the efficiency η (which includes interference) peak, as distance D is changing for the choice of the $m = 2$ disk of Fig. 24, and observe that their difference is actually significant. Then, in Fig. 27(b) we show the peak efficiency for the various frequency choices. For large distances, where efficiency is small and could use a boost, the improvement factor reaches a significant 2 for the particular system under examination. The same result is shown in Fig. 27(c) as a plot of the path of the efficiency on the U - V map, as distance is changing. Similar results are derived for the modes of different m-order. Physically, moving to higher frequencies increases role of radiative losses compared to absorption and thus interference can have a greater influence. At the optimal frequency f_η radiated power including interference is close to what it is at f_U, but absorbed power is much less, therefore the efficiency has been improved.

In some examples, instead of improving efficiency, one might care more about minimizing radiation. In that case, we calculate at the frequency f_U how much power is radiated when optimized under the conditions Eq.(57) compared to the power radiated when simply operating on resonance ($D_\circ = 0$) in the cases with and without interference (the latter case can be describing a case where the two disks do not interfere, because they are dissimilar, or due to decoherence issues etc.). We find in Fig. 28 that radiation can be suppressed by a factor of 1.6 by detuning the operating frequency towards the odd sub-radiant mode.

6 System Sensitivity to Extraneous Objects

In general, the overall performance of an example of the resonance-based wireless energy-transfer scheme depends strongly on the robustness of the resonant objects' resonances. Therefore, it is desirable to analyze the resonant objects' sensitivity to the near presence of random non-resonant extraneous objects. One appropriate analytical model is that of "perturbation theory" (PT), which suggests that in the presence of an extraneous perturbing object p the field amplitude $a_1(t)$ inside the resonant object 1 satisfies, to first order:

$$\frac{da_1}{dt} = -i\left(\omega_1 - i\Gamma_1\right)a_1 + i\left(\delta\kappa_{11(p)} + i\delta\Gamma_{1(p)}\right)a_1 \tag{61}$$

where again ω_1 is the frequency and Γ_1 the intrinsic (absorption, radiation etc.) loss rate, while $\delta\kappa_{11(p)}$ is the frequency shift induced onto 1 due to the presence of p and $\delta\Gamma_{1(p)}$ is the extrinsic due to p (absorption inside p, scattering from p etc.) loss rate. $\delta\Gamma_{1(p)}$ is defined as $\delta\Gamma_{1(p)} \equiv \Gamma_{1(p)} - \Gamma_1$, where $\Gamma_{1(p)}$ is the total perturbed loss rate in the presence of p. The first-order PT model is valid only for small perturbations. Nevertheless, the parameters $\delta\kappa_{11(p)}$, $\delta\Gamma_{1(p)}$ are well defined, even outside that regime, if a_1 is taken to be the amplitude of the exact perturbed mode. Note also that interference effects between the radiation field of the initial resonant-object mode and the field scattered off the extraneous object can for strong scattering (e.g. off metallic objects) result in total $\Gamma_{1,\text{rad}(p)}$ that are smaller than the initial $\Gamma_{1,\text{rad}}$ (namely $\delta\Gamma_{1,\text{rad}(p)}$ is negative).

It has been shown that a specific relation is desired between the resonant frequencies of the source and device-objects and the driving frequency. In some examples, all resonant objects must have the same eigenfrequency and this must be equal to the driving frequency. In some examples, when trying to optimize efficiency or suppress radiation by employing far-field interference, all resonant objects must have the same eigenfrequency and the driving frequency must be detuned from them by a particular amount. In some implementations, this frequency-shift can be "fixed" by applying to one or more resonant objects and the driving generator a feedback mechanism that corrects their frequencies. In some examples, the driving frequency from the generator can be fixed and only the resonant frequencies of the objects can be tuned with respect to this driving frequency.

The resonant frequency of an object can be tuned by, for example, adjusting the geometric properties of the object (e.g. the height of a self-resonant coil, the capacitor plate spacing of a capacitively-loaded loop or coil, the dimensions of the inductor of an inductively-loaded rod, the shape of a dielectric disc, etc.) or changing the position of a non-resonant object in the vicinity of the resonant object.

In some examples, referring to Fig. 29a, each resonant object is provided with an oscillator at fixed frequency and a monitor which determines the eigenfrequency of the object. At least one of the oscillator and the monitor is coupled to a frequency adjuster which can adjust the frequency of the resonant object. The frequency adjuster determines the difference between the fixed driving frequency and the object frequency and acts, as described above, to bring the object frequency into the required relation with respect to the fixed frequency. This technique assures that all resonant objects operate at the same fixed frequency, even in the presence of extraneous objects.

In some examples, referring to Fig. 29(b), during energy transfer from a source object to a device object, the device object provides energy or power to a load, and an efficiency monitor measures the efficiency of the energy-transfer or power-transmission. A frequency adjuster coupled to the load and the efficiency monitor acts, as described above, to adjust the frequency of the object to maximize the efficiency.

In other examples, the frequency adjusting scheme can rely on information exchange between the resonant objects. For example, the frequency of a source object can be monitored and transmitted to a device object, which is in turn synched to this frequency using frequency adjusters, as described above. In other embodiments the frequency of a single clock can be transmitted to multiple devices, and each device then synched to that frequency using frequency adjusters, as described above.

Unlike the frequency shift, the extrinsic perturbing loss due to the presence of extraneous perturbing objects can be detrimental to the functionality of the energy-transfer scheme, because it is difficult to remedy. Therefore, the total perturbed quality factors $Q_{(p)}$ (and the corresponding perturbed strong-coupling factor $U_{(p)}$ and the perturbed strong-interference factor $V_{(p)}$) should be quantified.

In some examples, a system for wireless energy-transfer uses primarily magnetic resonances, wherein the energy stored in the near field in the air region surrounding the resonator is predominantly magnetic, while the electric energy is stored primarily inside the resonator. Such resonances can exist in the quasi-static regime of operation ($r \ll \lambda$) that we are considering: for example, for coils with $h \ll 2r$, most of the electric field is localized within the self-capacitance of the coil or the externally loading capacitor and, for dielectric disks, with $\epsilon \gg 1$ the electric field is preferentially localized inside the disk. In some examples, the influence of extraneous objects on magnetic resonances is nearly absent. The reason is that extraneous non-conducting objects p that could interact with the magnetic field in the air region surrounding the resonator and act as a perturbation to the resonance are those having significant magnetic properties (magnetic permeability $Re\{\mu\}>1$ or magnetic loss $Im\{\mu\}>0$). Since almost all every-day non-conducting materials are non-magnetic but just dielectric, they respond to magnetic fields in the same way as free space, and thus will not disturb the resonance of the resonator. Extraneous conducting materials can however lead to some extrinsic losses due to the eddy currents induced inside them or on their surface (depending on their conductivity). However, even for such conducting materials, their presence will not be detrimental to the resonances, as long as they are not in very close proximity to the resonant objects.

The interaction between extraneous objects and resonant objects is reciprocal, namely, if an extraneous object does not influence a resonant object, then also the resonant object does not influence the extraneous object. This fact can be viewed in light of safety considerations for human beings. Humans are also non-magnetic and can sustain strong magnetic fields without undergoing any risk. A typical example, where magnetic fields $B{\sim}1T$ are safely used on humans, is the Magnetic Resonance Imaging (MRI) technique for medical testing. In contrast, the magnetic near-field required in typical embodiments in order to provide a few Watts of power to devices is only $B{\sim}10^{-4}T$, which is actually comparable to the magnitude of the Earth's magnetic field. Since, as explained above, a strong electric near-field is also not present and the radiation produced from this non-radiative scheme is minimal, the energy-transfer apparatus, methods and systems described herein is believed safe for living organisms.

6.1 Capacitively-loaded conducting loops or coils

In some examples, one can estimate the degree to which the resonant system of a capacitively-loaded conducting-wire coil has mostly magnetic energy stored in the space surrounding it. If one ignores the fringing electric field from the capacitor, the electric and magnetic energy densities in the space surrounding the coil come just from the electric and magnetic field produced by the current in the wire; note that in the far field, these two energy densities must be equal, as is always the case for radiative fields. By using the results for the fields produced by a subwavelength ($r \ll \lambda$) current loop (magnetic dipole) with $h=0$, we can calculate the ratio of electric to magnetic energy densities, as a function of distance D_p from the center of the loop (in the limit $r \ll D_p$) and the angle θ with respect to the loop axis:

$$\frac{w_e(x)}{w_m(x)} = \frac{\varepsilon_o |E(x)|^2}{\mu_o |H(x)|^2} = \frac{\left(1+\frac{1}{x^2}\right)\sin^2\theta}{\left(\frac{1}{x^2}+\frac{1}{x^4}\right)4\cos^2\theta + \left(1-\frac{1}{x^2}+\frac{1}{x^4}\right)\sin^2\theta};\ x = 2\pi\frac{D_p}{\lambda}$$
$$\Rightarrow \frac{\oiint_{S_p} w_e(x)\,dS}{\oiint_{S_p} w_m(x)\,dS} = \frac{1+\frac{1}{x^2}}{1+\frac{1}{x^2}+\frac{3}{x^4}};\ x = 2\pi\frac{D_p}{\lambda}, \qquad (62)$$

where the second line is the ratio of averages over all angles by integrating the electric and magnetic energy densities over the surface of a sphere of radius D_p. From Eq.(62) it is obvious that indeed for all angles in the near field ($x \ll 1$) the magnetic energy density is dominant, while in the far field ($x \gg 1$) they are equal as they should be. Also, the preferred positioning of the loop is such that objects which can interfere with its resonance lie close to its axis ($\theta = 0$), where there is no electric field. For example, using the systems described in Table 4, we can estimate from Eq.(62) that for the loop of $r = 30cm$ at a distance $D_p = 10r = 3m$ the ratio of average electric to average magnetic energy density would be $\sim 12\%$ and at $D_p = 3r = 90cm$ it would be $\sim 1\%$, and for the loop of $r = 10cm$ at a distance $D_p = 10r = 1m$ the ratio would be $\sim 33\%$ and at $D_p = 3r = 30cm$ it would be $\sim 2.5\%$. At closer distances this ratio is even smaller and thus the energy is predominantly magnetic in the near field, while in the radiative far field, where they are necessarily of the same order (ratio $\to 1$), both are very small, because the fields have significantly decayed, as capacitively-loaded coil systems are designed to radiate very little. Therefore, this is the criterion that qualifies this class of resonant system as a magnetic resonant system.

To provide an estimate of the effect of extraneous objects on the resonance of a capacitively-loaded loop including the capacitor fringing electric field, we use the perturbation theory formula, stated earlier, $\delta\Gamma_{1,abs(p)} = \omega_1 / 4 \cdot \int d^3\mathbf{r} \operatorname{Im}\left\{\varepsilon_p\left(\mathbf{r}\right)\right\} \left|\mathbf{E}_1\left(\mathbf{r}\right)\right|^2 \Big/ W$ with the computational FEFD results for the field of an example like the one shown in the plot of Fig. 5 and with a rectangular object of dimensions *30cm x 30cm x 1.5m* and permittivity *ε=49+16i* (consistent with human muscles) residing between the loops and almost standing on top of one capacitor (*~3cm* away from it) and find $\delta Q_{abs(human)} \sim 10^5$ and for *~10cm* away $\delta Q_{abs(human)} \sim 5 \cdot 10^5$. Thus, for ordinary distances (*~1m*) and placements (not immediately on top of the capacitor) or for most ordinary extraneous objects p of much smaller loss-tangent, we conclude that it is indeed fair to say that $\delta Q_{abs(p)} \to \infty$. The only perturbation that is expected to affect these resonances is a close proximity of large metallic structures.

Self-resonant coils can be more sensitive than capacitively-loaded coils, since for the former the electric field extends over a much larger region in space (the entire coil) rather than for the latter (just inside the capacitor). On the other hand, self-resonant coils can be simple to make and can withstand much larger voltages than most lumped capacitors. Inductively-loaded conducting rods can also be more sensitive than capacitively-loaded coils, since they rely on the electric field to achieve the coupling.

6.2 Dielectric disks

For dielectric disks, small, low-index, low-material-loss or far-away stray objects will induce small scattering and absorption. In such cases of small perturbations these extrinsic loss mechanisms can be quantified using respectively the analytical first-order perturbation theory formulas

$$\left[\delta Q_{1,rad(p)}\right]^{-1} \equiv 2\delta\Gamma_{1,rad(p)} / \omega_1 \propto \int d^3\mathbf{r}\left[\mathrm{Re}\left\{\varepsilon_p(\mathbf{r})\right\}\left|\mathbf{E}_1(\mathbf{r})\right|\right]^2 \Big/ W$$

$$\left[\delta Q_{1,abs(p)}\right]^{-1} \equiv 2\delta\Gamma_{1,abs(p)} / \omega_1 = \int d^3\mathbf{r}\,\mathrm{Im}\left\{\varepsilon_p(\mathbf{r})\right\}\left|\mathbf{E}_1(\mathbf{r})\right|^2 \Big/ 2W$$

where $W = \int d^3\mathbf{r}\varepsilon(\mathbf{r})\left|\mathbf{E}_1(\mathbf{r})\right|^2 / 2$ is the total resonant electromagnetic energy of the unperturbed mode. As one can see, both of these losses depend on the square of the resonant electric field tails **E**1 at the site of the extraneous object. In contrast, the coupling factor from object 1 to another resonant object 2 is, as stated earlier,

$$k_{12} = 2\kappa_{12} / \sqrt{\omega_1\omega_2} \approx \int d^3\mathrm{r}\varepsilon_2(\mathrm{r})\mathrm{E}_2^*(\mathrm{r})\mathrm{E}_1(\mathrm{r}) / \int d^3\mathrm{r}\varepsilon(\mathrm{r})\left|\mathrm{E}_1(\mathrm{r})\right|^2$$

and depends *linearly* on the field tails $\mathbf{E}_1$ of 1 inside 2. This difference in scaling gives us confidence that, for, for example, exponentially small field tails, coupling to other resonant objects should be much faster than all extrinsic loss rates ($\kappa_{12} \gg \delta\Gamma_{1,2(p)}$), at least for small perturbations, and thus the energy-transfer scheme is expected to be sturdy for this class of resonant dielectric disks.

However, we also want to examine certain possible situations where extraneous objects cause perturbations too strong to analyze using the above first-order perturbation theory approach. For example, we place a dielectric disk close to another off-resonance object of large $\mathrm{Re}\{\varepsilon\}$, $\mathrm{Im}\{\varepsilon\}$ and of same size but different shape (such as a human being *h*), as shown in Fig. 30a, and a roughened surface of large extent but of small $\mathrm{Re}\{\varepsilon\}$,

$\mathrm{Im}\{\varepsilon\}$ (such as a wall w), as shown in Fig. 30b. For distances $D_{h,w}/r = 10-3$ between the disk-center and the "human"-center or "wall", the numerical FDFD simulation results presented in Figs. 30a and 30b suggest that, the disk resonance seems to be fairly robust, since it is not detrimentally disturbed by the presence of extraneous objects, with the exception of the *very* close proximity of high-loss objects. To examine the influence of large perturbations on an entire energy-transfer system we consider two resonant disks in the close presence of both a "human" and a "wall". Comparing Table 8 to the table in Figure 30c, the numerical FDFD simulations show that the system performance deteriorates from $U{\sim}1-50$ to $U_{(hw)}{\sim}0.5-10$, i.e. only by acceptably small amounts.

In general, different examples of resonant systems have different degree of sensitivity to external perturbations, and the resonant system of choice depends on the particular application at hand, and how important matters of sensitivity or safety are for that application. For example, for a medical implantable device (such as a wirelessly powered artificial heart) the electric field extent must be minimized to the highest degree possible to protect the tissue surrounding the device. In such cases where sensitivity to external objects or safety is important, one should design the resonant systems so that the ratio of electric to magnetic energy density w_e/w_m is reduced or minimized at most of the desired (according to the application) points in the surrounding space.

7 Applications

The non-radiative wireless energy transfer techniques described above can enable efficient wireless energy-exchange between resonant objects, while suffering only modest transfer and dissipation of energy into other extraneous off-resonant objects. The technique is general, and can be applied to a variety of resonant systems in nature. In this Section, we identify a variety of applications that can benefit from or be designed to utilize wireless power transmission.

Remote devices can be powered directly, using the wirelessly supplied power or energy to operate or run the devices, or the devices can be powered by or through or in addition to a battery or energy storage unit, where the battery is occasionally being charged or re-charged wirelessly. The devices can be powered by hybrid battery/energy storage devices such as batteries with integrated storage capacitors and the like.

Furthermore, novel battery and energy storage devices can be designed to take advantage of the operational improvements enabled by wireless power transmission systems.

Devices can be turned off and the wirelessly supplied power or energy used to charge or recharge a battery or energy storage unit. The battery or energy storage unit charging or recharging rate can be high or low. The battery or energy storage units can be trickle charged or float charged. It would be understood by one of ordinary skill in the art that there are a variety of ways to power and/or charge devices, and the variety of ways could be applied to the list of applications that follows.

Some wireless energy transfer examples that can have a variety of possible applications include for example, placing a source (e.g. one connected to the wired electricity network) on the ceiling of a room, while devices such as robots, vehicles, computers, PDAs or similar are placed or move freely within the room. Other applications can include powering or recharging electric-engine buses and/or hybrid cars and medical implantable devices. Additional example applications include the ability to power or recharge autonomous electronics (e.g. laptops, cell-phones, portable music players, house-hold robots, GPS navigation systems, displays, etc), sensors, industrial and manufacturing equipment, medical devices and monitors, home appliances (e.g. lights, fans, heaters, displays, televisions, counter-top appliances, etc.), military devices, heated or illuminated clothing, communications and navigation equipment, including equipment built into vehicles, clothing and protective-wear such as helmets, body armor and vests, and the like, and the ability to transmit power to physically isolated devices such as to implanted medical devices, to hidden, buried, implanted or embedded sensors or tags, to and/or from roof-top solar panels to indoor distribution panels, and the like.

In some examples, far-field interference can be utilized by a system designer to suppress total radiation loss and/or to increase the system efficiency. In some examples, systems operating optimally closer to the radiative regime can benefit more from the presence of far-field interference, which leads to reduced losses for the sub-radiant normal mode of the coupled objects, and this benefit can be substantial.

A number of examples of the invention have been described. Nevertheless, it will be understood that various modifications can be made without departing from the spirit and scope of the invention.

WHAT IS CLAIMED IS:

1. An apparatus for use in wireless energy transfer, the apparatus comprising:

a first resonator structure configured for energy transfer with a second resonator structure, over a distance D larger than a characteristic size L_1 of said first resonator structure and larger than a characteristic size L_2 of said second resonator structure,

wherein the energy transfer has a rate κ and is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, wherein

said resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$ at least larger than 300, and

said resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$ at least larger than 300,

wherein the absolute value of the difference of said angular frequencies ω_1 and ω_2 is smaller than the broader of said resonant widths Γ_1 and Γ_2, and the quantity $\kappa / \sqrt{\Gamma_1 \Gamma_2}$ is at least larger than 20,

and further comprising a power supply coupled to the first structure and configured to drive the first resonator structure or the second resonator structure at an angular frequency away from the resonance angular frequencies and shifted towards a frequency corresponding to an odd normal mode for the resonator structures to reduce radiation from the resonator structures by destructive far-field interference.

2. The apparatus of claim 1, wherein the power supply is configured to drive the first resonator structure or the second resonator structure at the angular frequency away from the resonance angular frequencies and shifted towards the frequency corresponding to an odd normal mode for the resonator structures to substantially suppress radiation from the resonator structures by destructive far-field interference.

3. A method for wireless energy transfer involving a first resonator structure configured for energy transfer with a second resonator structure, over a distance D larger than a characteristic size L_1 of said first resonator structure and larger than a characteristic size L_2 of said second resonator structure, wherein the energy transfer has a rate κ and is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, wherein said resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$ at least larger than 300, and said resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$ at least larger than 300, wherein the absolute value of the difference of said angular frequencies ω_1 and ω_2 is smaller than the broader of said resonant widths Γ_1 and Γ_2, and the quantity $\kappa / \sqrt{\Gamma_1 \Gamma_2}$ is at least larger than 20, the method comprising:

driving the first resonator structure or the second resonator structure at an angular frequency away from the resonance angular frequencies and shifted towards a frequency corresponding to an odd normal mode for the resonator structures to reduce radiation from the resonator structures by destructive far-field interference.

4. The method of claim 3, wherein the first resonator structure or the second resonator structure is driven at the angular frequency away from the resonance angular frequencies and shifted towards the frequency corresponding to an odd normal mode for the resonator structures to substantially suppress radiation from the resonator structures by destructive far-field interference.

5. An apparatus for use in wireless energy transfer, the apparatus comprising:

a first resonator structure configured for energy transfer with a second resonator structure, over a distance D larger than a characteristic size L_1 of said first resonator structure and larger than a characteristic size L_2 of said second resonator structure,

wherein the energy transfer has a rate κ and is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, wherein

said resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$ at least larger than 300, and

said resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$ at least larger than 300,

wherein the absolute value of the difference of said angular frequencies ω_1 and ω_2 is smaller than the broader of said resonant widths Γ_1 and Γ_2, and the quantity $\kappa / \sqrt{\Gamma_1 \Gamma_2}$ is at least larger than 20,

and wherein for a desired range of the distances *D*, the resonance angular frequencies for the resonator structures increase transmission efficiency *T* by accounting for radiative interference, wherein the increase is relative to a transmission efficiency *T* calculated without accounting for the radiative interference.

6. The apparatus of claim 5, wherein the resonance angular frequencies for the resonator structures are selected by optimizing the transmission efficiency T to account for both a resonance quality factor *U* and an interference factor *V*.

7. A method for designing a wireless energy transfer apparatus, the apparatus including a first resonator structure configured for energy transfer with a second resonator structure, over a distance *D* larger than a characteristic size L_1 of said first resonator structure and larger than a characteristic size L_2 of said second resonator structure, wherein the energy transfer has a rate κ and is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, wherein said resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$ at least larger than 300, and said resonant field of the second

resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$ at least larger than 300, wherein the absolute value of the difference of said angular frequencies ω_1 and ω_2 is smaller than the broader of said resonant widths Γ_1 and Γ_2, and the quantity $\kappa / \sqrt{\Gamma_1 \Gamma_2}$ is at least larger than 20, the method comprising:

selecting the resonance angular frequencies for the resonator structures to substantially optimize the transmission efficiency by accounting for radiative interference between the resonator structures.

8. The method of claim 7, wherein the resonance angular frequencies for the resonator structures are selected by optimizing the transmission efficiency T to account for both a resonance quality factor *U* and an interference factor *V*.

9. An apparatus for use in wireless energy transfer, the apparatus comprising:

a first resonator structure configured for energy transfer with a second resonator structure over a distance *D*,

wherein the energy transfer is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, with a coupling factor *k*, wherein

said resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$, and is radiative in the far field, with an associated radiation quality factor $Q_{1,\mathrm{rad}} \geq Q_1$, and

said resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$, and is radiative in the far field, with an associated radiation quality factor $Q_{2,\mathrm{rad}} \geq Q_2$,

wherein an absolute value of a difference of said angular frequencies ω_1 and ω_2

is smaller than broader of said resonant widths Γ_1 and Γ_2, and an average resonant angular frequency is defined as $\omega_o = \sqrt{\omega_1 \omega_2}$, corresponding to an average resonant wavelength $\lambda_o = 2\pi c / \omega_o$, where c is the speed of light in free space, and a strong-coupling factor being defined as $U = k\sqrt{Q_1 Q_2}$,

wherein the apparatus is configured to employ interference between said radiative far fields of the resonant fields of the first and second resonator, with an interference factor V_{rad}, to reduce a total amount of radiation from the apparatus compared to an amount of radiation from the apparatus in the absence of interference, a strong-interference factor being defined as $V = V_{\text{rad}}\sqrt{(Q_1 / Q_{1,\text{rad}})(Q_2 / Q_{2,\text{rad}})}$.

10. The apparatus of claim 9, wherein $Q_1 / Q_{1,\text{rad}} \geq 0.01$ and $Q_2 / Q_{2,\text{rad}} \geq 0.01$.

11. The apparatus of claim 9, wherein $Q_1 / Q_{1,\text{rad}} \geq 0.1$ and $Q_2 / Q_{2,\text{rad}} \geq 0.1$.

12. The apparatus of claim 9, wherein D / λ_o is larger than 0.001 and the strong-interference factor V is larger than 0.01.

13. The apparatus of claim 9, wherein D / λ_o is larger than 0.001 and the strong-interference factor V is larger than 0.1.

14. The apparatus of claim 9, further comprising the second resonator structure.

15. The apparatus of claim 9, wherein, during operation, a power generator is coupled to one of the first and second resonant structure, with a coupling rate κ_g, and is configured to drive the resonator structure, to which it is coupled, at a driving frequency f, corresponding to a driving angular frequency $\omega = 2\pi f$,

wherein U_g is defined as κ_g / Γ_1, if the power generator is coupled to the first resonator structure and defined as κ_g / Γ_2, if the power generator is coupled to the second resonator structure.

16. The apparatus of claim 15, wherein the driving frequency is different from the resonance frequencies of the first and second resonator structures and is closer to a frequency corresponding to an odd normal mode of the system of the two resonator structures,

wherein the detuning of the first resonator from the driving frequency is defined as $D_1 = (\omega - \omega_1) / \Gamma_1$ and the detuning of the second resonator structure from the driving frequency is defined as $D_2 = (\omega - \omega_2) / \Gamma_2$.

17. The apparatus of claim 16, wherein D_1 is approximately equal to UV_{rad} and D_2 is approximately equal to UV_{rad}.

18. The apparatus of claim 15, wherein U_g is chosen to maximize the ratio of the energy-transfer efficiency to the radiation efficiency.

19. The apparatus of claim 17, wherein U_g is approximately equal to $\sqrt{1+U^2-V_{\text{rad}}^2U^2+V^2-2VV_{\text{rad}}}$.

20. The apparatus of claim 15, wherein *f* is at least larger than 100 kHz and smaller than 500MHz.

21. The apparatus of claim 15, wherein *f* is at least larger than 1MHz and smaller than 50MHz.

22. The apparatus of claim 15, further comprising the power generator.

23. The apparatus of claim 15, wherein, during operation, a power load is coupled to the resonant structure to which the power generator is not coupled, with a coupling rate

κ_l, and is configured to receive from the resonator structure, to which it is coupled, a usable power,

wherein U_l is defined as κ_l / Γ_1, if the power load is coupled to the first resonator structure and defined as κ_l / Γ_2, if the power load is coupled to the second resonator structure.

24. The apparatus of claim 23, wherein U_l is chosen to maximize the ratio of the energy-transfer efficiency to the radiation efficiency.

25. The apparatus of claim 24, wherein the driving frequency is different from the resonance frequencies of the first and second resonator structures and is closer to a frequency corresponding to an odd normal mode of the system of the two resonator structures,

wherein the detuning of the first resonator from the driving frequency is defined as $D_1 = (\omega - \omega_1) / \Gamma_1$ and is approximately equal to UV_{rad}, and the detuning of the second resonator structure from the driving frequency is defined as $D_2 = (\omega - \omega_2) / \Gamma_2$ and is approximately equal to UV_{rad},

and U_l is approximately equal to $\sqrt{1 + U^2 - V_{\text{rad}}^2 U^2 + V^2 - 2VV_{\text{rad}}}$.

26. The apparatus of claim 9, wherein at least one of the first and second resonator structures comprises a capacitively loaded loop or coil of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

27. The apparatus of claim 26, where the characteristic size of said loop or coil is less than 30 cm and the width of said conducting wire or Litz wire or ribbon is less than 2cm.

28. The apparatus of claim 26, where the characteristic size of said loop or coil is less than 1m and the width of said conducting wire or Litz wire or ribbon is less than 2cm.

29. The apparatus of claim 9, further comprising a feedback mechanism for maintaining the resonant frequency of one or more of the resonant objects.

30. The apparatus of claim 29, wherein the feedback mechanism comprises an oscillator with a fixed driving frequency and is configured to adjust the resonant frequency of the one or more resonant objects to be detuned by a fixed amount with respect to the fixed frequency.

31. An apparatus for use in wireless energy transfer, the apparatus comprising:

a first resonator structure configured for energy transfer with a second resonator structure over a distance *D*,

wherein the energy transfer is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, with a coupling factor *k*, wherein

said resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$, and is radiative in the far field, with an associated radiation quality factor $Q_{1,\mathrm{rad}} \geq Q_1$, and

said resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$, and is radiative in the far field, with an associated radiation quality factor $Q_{2,\mathrm{rad}} \geq Q_2$,

wherein an absolute value of a difference of said angular frequencies ω_1 and ω_2 is smaller than the broader of said resonant widths Γ_1 and Γ_2, and an average resonant angular frequency is defined as $\omega_o = \sqrt{\omega_1 \omega_2}$, corresponding to an average resonant wavelength $\lambda_o = 2\pi c / \omega_o$, where *c* is the speed of light in free space, and a strong-coupling factor is defined as $U = k\sqrt{Q_1 Q_2}$,

wherein the apparatus is configured to employ interference between said radiative far fields of the resonant fields of the first and second resonator, with an interference factor V_{rad}, to increase efficiency of energy transfer for the apparatus compared to efficiency for the apparatus in the absence of interference, the strong-interference factor being defined as $V = V_{\text{rad}}\sqrt{(Q_1 / Q_{1,\text{rad}})(Q_2 / Q_{2,\text{rad}})}$.

32. The apparatus of claim 31, wherein $Q_1 / Q_{1,\text{rad}} \geq 0.05$ and $Q_2 / Q_{2,\text{rad}} \geq 0.05$.

33. The apparatus of claim 31, wherein $Q_1 / Q_{1,\text{rad}} \geq 0.5$ and $Q_2 / Q_{2,\text{rad}} \geq 0.5$.

34. The apparatus of claim 31, wherein D / λ_o is larger than 0.01 and the strong-interference factor V is larger than 0.05.

35. The apparatus of claim 31, wherein D / λ_o is larger than 0.01 and the strong-interference factor V is larger than 0.5.

36. The apparatus of claim 31, further comprising the second resonator structure.

37. The apparatus of claim 31, wherein, during operation, a power generator is coupled to one of the first and second resonant structure, with a coupling rate κ_g, and is configured to drive the resonator structure, to which it is coupled, at a driving frequency *f*, corresponding to a driving angular frequency $\omega = 2\pi f$,

wherein U_g is defined as κ_g / Γ_1, if the power generator is coupled to the first resonator structure and defined as κ_g / Γ_2, if the power generator is coupled to the second resonator structure.

38. The apparatus of claim 37, wherein the driving frequency is different from the resonance frequencies of the first and second resonator structures and is closer to a frequency corresponding to an odd normal mode of the system of the two resonator structures,

wherein the detuning of the first resonator from the driving frequency is defined as $D_1 = (\omega - \omega_1)/\Gamma_1$ and the detuning of the second resonator structure from the driving frequency is defined as $D_2 = (\omega - \omega_2)/\Gamma_2$.

39. The apparatus of claim 38, wherein D_1 is approximately equal to UV and D_2 is approximately equal to UV.

40. The apparatus of claim 37, wherein U_g is chosen to maximize the energy-transfer efficiency.

41. The apparatus of claim 39, wherein U_g is approximately equal to $\sqrt{(1+U^2)(1-V^2)}$.

42. The apparatus of claim 37, wherein *f* is at least larger than 100 kHz and smaller than 500MHz.

43. The apparatus of claim 37, wherein *f* is at least larger than 1MHz and smaller than 50MHz.

44. The apparatus of claim 37, further comprising the power generator.

45. The apparatus of claim 37, wherein, during operation, a power load is coupled to the resonant structure to which the power generator is not coupled, with a coupling rate κ_l, and is configured to receive from the resonator structure, to which it is coupled, a usable power,

wherein U_l is defined as κ_l / Γ_1, if the power load is coupled to the first resonator structure and defined as κ_l / Γ_2, if the power load is coupled to the second resonator structure.

46. The apparatus of claim 45, wherein U_l is chosen to maximize the energy-transfer efficiency.

47. The apparatus of claim 46, wherein the driving frequency is different from the resonance frequencies of the first and second resonator structures and is closer to a frequency corresponding to an odd normal mode of the system of the two resonator structures,

wherein the detuning of the first resonator from the driving frequency is defined as $D_1 = (\omega - \omega_1)/\Gamma_1$ and is approximately equal to UV, and the detuning of the second resonator structure from the driving frequency is defined as $D_2 = (\omega - \omega_2)/\Gamma_2$ and is approximately equal to UV,

and U_l is approximately equal to $\sqrt{(1+U^2)(1-V^2)}$.

48. The apparatus of claim 31, wherein at least one of the first and second resonator structures comprises a capacitively loaded loop or coil of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

49. The apparatus of claim 48, where the characteristic size of said loop or coil is less than 30 cm and the width of said conducting wire or Litz wire or ribbon is less than 2cm.

50. The apparatus of claim 48, where the characteristic size of said loop or coil is less than 1m and the width of said conducting wire or Litz wire or ribbon is less than 2cm.

51. The apparatus of claim 31, further comprising a feedback mechanism for maintaining the resonant frequency of one or more of the resonant objects.

52. The apparatus of claim 51, wherein the feedback mechanism comprises an oscillator with a fixed driving frequency and is configured to adjust the resonant frequency of the one or more resonant objects to be detuned by a fixed amount with

respect to the fixed frequency.

53. The apparatus of claim 51, where the feedback mechanism is configured to monitor an efficiency of the energy transfer, and adjust the resonant frequency of the one or more resonant objects to maximize the efficiency.

54. The apparatus of claim 31, wherein the resonance angular frequencies for the resonator structures are selected to optimize the energy-transfer efficiency by accounting for both the strong-coupling factor U and the strong-interference interference factor V.

55. A method for wireless energy transfer, the method comprising:

providing a first resonator structure configured for energy transfer with a second resonator structure over a distance D,

wherein the energy transfer is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, with a coupling factor k, wherein

said resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$, and is radiative in the far field, with an associated radiation quality factor $Q_{1,\text{rad}} \geq Q_1$, and

said resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$, and is radiative in the far field, with an associated radiation quality factor $Q_{2,\text{rad}} \geq Q_2$,

wherein an absolute value of a difference of said angular frequencies ω_1 and ω_2 is smaller than broader of said resonant widths Γ_1 and Γ_2, and an average resonant angular frequency is defined as $\omega_o = \sqrt{\omega_1\omega_2}$, corresponding to an average resonant wavelength $\lambda_o = 2\pi c / \omega_o$, where c is the speed of light in free space, and the strong-coupling factor is defined as $U = k\sqrt{Q_1 Q_2}$, and employing interference between said radiative far fields of the resonant fields of the first and second resonator, with an interference factor V_{rad}, to reduce a total amount of radiation from the first and second resonator compared to an amount of radiation from the first and second resonator in the absence of interference, a strong-interference factor being defined as $V = V_{\text{rad}}\sqrt{(Q_1 / Q_{1,\text{rad}})(Q_2 / Q_{2,\text{rad}})}$.

56. The method of claim 55, wherein $Q_1 / Q_{1,\text{rad}} \geq 0.01$ and $Q_2 / Q_{2,\text{rad}} \geq 0.01$.

57. The method of claim 55, wherein, during operation, a power generator is coupled to one of the first and second resonant structure and is configured to drive the resonator structure, to which it is coupled, at a driving frequency *f*, corresponding to a driving angular frequency $\omega = 2\pi f$,

wherein the driving frequency is different from the resonance frequencies of the first and second resonator structures and is closer to a frequency corresponding to an odd normal mode of the system of the two resonator structures.

58. The method of claim 57, wherein, during operation, a power load is coupled to the resonant structure to which the power generator is not coupled and is configured to receive from the resonator structure, to which it is coupled, a usable power.

59. A method for wireless energy transfer, the method comprising:

providing a first resonator structure configured for energy transfer with a second resonator structure over a distance *D*,

wherein the energy transfer is mediated by evanescent-tail coupling of a resonant field of the first resonator structure and a resonant field of the second resonator structure, with a coupling factor *k*, wherein

said resonant field of the first resonator structure has a resonance angular frequency ω_1, a resonance frequency-width Γ_1, and a resonance quality factor $Q_1 = \omega_1 / 2\Gamma_1$, and is radiative in the far field, with an associated radiation quality factor

$Q_{1,\text{rad}} \geq Q_1$, and

said resonant field of the second resonator structure has a resonance angular frequency ω_2, a resonance frequency-width Γ_2, and a resonance quality factor $Q_2 = \omega_2 / 2\Gamma_2$, and is radiative in the far field, with an associated radiation quality factor $Q_{2,\text{rad}} \geq Q_2$,

wherein an absolute value of the difference of said angular frequencies ω_1 and ω_2 is smaller than the broader of said resonant widths Γ_1 and Γ_2,and an average resonant angular frequency is defined as $\omega_o = \sqrt{\omega_1\omega_2}$, corresponding to an average resonant wavelength $\lambda_o = 2\pi c / \omega_o$, where c is the speed of light in free space, and the strong-coupling factor is defined as $U = k\sqrt{Q_1 Q_2}$, and

employing interference between said radiative far fields of the resonant fields of the first and second resonator, with an interference factor V_{rad}, to increase efficiency of energy transfer between the first and second resonator compared to efficiency of energy transfer between the first and second resonator in the absence of interference, a strong-interference factor being defined as $V = V_{\text{rad}}\sqrt{\left(Q_1 / Q_{1,\text{rad}}\right)\left(Q_2 / Q_{2,\text{rad}}\right)}$.

60. The method of claim 59, wherein $Q_1 / Q_{1,\text{rad}} \geq 0.05$ and $Q_2 / Q_{2,\text{rad}} \geq 0.05$.

61. The method of claim 59, wherein, during operation, a power generator is coupled to one of the first and second resonant structure and is configured to drive the resonator structure, to which it is coupled, at a driving frequency *f*, corresponding to a driving angular frequency $\omega = 2\pi f$,

wherein the driving frequency is different from the resonance frequencies of the first and second resonator structures and is closer to a frequency corresponding to an odd normal mode of the system of the two resonator structures.

62. The method of claim 61, wherein, during operation, a power load is coupled to the resonant structure to which the power generator is not coupled and is configured to

receive from the resonator structure, to which it is coupled, a usable power.

63. The method of claim 59, wherein the resonance angular frequencies for the resonator structures are selected to optimize the energy-transfer efficiency by accounting for both the strong-coupling factor U and the strong-interference interference factor V.

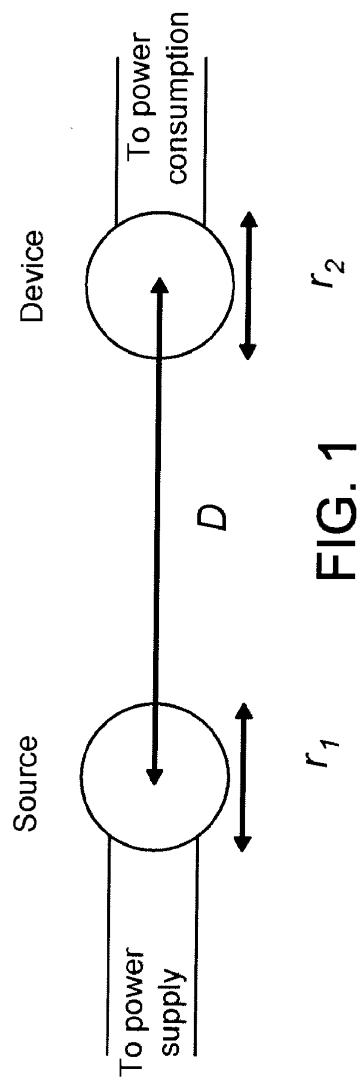


FIG. 1

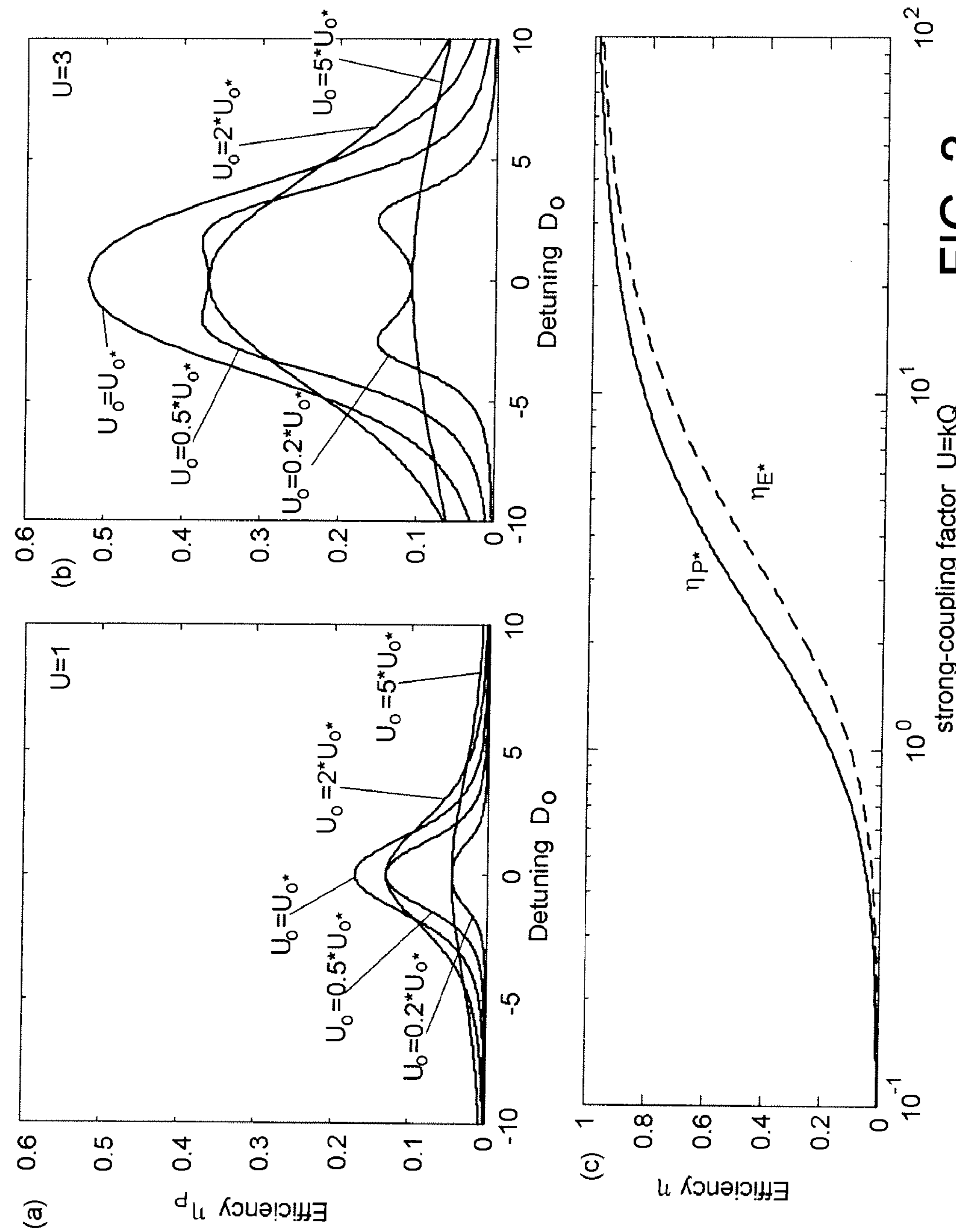

(a)
Efficiency η_P
U=1
Detuning D_o
$U_o=U_o*$
$U_o=0.5*U_o*$
$U_o=0.2*U_o*$
$U_o=2*U_o*$
$U_o=5*U_o*$
(b)
U=3
Detuning D_o
$U_o=U_o*$
$U_o=0.5*U_o*$
$U_o=0.2*U_o*$
$U_o=2*U_o*$
$U_o=5*U_o*$
(c)
Efficiency η
η_{P*}
η_{E*}
strong-coupling factor U=kQ

FIG. 2

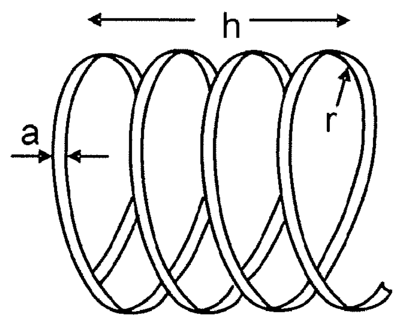


FIG. 3

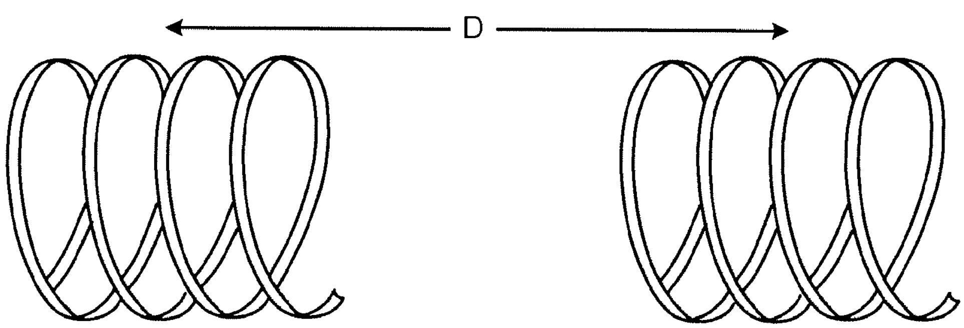


FIG. 4

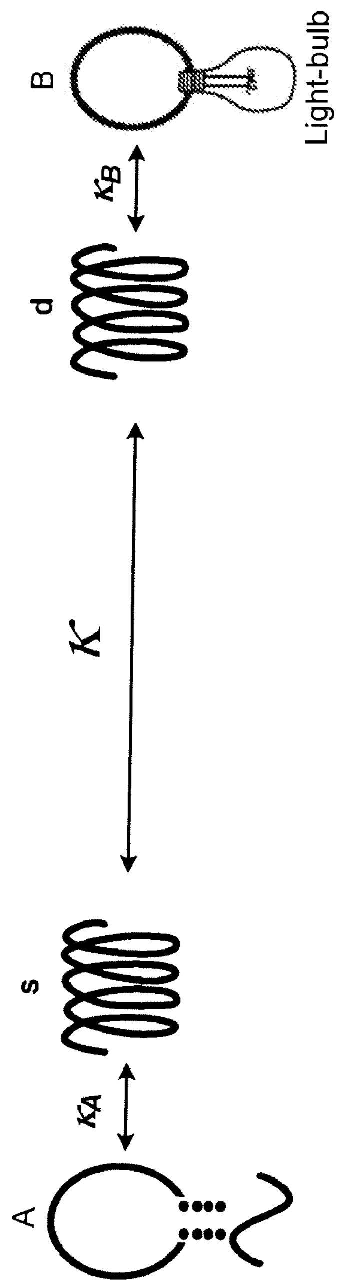
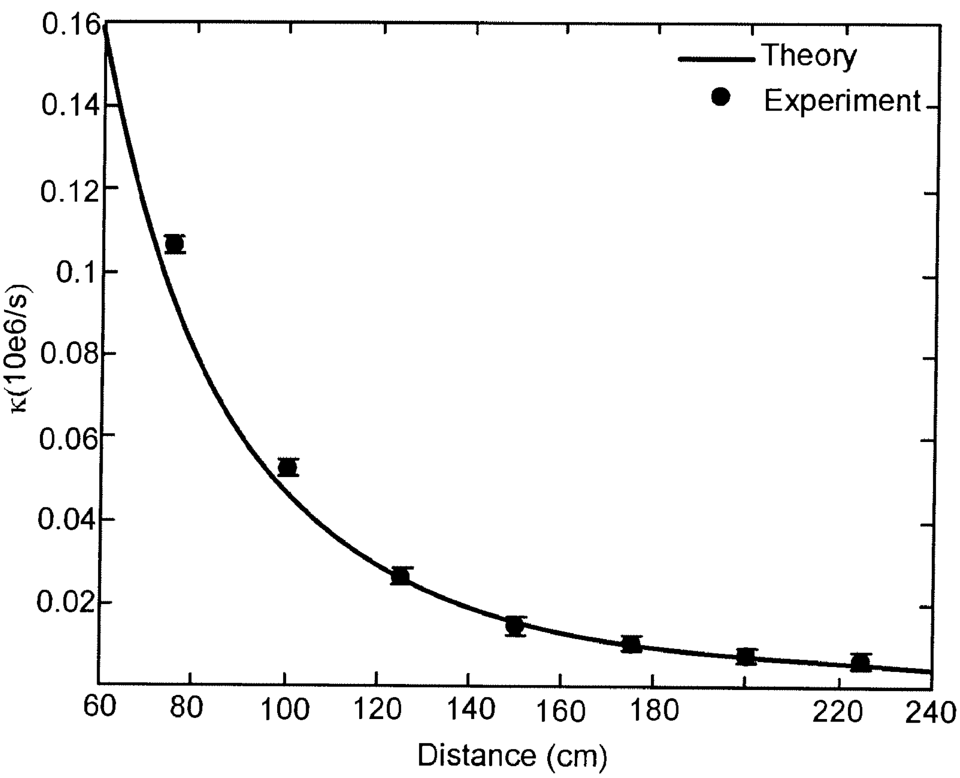
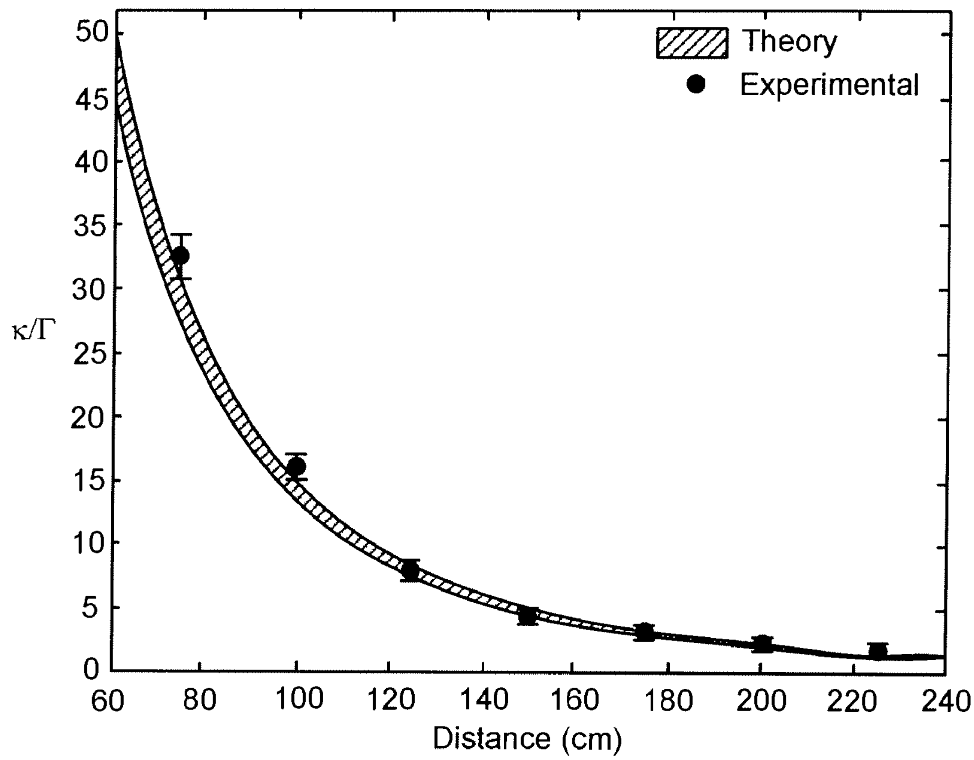


FIG. 5



Comparison of experimental and theoretical values for κ as a function of the separation between the source and device coils.

FIG. 6



Comparison of experimental and theoretical values for the parameter κ/Γ as a function of the separation between the two coils. The theory values are obtained by using the theoretically obtained κ and the experimentally measured Γ. The shaded area represents the spread in the theoretical κ/Γ due to the ~5% uncertainty in Q.

FIG.7

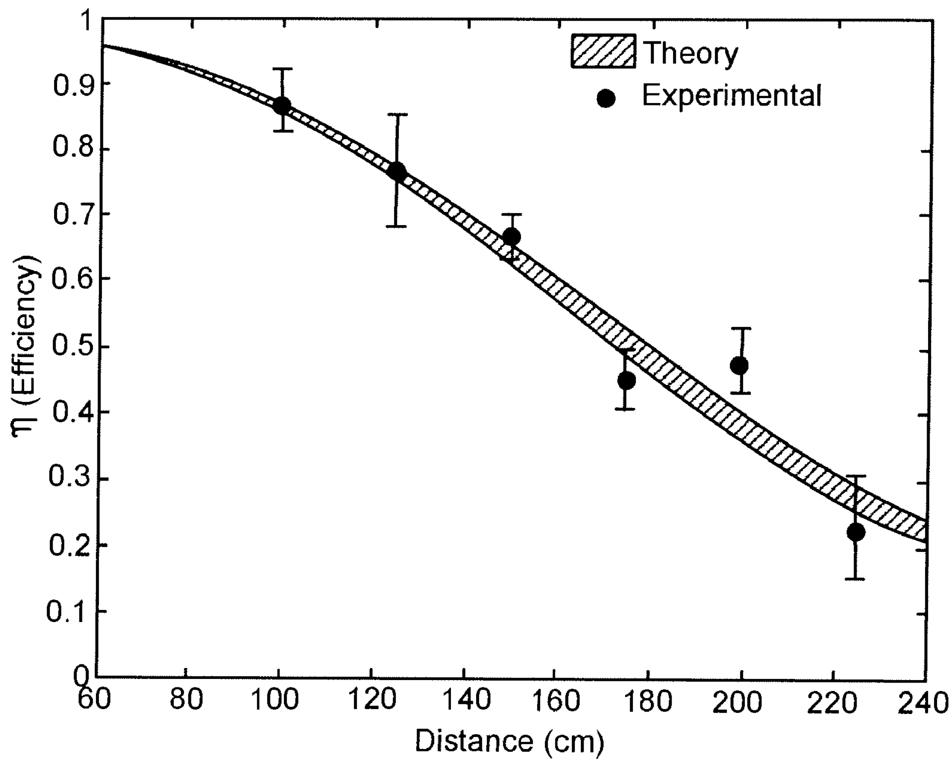

Theory
Experimental
η (Efficiency)
Distance (cm)
0
0.1
0.2
0.3
0.4
0.5
0.6
0.7
0.8
0.9
1
60
80
100
120
140
160
180
200
220
240

FIG. 8

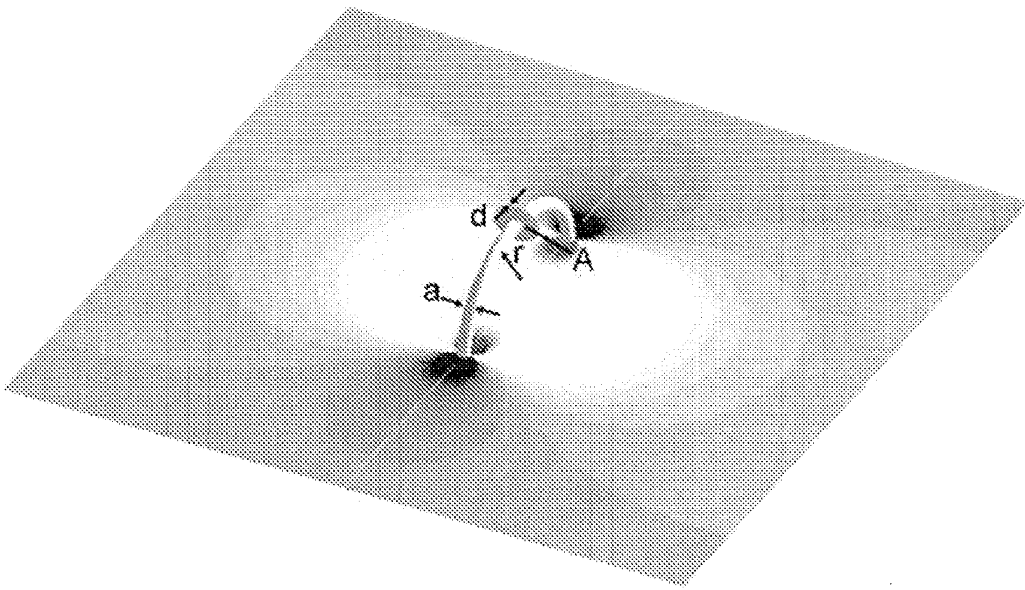


FIG. 9

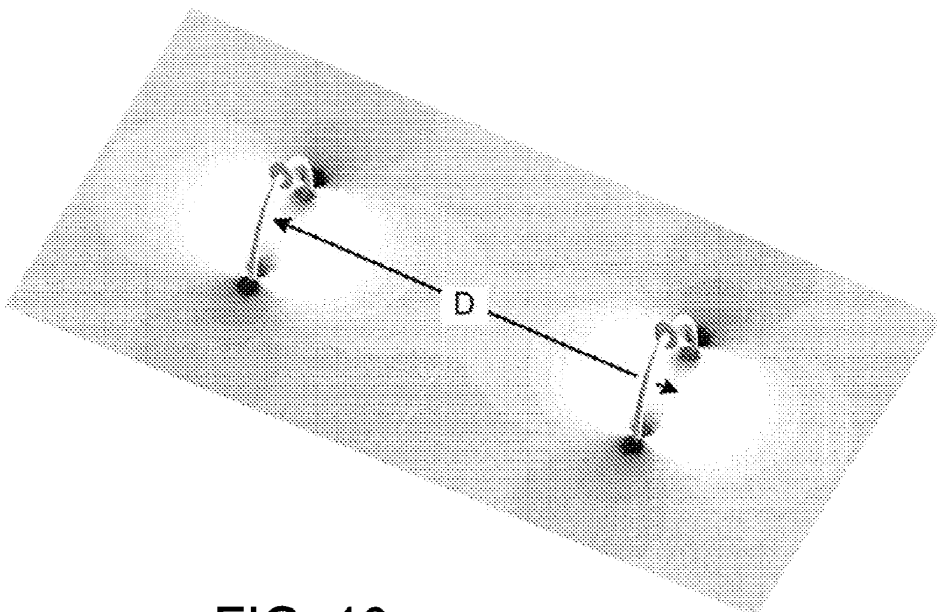


FIG. 10

SUBSTITUTE SHEET (RULE 26)

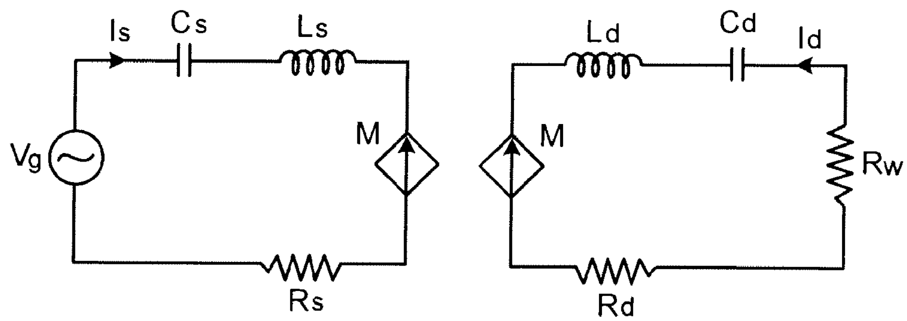


FIG. 11

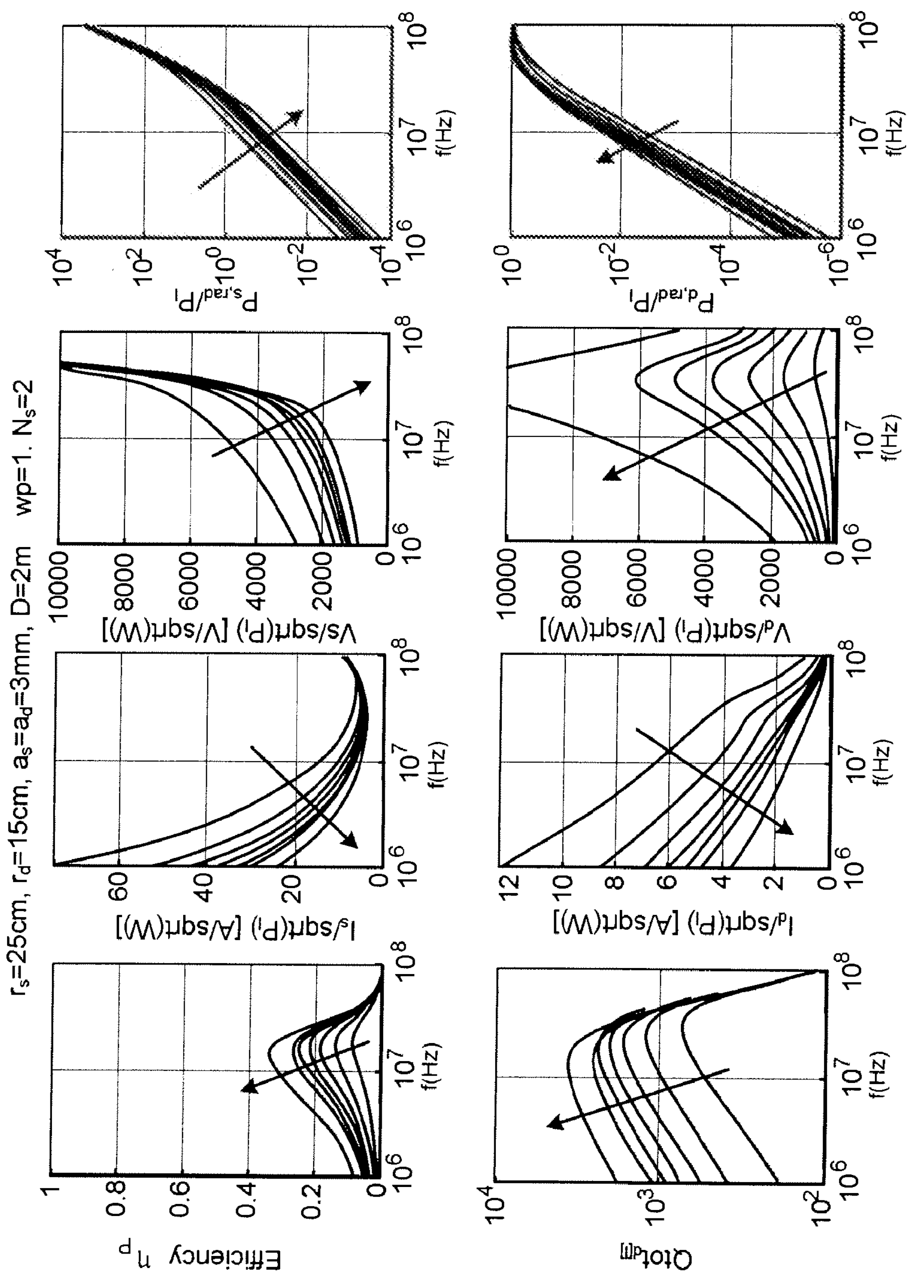

rs=25cm, rd=15cm, as=ad=3mm, D=2m wp=1. Ns=2
Efficiency ηp
Is/sqrt(Pl) [A/sqrt(W)]
Vs/sqrt(Pl) [V/sqrt(W)]
Ps,rad/Pl
Qtot[dl]
Id/sqrt(Pl) [A/sqrt(W)]
Vd/sqrt(Pl) [V/sqrt(W)]
Pd,rad/Pl
f(Hz)

FIG. 12

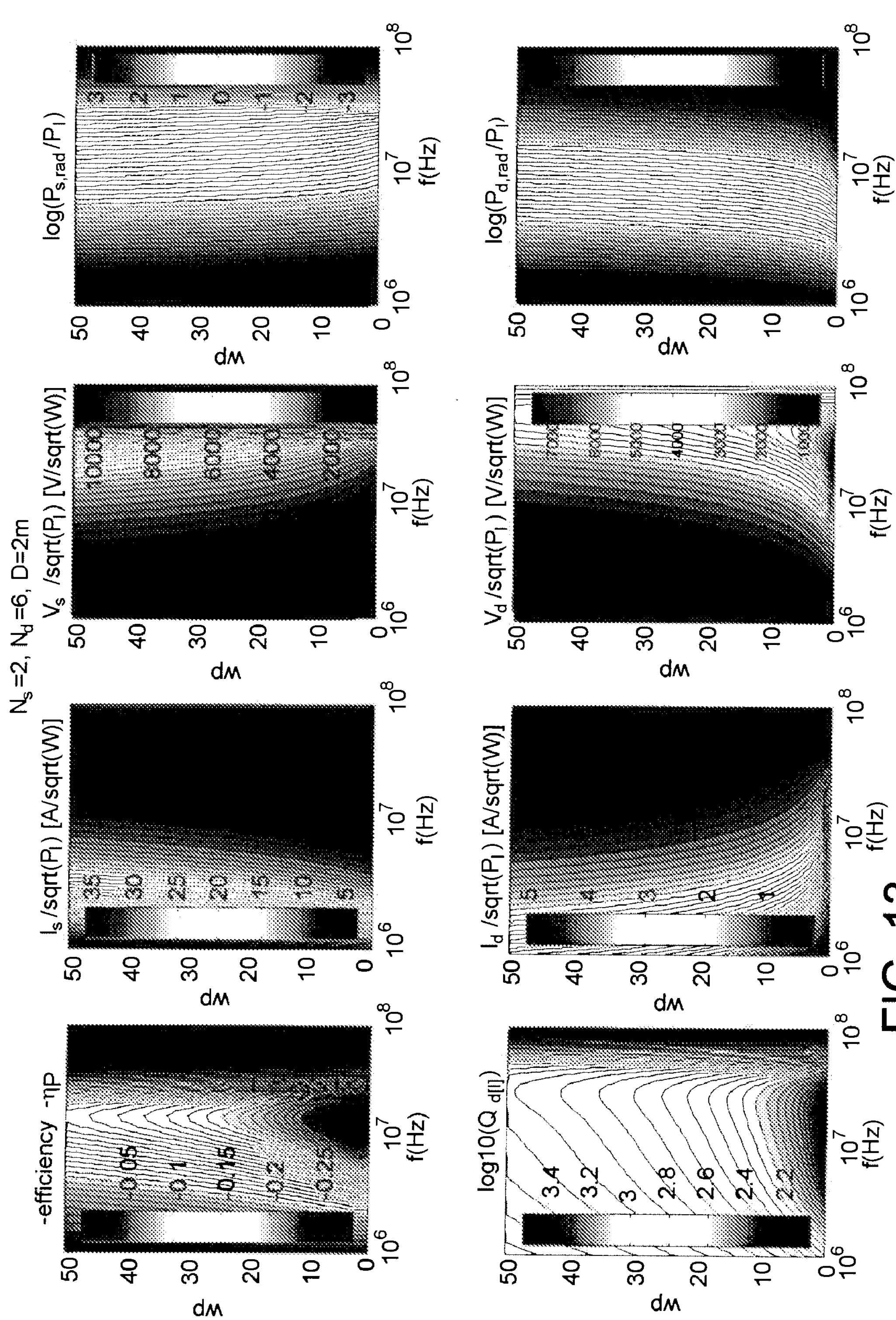

Ns=2, Nd=6, D=2m
-efficiency -ηP
Is /sqrt(Pl) [A/sqrt(W)]
Vs /sqrt(Pl) [V/sqrt(W)]
log(Ps,rad/Pl)
log10(Qd[l])
Id /sqrt(Pl) [A/sqrt(W)]
Vd /sqrt(Pl) [V/sqrt(W)]
log(Pd,rad/Pl)
wp
f(Hz)

FIG. 13

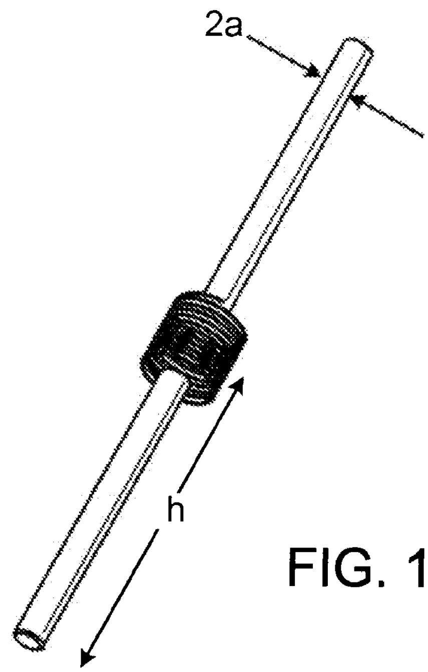


FIG. 14

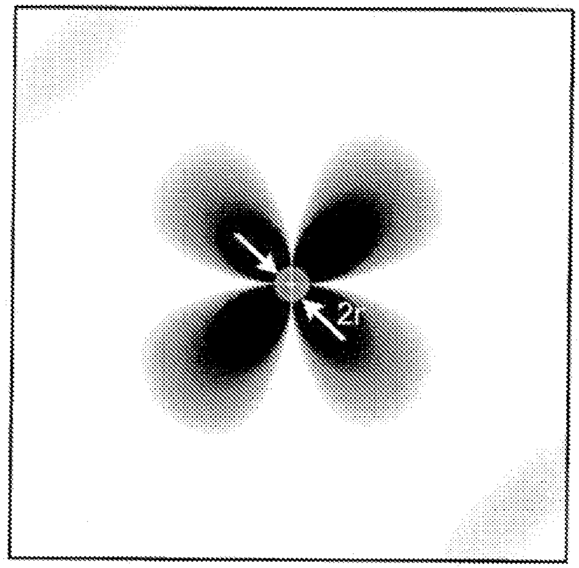


FIG. 15A

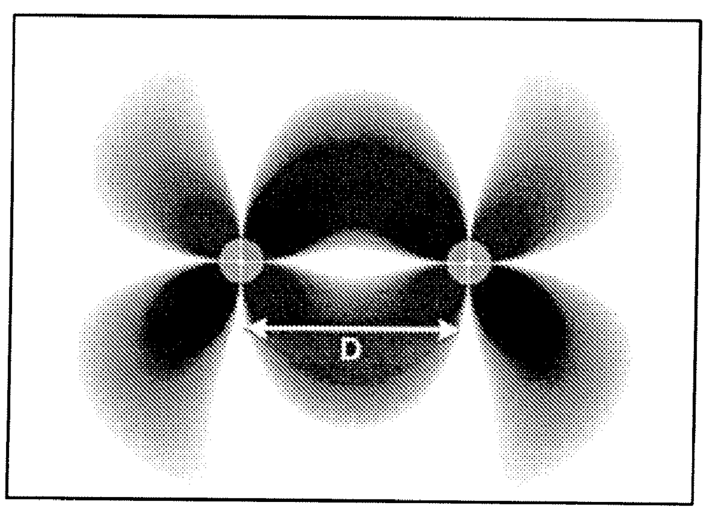


FIG. 15B

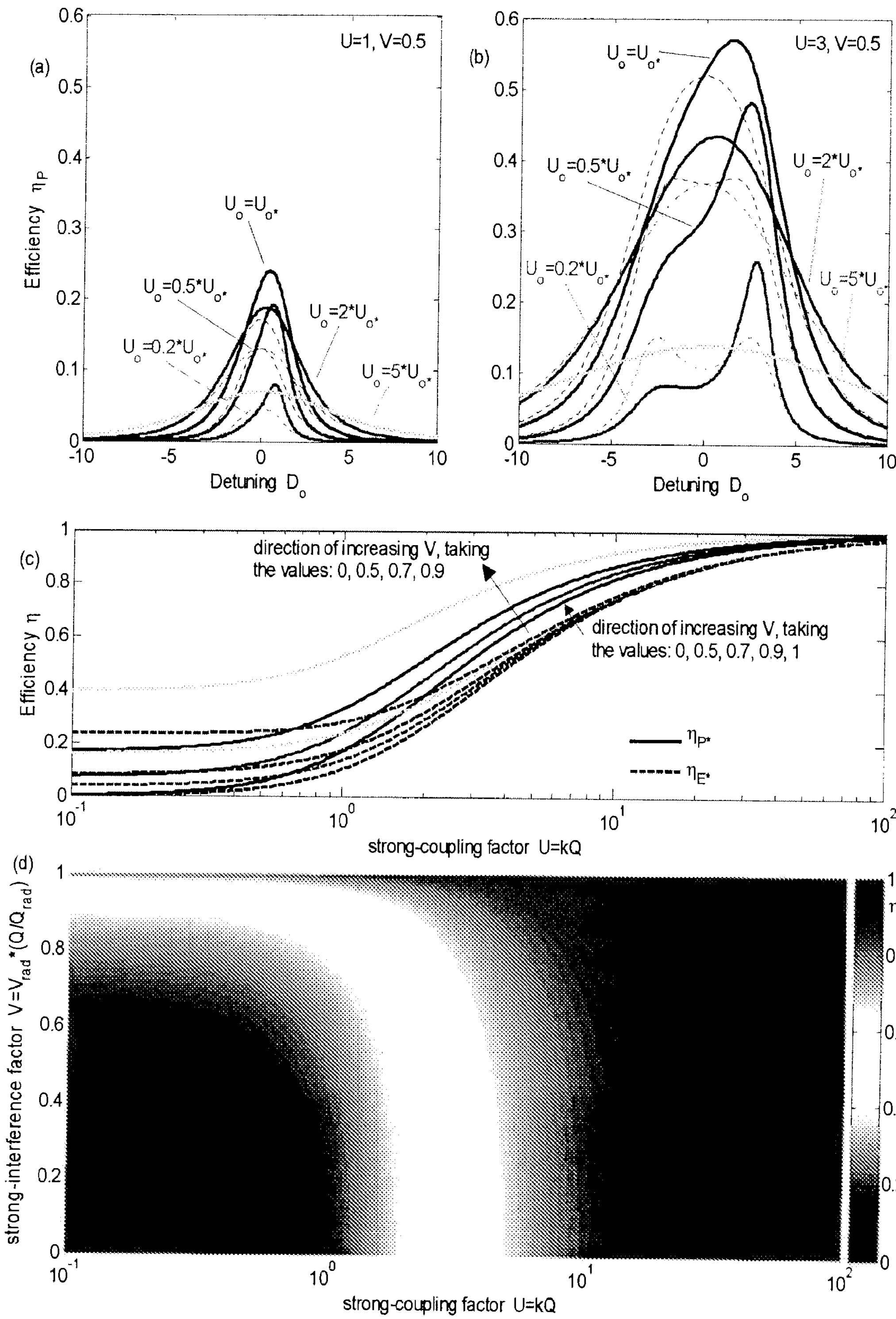

(a)
U=1, V=0.5
Efficiency η_P
U_o=U_o*
U_o=0.5*U_o*
U_o=2*U_o*
U_o=0.2*U_o*
U_o=5*U_o*
Detuning D_o
(b)
U=3, V=0.5
U_o=U_o*
U_o=0.5*U_o*
U_o=2*U_o*
U_o=0.2*U_o*
U_o=5*U_o*
Detuning D_o
(c)
Efficiency η
direction of increasing V, taking the values: 0, 0.5, 0.7, 0.9
direction of increasing V, taking the values: 0, 0.5, 0.7, 0.9, 1
η_P*
η_E*
strong-coupling factor U=kQ
(d)
strong-interference factor V=V_rad*(Q/Q_rad)
η_P*
strong-coupling factor U=kQ

FIG. 16

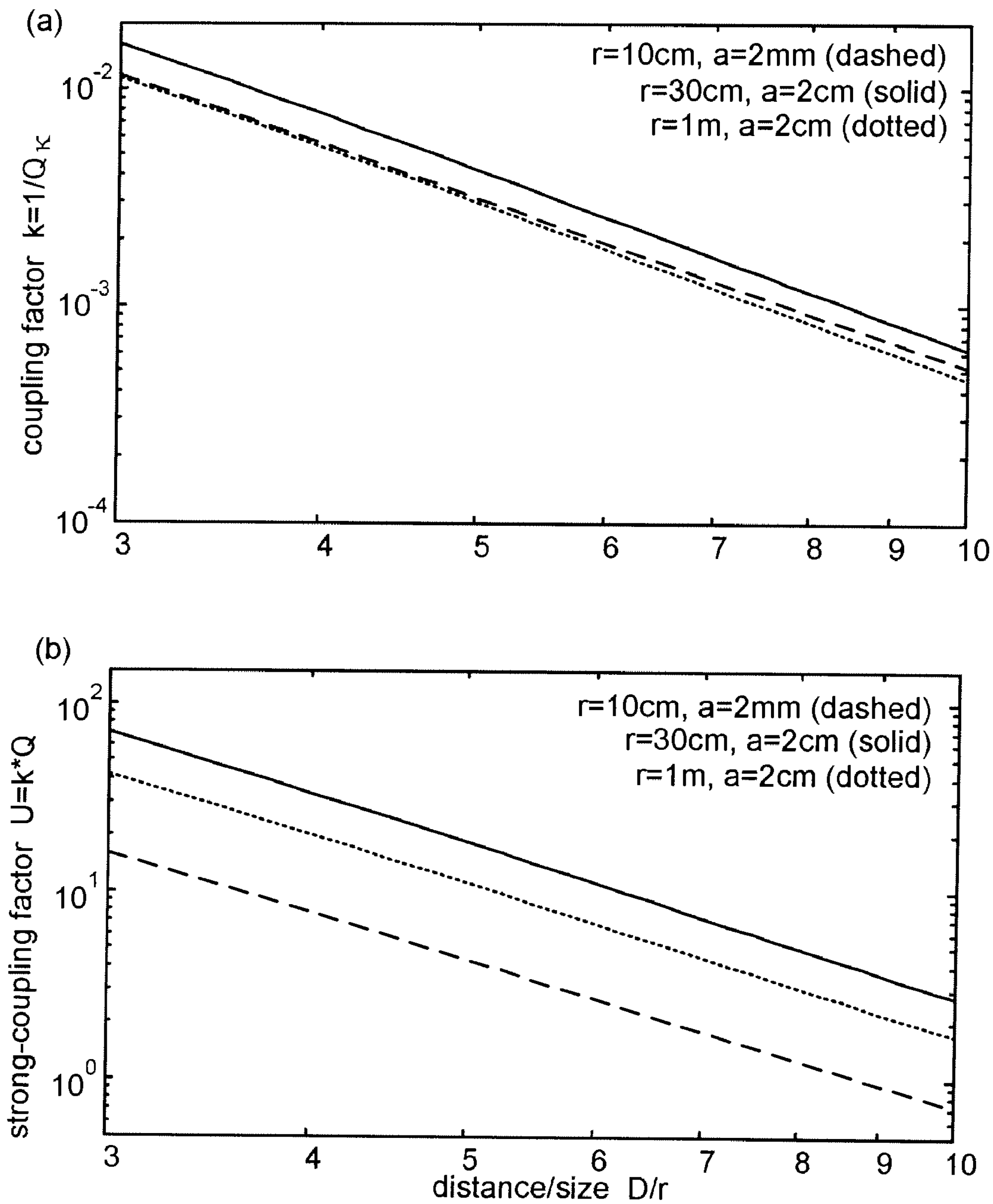

(a)
r=10cm, a=2mm (dashed)
r=30cm, a=2cm (solid)
r=1m, a=2cm (dotted)
coupling factor $k=1/Q_\kappa$
10^{-2}
10^{-3}
10^{-4}
3
4
5
6
7
8
9
10
(b)
r=10cm, a=2mm (dashed)
r=30cm, a=2cm (solid)
r=1m, a=2cm (dotted)
strong-coupling factor U=k*Q
10^{2}
10^{1}
10^{0}
3
4
5
6
7
8
9
10
distance/size D/r

FIG. 17

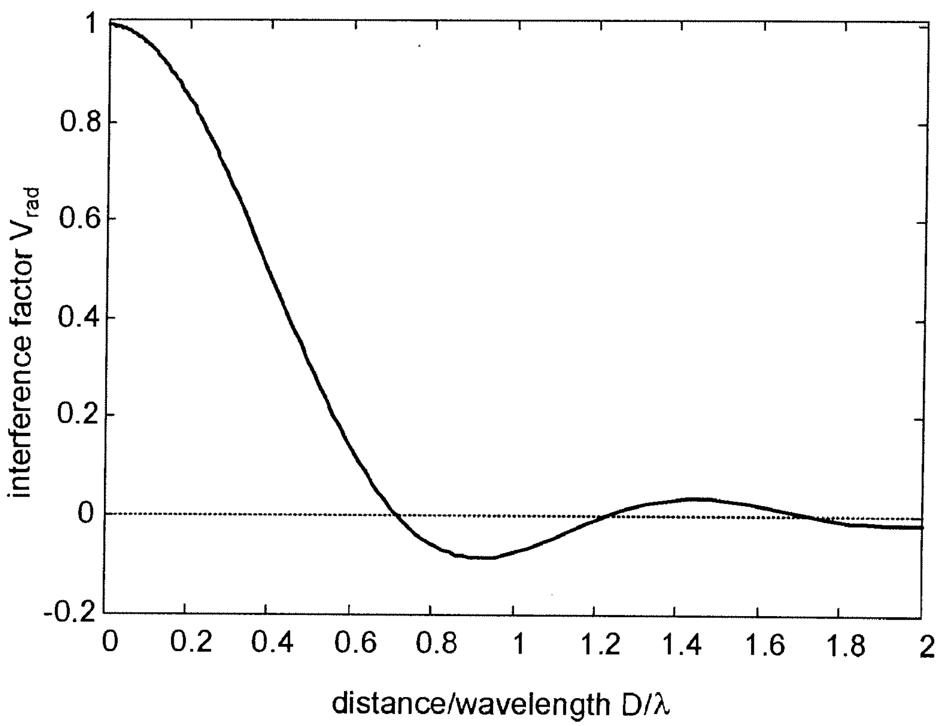

1
0.8
0.6
0.4
0.2
0
-0.2
interference factor V_{rad}
0
0.2
0.4
0.6
0.8
1
1.2
1.4
1.6
1.8
2
distance/wavelength D/λ

FIG. 18

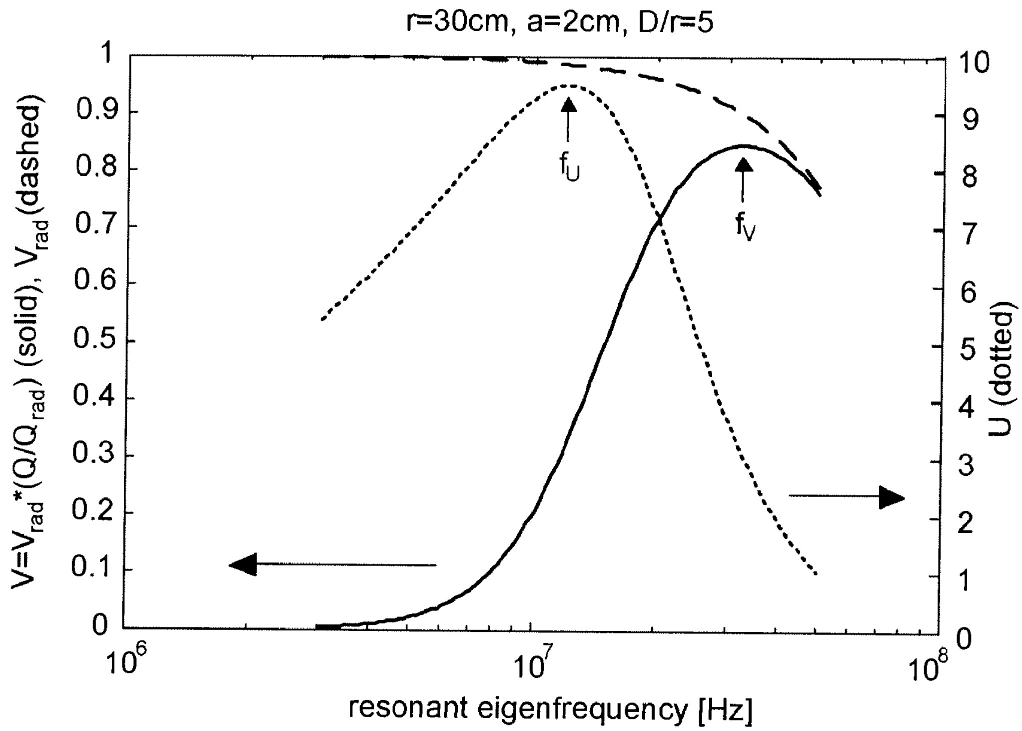
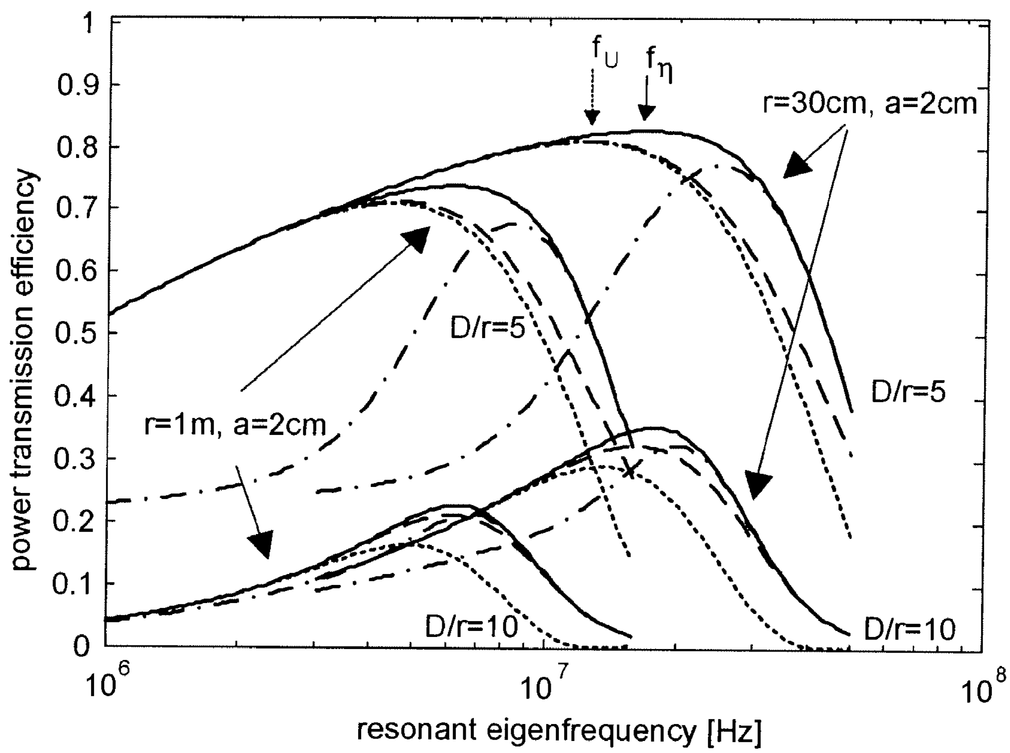
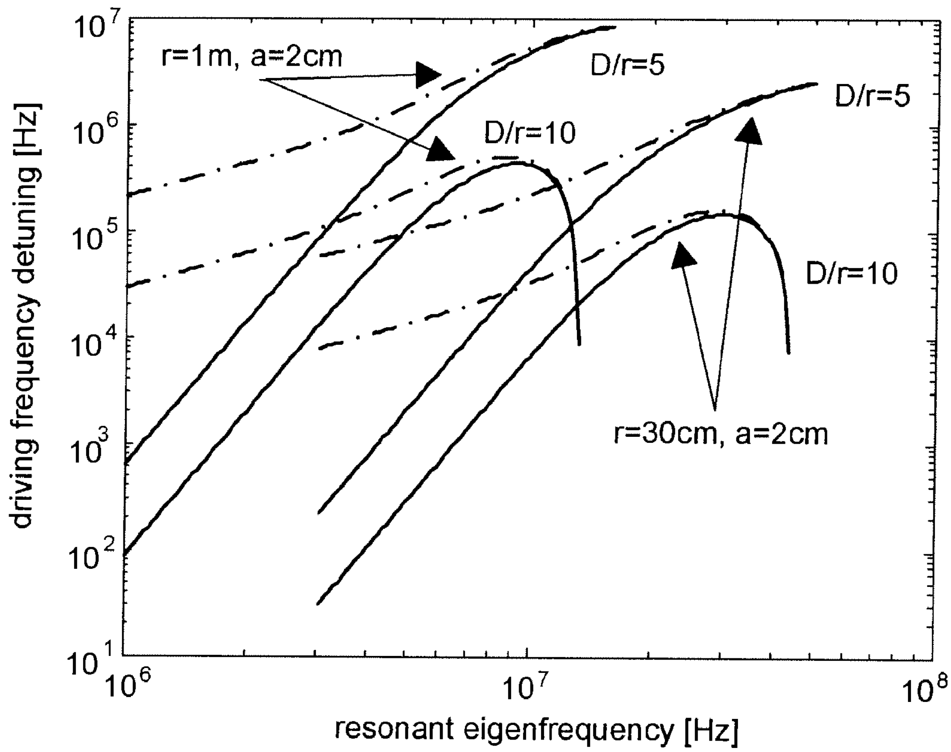

r=30cm, a=2cm, D/r=5
1
0.9
0.8
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0
10
9
8
7
6
5
4
3
2
1
0
$V=V_{rad}*(Q/Q_{rad})$ (solid), V_{rad}(dashed)
U (dotted)
f_U
f_V
10^6
10^7
10^8
resonant eigenfrequency [Hz]

FIG. 19



power transmission efficiency
resonant eigenfrequency [Hz]
f_U
f_η
r=30cm, a=2cm
r=1m, a=2cm
D/r=5
D/r=5
D/r=10
D/r=10

FIG. 20



r=1m, a=2cm
D/r=5
D/r=5
D/r=10
D/r=10
r=30cm, a=2cm
driving frequency detuning [Hz]
resonant eigenfrequency [Hz]
10^7
10^6
10^5
10^4
10^3
10^2
10^1
10^6
10^7
10^8

FIG. 21

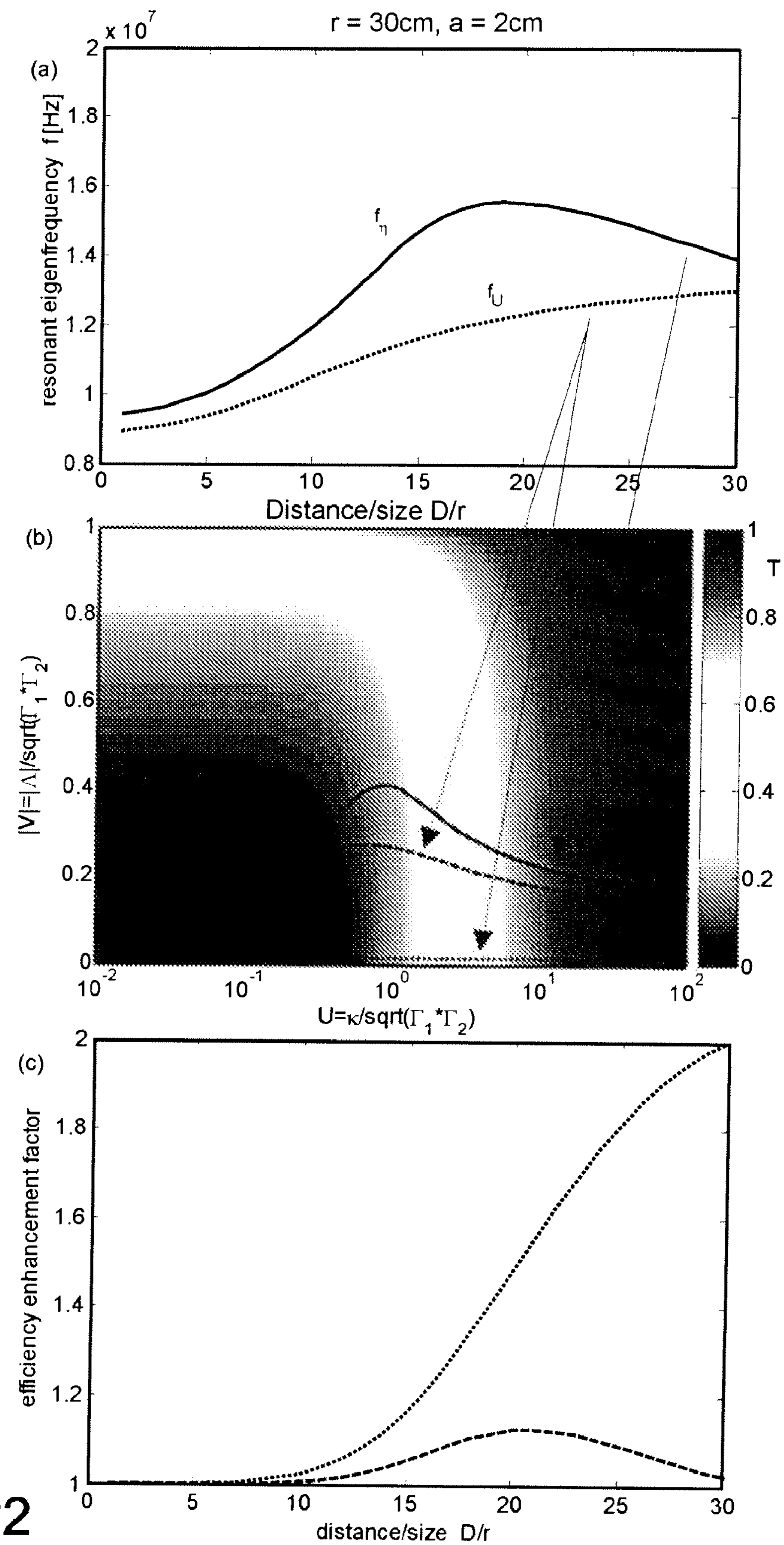
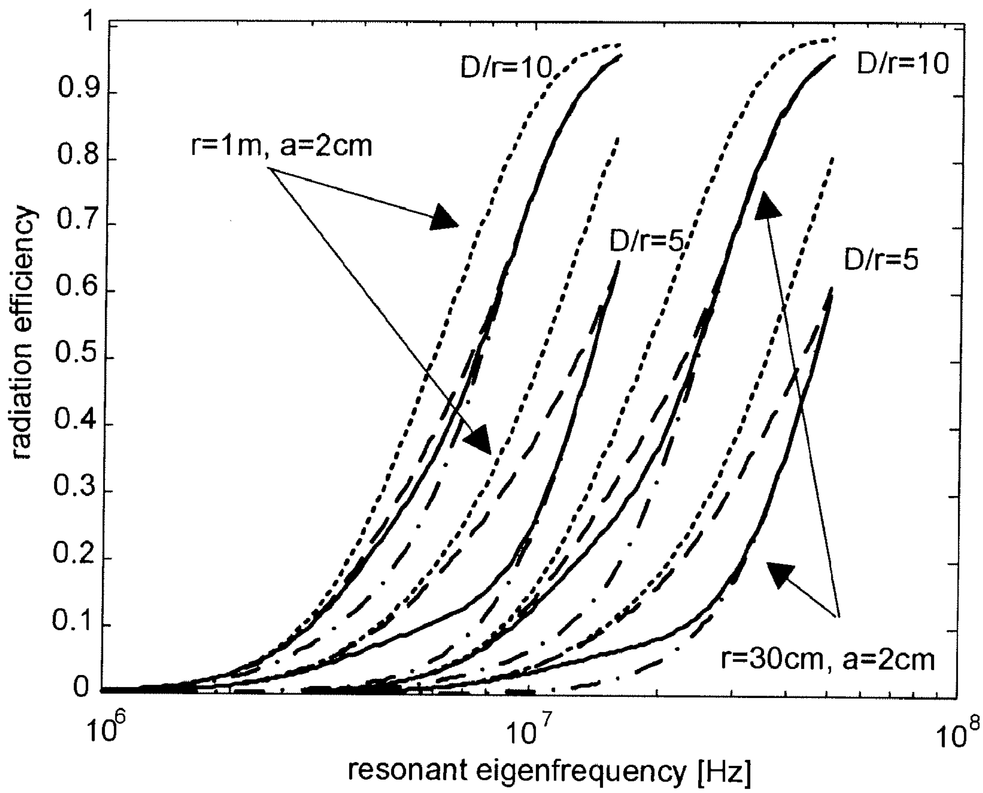

r = 30cm, a = 2cm
(a)
x 10^7
resonant eigenfrequency f [Hz]
f_η
f_U
Distance/size D/r
(b)
T
|V|=|Λ|/sqrt(Γ1*Γ2)
U=κ/sqrt(Γ1*Γ2)
(c)
efficiency enhancement factor
distance/size D/r

FIG. 22



1
0.9
0.8
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0
radiation efficiency
10^6
10^7
10^8
resonant eigenfrequency [Hz]
D/r=10
r=1m, a=2cm
D/r=5
D/r=10
D/r=5
r=30cm, a=2cm

FIG. 23

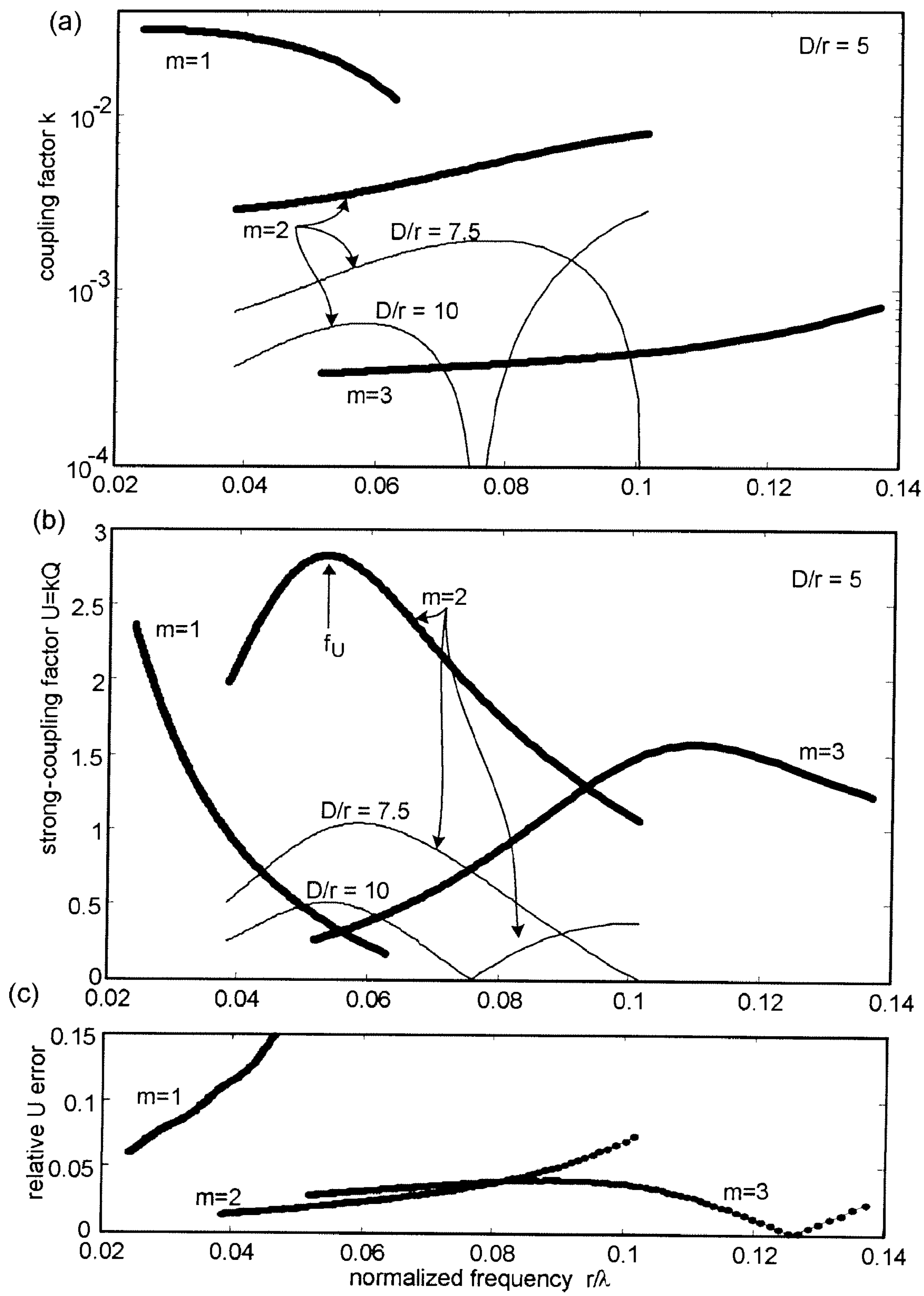

(a)
D/r = 5
m=1
coupling factor k
10^{-2}
10^{-3}
10^{-4}
m=2
D/r = 7.5
D/r = 10
m=3
0.02
0.04
0.06
0.08
0.1
0.12
0.14
(b)
D/r = 5
strong-coupling factor U=kQ
m=1
m=2
f_U
m=3
D/r = 7.5
D/r = 10
0
0.5
1
1.5
2
2.5
3
(c)
relative U error
0.15
0.1
0.05
m=1
m=2
m=3
normalized frequency r/λ

FIG. 24

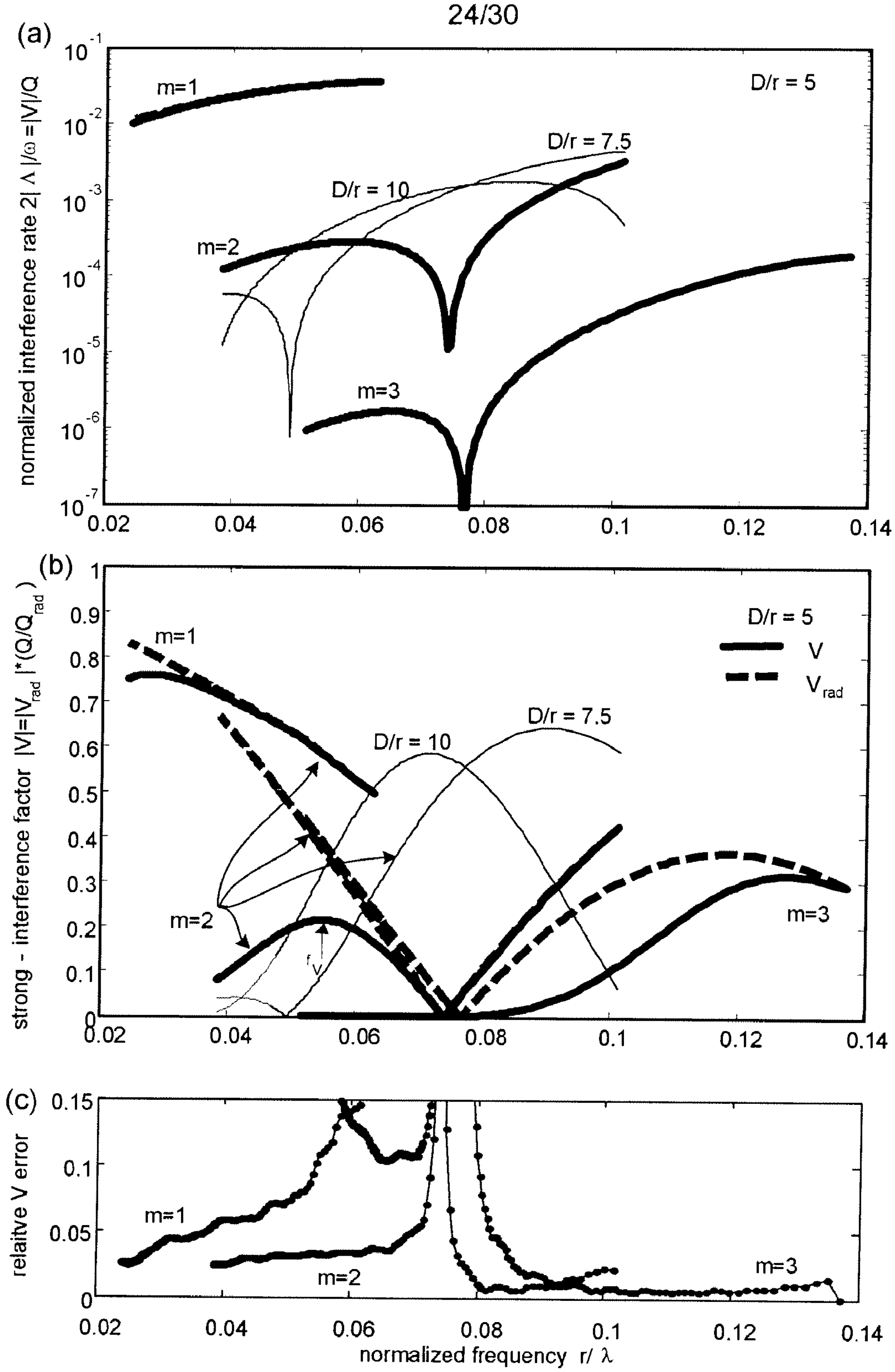
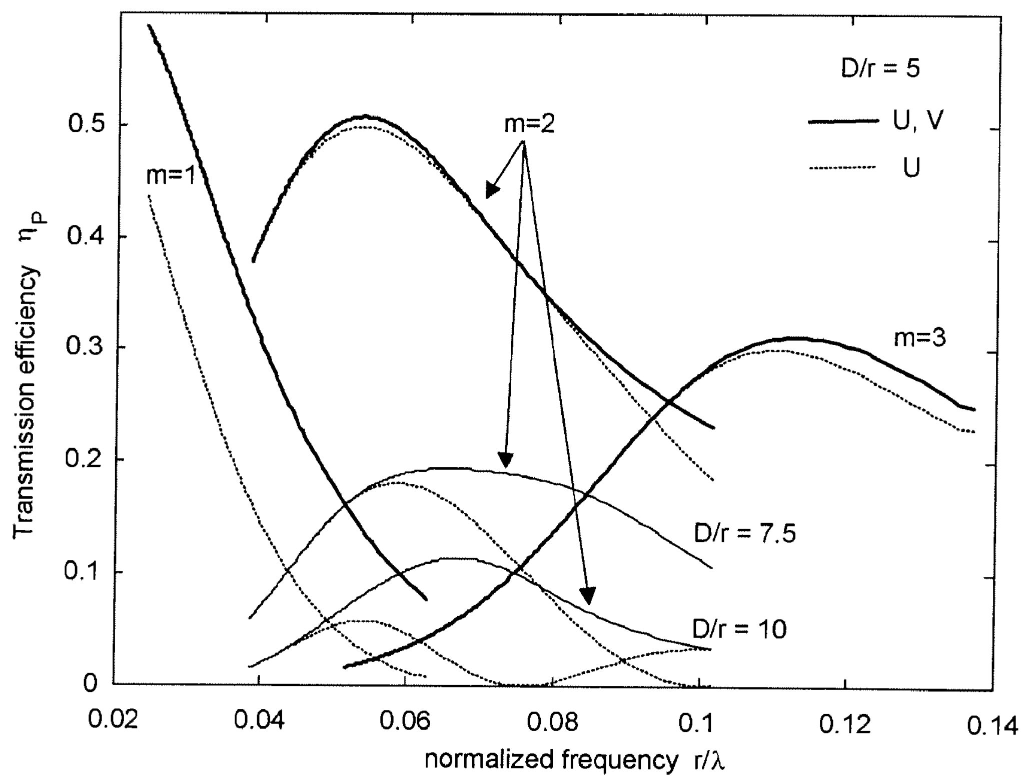

(a)
normalized interference rate 2| Λ |/ω =|V|/Q
m=1
D/r = 5
D/r = 7.5
D/r = 10
m=2
m=3
(b)
strong - interference factor |V|=|V_rad |*(Q/Q_rad)
D/r = 5
V
V_rad
m=1
D/r = 10
D/r = 7.5
m=2
m=3
f_V
(c)
relative V error
m=1
m=2
m=3
normalized frequency r/ λ

FIG. 25



D/r = 5
U, V
U
m=2
m=1
m=3
D/r = 7.5
D/r = 10
Transmission efficiency η_P
0.5
0.4
0.3
0.2
0.1
0
0.02
0.04
0.06
0.08
0.1
0.12
0.14
normalized frequency r/λ

FIG. 26

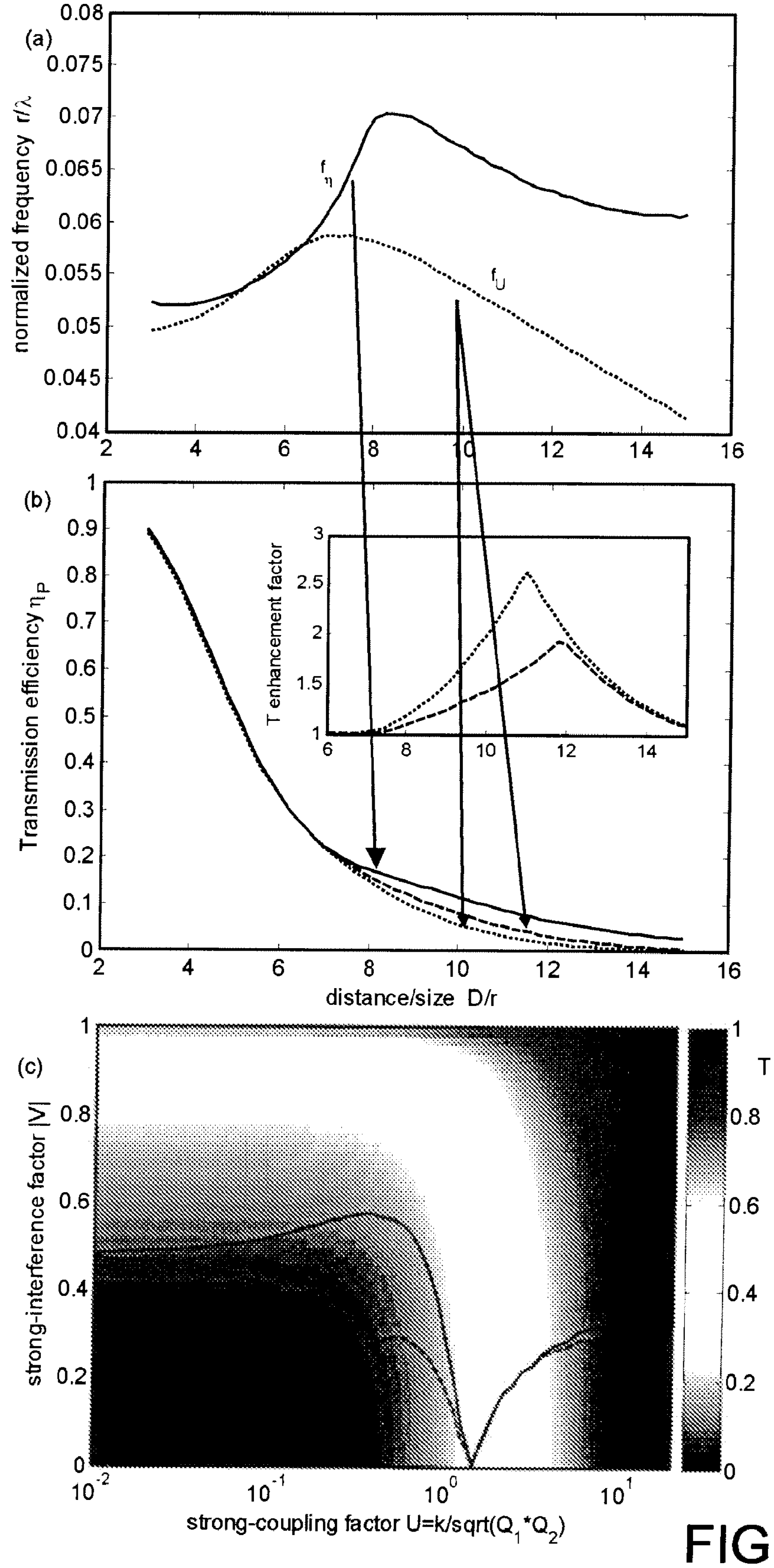
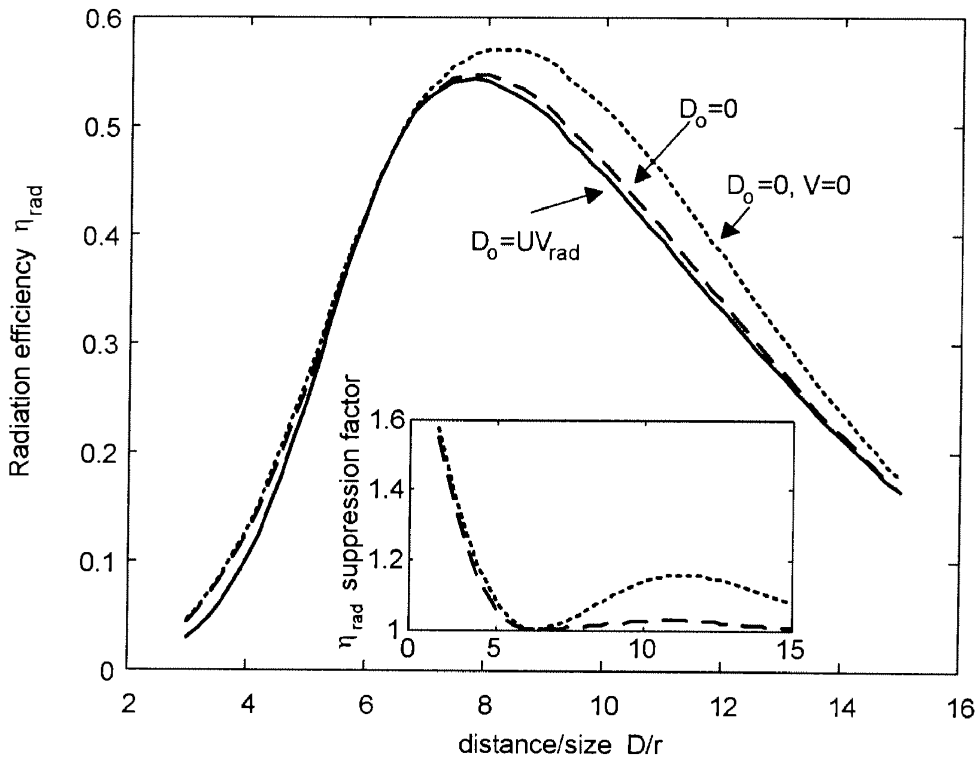

(a)
normalized frequency r/λ
0.08
0.075
0.07
0.065
0.06
0.055
0.05
0.045
0.04
2
4
6
8
10
12
14
16
fη
fU
(b)
Transmission efficiency ηP
T enhancement factor
distance/size D/r
(c)
strong-interference factor |V|
strong-coupling factor U=k/sqrt(Q1*Q2)
T

FIG. 27



0.6
0.5
0.4
0.3
0.2
0.1
0
Radiation efficiency η_{rad}
2
4
6
8
10
12
14
16
distance/size D/r
D_o=0
D_o=0, V=0
D_o=UV_{rad}
η_{rad} suppression factor
1.6
1.4
1.2
1
0
5
10
15

FIG. 28

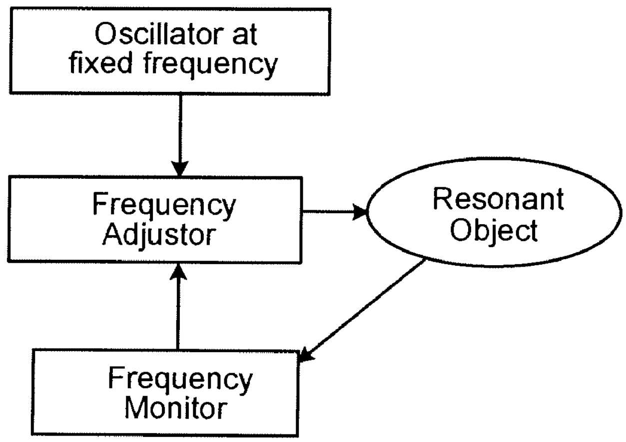


FIG. 29A

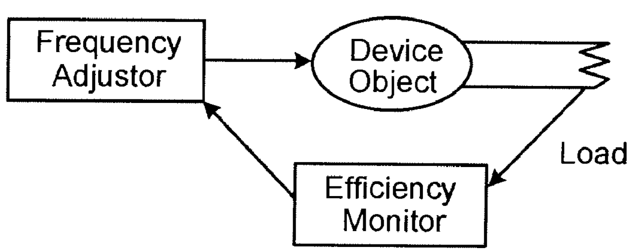
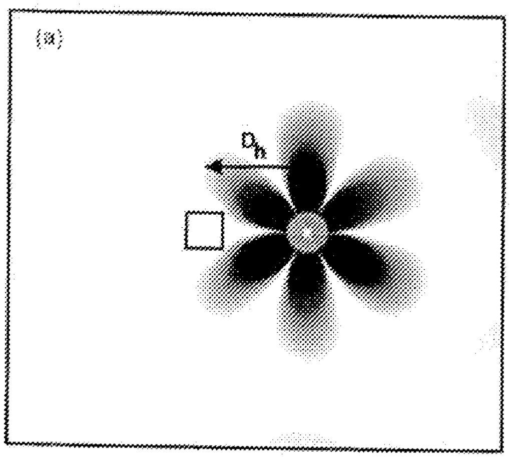
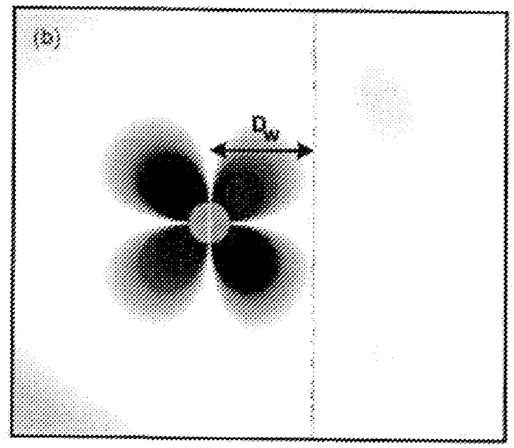


FIG. 29B



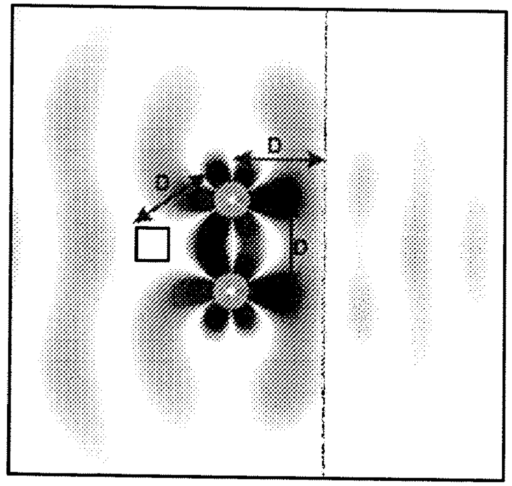
Disk with "human"	D_h/r	$Q_{c\text{-}h}^{abc}$	$Q_{c[h]}^{rad}$	$Q_{c[h]}$
Re{ε}=147.7, m=2 $\lambda/r \approx 20$ $Q_c^{abc} \approx 10096$	3	230	981	
	5	2917	1984	
	7	11573	2230	
	10	41496	2201	
Re{ε}=65.6, m=3 $\lambda/r \approx 10$ $Q_c^{abc} \approx 10096$	3	1827	6197	
	5	**58431**	**11808**	
	7	249748	9931	
	10	867552	9078	

FIG. 30A



Disk with "human"	D_w/r	$Q_{c\text{-}w}^{abc}$	$Q_{c[w]}^{rad}$	$Q_{c[w]}$
Re{ε}=147.7, m=2 $\lambda/r \approx 20$ $Q_c^{abc} \approx 10098$	3	16725	1235	1033
	5	**31659**	**1922**	**1536**
	7	49440	2389	1859
	10	82839	2140	1729
Re{ε}=65.6, m=3 $\lambda/r \approx 10$ $Q_c^{abc} \approx 10097$	3	53154	6228	3592
	5	127402	10988	5053
	7	159192	10168	4910
	10	191506	9510	4775

FIG. 30B



Two disks with "human" and "wall"	D/r	$Q^{abc}_{c\text{-}h}$	$Q^{abc}_{c\text{-}w}$	$Q^{rad}_{c[hw]}$	$Q_{w[hw]} = \omega/2\Gamma$	$\omega/2\kappa_{[hw]}$	$\kappa_{[hw]}/\Gamma_{c[hw]}$
Re{ε}=147.7, m=2	3	3300	12774	536	426	48	8.8
$\lambda/r \approx 20$	5	5719	26333	1600	1068	322	3.3
$Q^{abc}_c \approx 10100$	7	13248	50161	3542	2097	973	2.2
	10	18447	68460	3624	2254	1768	1.3
Re{ε}=65.6, m=3	3	2088	36661	6764	1328	141	9.4
$\lambda/r \approx 10$	**5**	**72137**	**90289**	**11945**	**4815**	**2114**	**2.3**
$Q^{abc}_c \approx 10100$	7	237822	129094	12261	5194	8307	0.6

FIG. 30C

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US 09/43970

A. CLASSIFICATION OF SUBJECT MATTER
IPC(8) - H01P 7/00 (2009.01)
USPC - 333/219
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC(8) - H01P 7/00 (2009.01)
USPC - 333/219

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
USPC - 333/219, 230

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)
PubWEST (PGPB,USPT,EPAB,JPAB); Google Patents; Google Scholar
Search Terms Used: resonance, resonator, energy, electricity, power, transfer, wireless, frequency, wavelength, angular, width, quality, factor, far, field, interference, generator, source, driving, drive

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X ---- Y	US 2007/0222542 A1 (JOANNOPOULOS et al.) 27 September 2007 (27.09.2007), para [0005], [0006], [0012], [0015], [0016], [0019], [0021], [0022], [0024], [0027], [0028], [0029], [0031], [0035], [0042]	1-8 ---- 9-63
Y	US 2004/0113847 A1 (QI et al.) 17 June 2004 (17.06.2004), para [0015], [0023]	9-63
Y	US 5,437,057 A (RICHLEY et al.) 25 July 1995 (25.07.1995), col 9, ln 6-18	15-25, 37-47, 57, 58, 61, 62

☐ Further documents are listed in the continuation of Box C. ☐

* Special categories of cited documents:
"A" document defining the general state of the art which is not considered to be of particular relevance
"E" earlier application or patent but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed
"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
07 July 2009 (07.07.2009)	14 JUL 2009

Name and mailing address of the ISA/US	Authorized officer:
Mail Stop PCT, Attn: ISA/US, Commissioner for Patents P.O. Box 1450, Alexandria, Virginia 22313-1450 Facsimile No. 571-273-3201	Lee W. Young PCT Helpdesk: 571-272-4300 PCT OSP: 571-272-7774

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Published:
— *with international search report (Art. 21(3))*

(54) Title: EFFICIENT NEAR-FIELD WIRELESS ENERGY TRANSFER USING ADIABATIC SYSTEM VARIATIONS

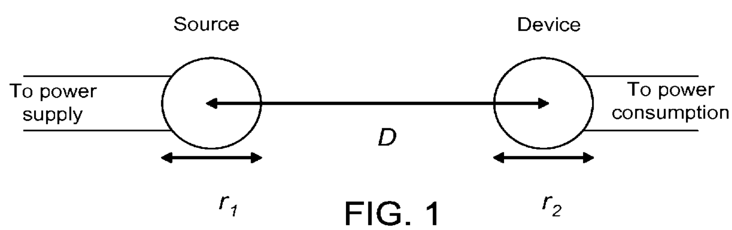


FIG. 1

(57) Abstract: Disclosed is a method for transferring energy wirelessly including transferring energy wirelessly from a first resonator structure to an intermediate resonator structure, wherein the coupling rate between the first resonator structure and the intermediate resonator structure is κ_{1B}, transferring energy wirelessly from the intermediate resonator structure to a second resonator structure, wherein the coupling rate between the intermediate resonator structure and the second resonator structure is κ_{B2} and during the wireless energy transfers, adjusting at least one of the coupling rates κ_{1B} and κ_{B2} to reduce energy accumulation in the intermediate resonator structure and improve wireless energy transfer from the first resonator structure to the second resonator structure through the intermediate resonator structure.

WO 2010/039967 A1

Attorney Docket No. 01997-0366WO1
MIT Case No 13441 (WiT MIT 4)

Efficient Near-Field Wireless Energy Transfer Using Adiabatic System Variations

CROSS REFERENCE TO RELATED APPLICATIONS

Pursuant to U.S.C. § 119(e), this application claims priority to U.S. Provisional Application Serial No. 61/101,809, filed October 1, 2008. The contents of the prior application is incorporated herein by reference in its entirety.

BACKGROUND

The disclosure relates to wireless energy transfer. Wireless energy transfer can for example, be useful in such applications as providing power to autonomous electrical or electronic devices.

Radiative modes of omni-directional antennas (which work very well for information transfer) are not suitable for such energy transfer, because a vast majority of energy is wasted into free space. Directed radiation modes, using lasers or highly-directional antennas, can be efficiently used for energy transfer, even for long distances (transfer distance $L_{TRANS} \gg L_{DEV}$, where L_{DEV} is the characteristic size of the device and/or the source), but may require existence of an uninterruptible line-of-sight and a complicated tracking system in the case of mobile objects. Some transfer schemes rely on induction, but are typically restricted to very close-range ($L_{TRANS} \ll L_{DEV}$) or low power (~mW) energy transfers.

The rapid development of autonomous electronics of recent years (e.g. laptops, cell-phones, house-hold robots, that all typically rely on chemical energy storage) has led to an increased need for wireless energy transfer.

SUMMARY

Disclosed is a method for transferring energy wirelessly. The method includes i) transferring energy wirelessly from a first resonator structure to an intermediate resonator structure, wherein the coupling rate between the first resonator structure and the intermediate resonator structure is κ_{1B}; ii) transferring energy wirelessly from the intermediate resonator structure to a second resonator structure, wherein the coupling rate

between the intermediate resonator structure and the second resonator structure is κ_{B2}; and iii) during the wireless energy transfers, adjusting at least one of the coupling rates κ_{1B} and κ_{B2} to reduce energy accumulation in the intermediate resonator structure and improve wireless energy transfer from the first resonator structure to the second resonator structure through the intermediate resonator structure.

Embodiments of the method may include one or more of the following features.

The adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} can be selected to minimize energy accumulation in the intermediate resonator structure and cause wireless energy transfer from the first resonator structure to the second resonator structure.

The adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} can be selected to maintain energy distribution in the field of the three-resonator system in an eigenstate having substantially no energy in the intermediate resonator structure. For example, the adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} can further cause the eigenstate to evolve substantially adiabatically from an initial state with substantially all energy in the resonator structures in the first resonator structure to a final state with substantially all of the energy in the resonator structures in the second resonator structure.

The adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} can be selected to include adjustments of both coupling rates κ_{1B} and κ_{B2} during wireless energy transfer.

The resonator structures can each have a quality factor larger than 10.

The first and second resonator structures can each have a quality factor greater than 50.

The first and second resonator structures can each have a quality factor greater than 100.

The resonant energy in each of the resonator structures can include electromagnetic fields. For example, the maximum value of the coupling rate κ_{1B} and the maximum value of the coupling rate κ_{B2} for inductive coupling between the intermediate resonator structure and each of the first and second resonator structures can each be larger than twice the loss rate Γ for each of the first and second resonators. Moreover, The maximum value of the coupling rate κ_{1B} and the maximum value of the coupling rate κ_{B2} for inductive coupling between the intermediate resonator structure and each of the first and second resonator structures can each be larger than four (4) times the loss rate Γ for each of the first and second resonators.

Each resonator structure can have a resonant frequency between 50 KHz and 500 MHz.

The maximum value of the coupling rate κ_{1B} and the maximum value of the coupling rate κ_{B2} can each be at least five (5) times greater than the coupling rate between the first resonator structure and the second resonator structure.

The intermediate resonator structure can have a rate of radiative energy loss that is at least twenty (20) times greater than that for either the first resonator structure or the second resonator structure.

The first and second resonator structures can be substantially identical.

The adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} can be selected to cause peak energy accumulation in the intermediate resonator structure to be less than five percent (5%) of the peak total energy in the three resonator structures.

The adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} can be selected to cause peak energy accumulation in the intermediate resonator structure during the wireless energy transfers to be less than ten percent (10%) of the peak total energy in the three resonator structures.

Adjusting at least one of the coupling rates κ_{1B} and κ_{B2} can include adjusting a relative position and/or orientation between one or more pairs of the resonator structures. Moreover, adjusting at least one of the coupling rates κ_{1B} and κ_{B2} can include adjusting a resonator property of one or more of the resonator structures, such as mutual inductance.

The resonator structures can include a capacitively loaded loop or coil of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

The resonator structures can include an inductively loaded rod of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

The wireless energy transfers are non-radiative energy transfers mediated by a coupling of a resonant field evanescent tail of the first resonator structure and a resonant field evanescent tail of the intermediate resonator structure and a coupling of the resonant field evanescent tail of the intermediate resonator structure and a resonant field evanescent tail of the second resonator structure.

The adjustment of the at least one of the coupling rates can define a first mode of operation, wherein the reduction in the energy accumulation in the intermediate resonator structure is relative to energy accumulation in the intermediate resonator structure for a second mode of operation of wireless energy transfer among the three resonator structures having a coupling rate κ'_{1B} for wireless energy transfer from the first resonator structure to the intermediate resonator structure and a coupling rate κ'_{B2} for wireless energy transfer from the intermediate resonator structure to the second resonator structure with κ'_{1B} and κ'_{B2} each being substantially constant during the second mode of wireless energy transfer, and wherein the adjustment of the coupling rates κ_{1B} and κ_{2B} in the first mode of operation can be selected to $\kappa_{1B}, \kappa_{B2} < \sqrt{\left(\kappa'^{2}_{1B} + \kappa'^{2}_{B2}\right)/2}$. Moreover, the first mode of operation can have a greater efficiency of energy transferred from the first resonator to the second resonator compared to that for the second mode of operation. Further, the first and second resonator structures can be substantially identical and each one can have a loss rate Γ_A, the intermediate resonator structure can have a loss rate Γ_B, and wherein Γ_B/Γ_A can be greater than 50.

Also, a ratio of energy lost to radiation and total energy wirelessly transferred between the first and second resonator structures in the first mode of operation is less than that for the second mode of operation. Moreover, the first and second resonator structures can be substantially identical and each one can have a loss rate Γ_A and a loss rate only due to radiation $\Gamma_{A,rad}$, the intermediate resonator structure can have a loss rate Γ_B and a loss rate only due to radiation $\Gamma_{B,rad}$ and wherein $\Gamma_{B,rad}/\Gamma_B > \Gamma_{A,rad}/\Gamma_A$.

The first mode of operation the intermediate resonator structure interacts less with extraneous objects than it does in the second mode of operation.

During the wireless energy transfer from the first resonator structure to the second resonator structure at least one of the coupling rates can be adjusted so that $\kappa_{1B} << \kappa_{B2}$ at a start of the energy transfer and $\kappa_{1B} >> \kappa_{B2}$ by a time a substantial portion of the energy has been transferred from the first resonator structure to the second resonator structure.

The coupling rate κ_{B2} can be maintained at a fixed value and the coupling rate κ_{1B} is increased during the wireless energy transfer from the first resonator structure to second resonator structure.

The coupling rate κ_{1B} can be maintained at a fixed value and the coupling rate κ_{B2} is decreased during the wireless energy transfer from the first resonator structure to second resonator structure.

During the wireless energy transfer from the first resonator structure to second resonator structure, the coupling rate κ_{1B} can be increased and the coupling rate κ_{B2} is decreased.

The method may further include features corresponding to those listed for one or more of the apparatuses and methods described below.

In another aspect, disclosed is an apparatus including: first, intermediate, and second resonator structures, wherein a coupling rate between the first resonator structure and the intermediate resonator structure is κ_{1B} and a coupling rate between the intermediate resonator structure and the second resonator structure is κ_{B2}; and means for adjusting at least one of the coupling rates κ_{1B} and κ_{B2} during wireless energy transfers among the resonator structures to reduce energy accumulation in the intermediate resonator structure and improve wireless energy transfer from the first resonator structure to the second resonator structure through the intermediate resonator structure.

Embodiments for the apparatus can include one or more of the following features.

The means for adjusting can include a rotation stage for adjusting the relative orientation of the intermediate resonator structure with respect to the first and second resonator structures.

The means for adjusting can include a translation stage for moving the first and/or second resonator structures relative to the intermediate resonator structure.

The means for adjusting can include a mechanical, electro-mechanical, or electrical staging system for dynamically adjusting the effective size of one or more of the resonator structures.

The apparatus may further include features corresponding to those listed for the method described above, and one or more of the apparatuses and methods described below.

In another aspect, a method for transferring energy wirelessly includes i): transferring energy wirelessly from a first resonator structure to a intermediate resonator structure, wherein the coupling rate between the first resonator structure and the intermediate resonator structure is κ_{1B}; ii) transferring energy wirelessly from the intermediate resonator structure to a second resonator, wherein the coupling rate between the intermediate resonator structure and the second resonator structure is κ_{B2}; and iii) during the wireless energy transfers, adjusting at least one of the coupling rates κ_{1B} and κ_{B2} to cause an energy distribution in the field of the three-resonator system to have substantially no energy in the intermediate resonator structure while wirelessly transferring energy from the first resonator structure to the second resonator structure through the intermediate resonator structure.

Embodiments for the method above can include one or more of the following features.

Having substantially no energy in the intermediate resonator structure can mean that peak energy accumulation in the intermediate resonator structure is less than ten percent (10%) of the peak total energy in the three resonator structures throughout the wireless energy transfer.

Having substantially no energy in the intermediate resonator structure can mean that peak energy accumulation in the intermediate resonator structure is less than five percent (5%) of the peak total energy in the three resonator structures throughout the wireless energy transfer.

The adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} can be selected to maintain the energy distribution in the field of the three-resonator system in an eigenstate having the substantially no energy in the intermediate resonator structure.

The adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} can be selected to further cause the eigenstate to evolve substantially adiabatically from an initial state with substantially all energy in the resonator structures in the first resonator structure to a final state with substantially all of the energy in the resonator structures in the second resonator structure.

The adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} can include adjustments of both coupling rates κ_{1B} and κ_{B2} during wireless energy transfers.

The resonant energy in each of the resonator structures comprises electromagnetic fields. For example, the maximum value of the coupling rate κ_{1B} and the maximum value of the coupling rate κ_{B2} for inductive coupling between the intermediate resonator structure and each of the first and second resonator structures can each be larger than twice the loss rate Γ for each of the first and second resonators. Moreover, the maximum value of the coupling rate κ_{1B} and the maximum value of the coupling rate κ_{B2} for inductive coupling between the intermediate resonator structure and each of the first and second resonator structures can each be larger than four (4) times the loss rate Γ for each of the first and second resonators.

The resonator structure can have a resonant frequency between 50 KHz and 500 MHz.

The maximum value of the coupling rate κ_{1B} and the maximum value of the coupling rate κ_{B2} can each be at least five (5) times greater than the coupling rate between the first resonator structure and the second resonator structure.

The intermediate resonator structure can have a rate of radiative energy loss that is at least twenty (20) times greater than that for either the first resonator structure or the second resonator structure.

The first and second resonator structures can be substantially identical.

Adjusting at least one of the coupling rates κ_{1B} and κ_{B2} can include adjusting a relative position and/or orientation between one or more pairs of the resonator structures.

Adjusting at least one of the coupling rates κ_{1B} and κ_{B2} can include adjusting a resonator property of one or more of the resonator structures, such as mutual inductance.

The resonator structures can include a capacitively loaded loop or coil of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

The resonator structures can include an inductively loaded rod of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

The wireless energy transfers can be non-radiative energy transfers mediated by a coupling of a resonant field evanescent tail of the first resonator structure and a resonant field evanescent tail of the intermediate resonator structure and a coupling of the resonant field evanescent tail of the intermediate resonator structure and a resonant field evanescent tail of the second resonator structure.

The first and second resonator structures can each have a quality factor greater than 50.

The first and second resonator structures can each have a quality factor greater than 100.

The adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} can be selected to cause the energy distribution in the field of the three-resonator system to have substantially no energy in the intermediate resonator structure improves wireless energy transfer between the first and second resonator structures.

The adjustment of the at least one of the coupling rates can be selected to define a first mode of operation, wherein energy accumulation in the intermediate resonator structure during the wireless energy transfer from the first resonator structure to second resonator structure is smaller than that for a second mode of operation of wireless energy transfer among the three resonator structures having a coupling rate κ'_{1B} for wireless energy transfer from the first resonator structure to the intermediate resonator structure and a coupling rate κ'_{B2} for wireless energy transfer from the intermediate resonator structure to the second resonator structure with κ'_{1B} and κ'_{B2} each being substantially constant during the second mode of wireless energy transfer, and wherein the adjustment of the coupling rates κ_{1B} and κ_{B2} in the first mode of operation can be selected to satisfy $\kappa_{1B}, \kappa_{B2} < \sqrt{\left(\kappa'^{2}_{1B} + \kappa'^{2}_{B2}\right)/2}$.

The first mode of operation can have a greater efficiency of energy transferred from the first resonator to the second resonator compared to that for the second mode of operation.

The first and second resonator structures can be substantially identical and each one can have a loss rate Γ_A, the intermediate resonator structure can have a loss rate Γ_B, and wherein Γ_B / Γ_A can be greater than 50.

A ratio of energy lost to radiation and total energy wirelessly transferred between the first and second resonator structures in the first mode of operation can be less than that for the second mode of operation.

The first and second resonator structures can be substantially identical and each one can have a loss rate Γ_A and a loss rate only due to radiation $\Gamma_{A,rad}$, the intermediate resonator structure can have a loss rate Γ_B and a loss rate only due to radiation $\Gamma_{B,rad}$ and wherein $\Gamma_{B,rad} / \Gamma_B > \Gamma_{A,rad} / \Gamma_A$.

The first mode of operation the intermediate resonator structure interacts less with extraneous objects than it does in the second mode of operation.

During the wireless energy transfer from the first resonator structure to the second resonator structure at least one of the coupling rates can be adjusted so that $\kappa_{1B} << \kappa_{B2}$ at a start of the energy transfer and $\kappa_{1B} >> \kappa_{B2}$ by a time a substantial portion of the energy has been transferred from the first resonator structure to the second resonator structure.

The coupling rate κ_{B2} can be maintained at a fixed value and the coupling rate κ_{1B} can be increased during the wireless energy transfer from the first resonator structure to second resonator structure.

The coupling rate κ_{1B} can be maintained at a fixed value and the coupling rate κ_{B2} can be decreased during the wireless energy transfer from the first resonator structure to second resonator structure.

During the wireless energy transfer from the first resonator structure to second resonator structure, the coupling rate κ_{1B} can be increased and the coupling rate κ_{B2} can be decreased.

The method may further include features corresponding to those listed for the apparatus and method described above, and one or more of the apparatuses and methods described below.

In another aspect, disclosed is an apparatus including: first, intermediate, and second resonator structures, wherein a coupling rate between the first resonator structure and the intermediate resonator structure is κ_{1B} and a coupling rate between the intermediate resonator structure and the second resonator structure is κ_{B2}; and means for adjusting at least one of the coupling rates κ_{1B} and κ_{B2} during wireless energy transfers among the resonator structures to cause an energy distribution in the field of the three-resonator system to have substantially no energy in the intermediate resonator structure while wirelessly transferring energy from the first resonator structure to the second resonator structure through the intermediate resonator structure.

Embodiments for the apparatus can include one or more of the following features.

Having substantially no energy in the intermediate resonator structure can mean that peak energy accumulation in the intermediate resonator structure is less than ten percent (10%) of the peak total energy in the three resonator structures throughout the wireless energy transfers.

Having substantially no energy in the intermediate resonator structure can mean that peak energy accumulation in the intermediate resonator structure is less than five percent (5%) of the peak total energy in the three resonator structures throughout the wireless energy transfers.

The means for adjusting can be configured to maintain the energy distribution in the field of the three-resonator system in an eigenstate having the substantially no energy in the intermediate resonator structure.

The means for adjusting can include a rotation stage for adjusting the relative orientation of the intermediate resonator structure with respect to the first and second resonator structures.

The means for adjusting can include a translation stage for moving the first and/or second resonator structures relative to the intermediate resonator structure.

The means for adjusting can include a mechanical, electro-mechanical, or electrical staging system for dynamically adjusting the effective size of one or more of the resonator structures.

The resonator structures can include a capacitively loaded loop or coil of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

The resonator structures can include an inductively loaded rod of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

A source can be coupled to the first resonator structure and a load can be coupled to the second resonator structure.

The apparatus may further include features corresponding to those listed for the apparatus and methods described above, and the apparatus and method described below.

In another aspect, disclosed is a method for transferring energy wirelessly that includes: i) transferring energy wirelessly from a first resonator structure to a intermediate resonator structure, wherein the coupling rate between the first resonator structure and the intermediate resonator structure is κ_{1B}; ii) transferring energy wirelessly from the intermediate resonator structure to a second resonator, wherein the coupling rate between the intermediate resonator structure and the second resonator structure with a coupling rate is κ_{B2}; and iii) during the wireless energy transfers, adjusting at least one of the coupling rates κ_{1B} and κ_{B2} to define a first mode of operation in which energy accumulation in the intermediate resonator structure is reduced relative to that for a second mode of operation of wireless energy transfer among the three resonator structures having a coupling rate κ'_{1B} for wireless energy transfer from the first resonator structure to the intermediate resonator structure and a coupling rate κ'_{B2} for wireless energy transfer from the intermediate resonator structure to the second resonator structure with κ'_{1B} and κ'_{B2} each being substantially constant during the second mode of wireless energy transfer, and wherein the adjustment of the coupling rates κ_{1B} and κ_{B2} in the first mode of operation can be selected to satisfy $\kappa_{1B}, \kappa_{B2} < \sqrt{\left(\kappa'^{2}_{1B} + \kappa'^{2}_{B2}\right)/2}$.

The method may further include features corresponding to those listed for the apparatuses and methods described above.

In another aspect, disclosed is an apparatus that includes: first, intermediate, and second resonator structures, wherein a coupling rate between the first resonator structure and the intermediate resonator structure is κ_{1B} and a coupling rate between the intermediate resonator structure and the second resonator structure is κ_{B2}; and means for adjusting at least one of the coupling rates κ_{1B} and κ_{B2} during wireless energy transfers among the resonator structures to define a first mode of operation in which energy accumulation in the intermediate resonator structure is reduced relative to that for a second mode of operation for wireless energy transfer among the three resonator structures having a coupling rate κ'_{1B} for wireless energy transfer from the first resonator structure to the intermediate resonator structure and a coupling rate κ'_{B2} for wireless energy transfer from the intermediate resonator structure to the second resonator structure with κ'_{1B} and κ'_{B2} each being substantially constant during the second mode of wireless energy transfer, and wherein the adjustment of the coupling rates κ_{12} and κ_{B2} in the first mode of operation can be selected to satisfy $\kappa_{1B}, \kappa_{B2} < \sqrt{\left(\kappa'^{2}_{1B} + \kappa'^{2}_{B2}\right)/2}$.

The apparatus may further include features corresponding to those listed for the apparatuses and methods described above.

Unless otherwise defined, all technical and scientific terms used herein have the same meaning as commonly understood by one of ordinary skill in the art. Although methods and materials similar or equivalent to those described herein can be used in the practice or testing of the present disclosure, suitable methods and materials are described below. All publications, patent applications, patents, and other references mentioned herein are incorporated by reference in their entirety. In case of conflict, the present specification, including definitions, will control. In addition, the materials, methods, and examples are illustrative only and not intended to be limiting.

The details of one or more embodiments are set forth in the accompanying drawings and the description below, including the documents appended hereto. Other features and advantages will be apparent from this disclosure and from the claims.

BRIEF DESCRIPTION OF THE DRAWINGS

Fig. 1 shows a schematic of an example wireless energy transfer scheme.

Figs. 2(a)-(b) show the efficiency of power transmission η_{P} for (a) $U=1$ and (b) $U=3$, as a function of the frequency detuning D_o and for different values of the loading rate U_o.

Fig. 2(c) shows the optimal (for zero detuning and under conditions of impedance matching) efficiency for energy transfer $\eta_{\mathrm{E}*}$ and power transmission $\eta_{\mathrm{P}*}$, as a function of the coupling-to-loss figure-of-merit U.

Fig. 3 shows an example of a self-resonant conducting-wire coil.

Fig. 4 shows an example of a wireless energy transfer scheme featuring two self-resonant conducting-wire coils.

Fig. 5 is a schematic of an experimental system demonstrating wireless energy transfer.

Fig. 6 shows a comparison between experimental and theoretical results for the coupling rate of the system shown schematically in Fig. 5.

Fig. 7 shows a comparison between experimental and theoretical results for the strong-coupling factor of the system shown schematically in Fig. 5.

Fig. 8 shows a comparison between experimental and theoretical results for the power-transmission efficiency of the system shown schematically in Fig. 5.

Fig. 9 shows an example of a capacitively-loaded conducting-wire coil, and illustrates the surrounding field.

Fig. 10 shows an example wireless energy transfer scheme featuring two capacitively-loaded conducting-wire coils, and illustrates the surrounding field.

Fig. 11 illustrates an example circuit model for wireless energy transfer.

Fig. 12 shows the efficiency, total (loaded) device Q, and source and device currents, voltages and radiated powers (normalized to 1Watt of output power to the load) as functions of the resonant frequency, for a particular choice of source and device loop dimensions, wp and N_s and different choices of N_d=1,2,3,4,5,6,10.

Fig.13 shows the efficiency, total (loaded) device Q, and source and device currents, voltages and radiated powers (normalized to 1Watt of output power to the load) as functions of frequency and wp for a particular choice of source and device loop dimensions, and number of turns N_s and N_d.

Fig. 14 shows an example of an inductively-loaded conducting-wire coil.

Fig. 15 shows (a) an example of a resonant dielectric disk, and illustrates the surrounding field and (b) a wireless energy transfer scheme featuring two resonant dielectric disks, and illustrates the surrounding field.

Fig. 16 shows a schematic of an example wireless energy transfer scheme with one source resonator and one device resonator exchanging energy indirectly through an intermediate resonator.

Fig. 17 shows an example of a wireless energy transfer system: (a) (Left) Schematic of loops configuration in two-object direct transfer. (Right) Time evolution of energies in the two-object direct energy transfer case. (b) (Left) Schematic of three-loops configuration in the constant-κ case. (Right) Dynamics of energy transfer for the configuration in (b. Left). Note that the total energy transferred E_2 is 2 times larger than in (a. Right), but at the price of the total energy radiated being 4 times larger. (c) (Left)

Loop configuration at t=0 in the adiabatic-κ scheme. (Center) Dynamics of energy transfer with adiabatically rotating loops. (Right) Loop configuration at $t=t_{EIT}$. Note that E_2 is comparable to (b. Right), but the radiated energy is now much smaller: In fact, it is comparable to (a. Right).

Fig. 18 shows a schematic of an example wireless energy transfer scheme with one source resonator and one device resonator exchanging energy indirectly through an intermediate resonator, where an adjustment system is used to rotate the resonator structures to dynamically adjust their coupling rates.

Fig. 19 shows an example of a temporal variation of the coupling rates in a wireless energy transfer system as in Fig. 18 to achieve an adiabatic transfer of energy from the source object R_1 to the device object R_2.

Fig. 20 shows the energy distribution in a wireless energy transfer system as in Fig. 18 as a function of time when the coupling rates are time-varying, for $\Gamma_A=0$, $\kappa/\Gamma_B=10$, $\kappa_{1B}=\kappa\sin[\pi t/(2t_{EIT})]$, and $\kappa_{B2}=\kappa\cos[\pi t/(2t_{EIT})]$.

Figs. 21(a)-(f) show a comparison between the adiabatic-κ and constant-κ energy transfer schemes, in the general case: (a) Optimum E_2 (%) in adiabatic-κ transfer, (b) Optimum E_2 (%) in constant-κ transfer, (c) $(E_2)_{\text{adiabatic-}\kappa}/(E_2)_{\text{constant-}\kappa}$, (d) Energy lost (%) at optimum adiabatic-κ transfer, (e) Energy lost (%) at optimum constant-κ transfer, (f) $(E_{lost})_{\text{constant-}\kappa}/(E_{lost})_{\text{adiabatic-}\kappa}$.

Fig. 22(a)-(e) show a comparison between radiated energies in the adiabatic-κ and constant-κ energy transfer schemes: (a) E_{rad}(%) in the constant-scheme for $\Gamma_B/\Gamma_A=500$ and $\Gamma_{rad}^{A}=0$, (b) E_{rad}(%) in the adiabatic-κ scheme for $\Gamma_B/\Gamma_A=500$ and $\Gamma_{rad}^{A}=0$, (c) $(E_{rad})_{\text{constant-}\kappa}/(E_{rad})_{\text{adiabatic-}\kappa}$ for $\Gamma_B/\Gamma_A=50$, (d) $(E_{rad})_{\text{constant-}\kappa}/(E_{rad})_{\text{adiabatic-}\kappa}$ for $\Gamma_B/\Gamma_A=500$, (e) $[(E_{rad})_{\text{constant-}\kappa}/(E_{rad})_{\text{adiabatic-}\kappa}]$ as function of κ/Γ_B and Γ_B/Γ_A, for $\Gamma_{rad}^{A}=0$.

Figs. 23(a)-(b) show schematics for frequency control mechanisms.

Figs. 24(a)-(c) illustrate a wireless energy transfer scheme using two dielectric disks in the presence of various extraneous objects.

DETAILED DESCRIPTION

Efficient wireless energy-transfer between two similar-frequency resonant objects can be achieved at mid-range distances, provided these resonant objects are designed to operate in the 'strong-coupling' regime. 'Strong coupling' can be realized for a wide variety of resonant objects, including electromagnetic resonant objects such as inductively-loaded conducting rods and dielectric disks. Recently, we have demonstrated wireless energy transfer between strongly coupled electromagnetic self-resonant conducting coils and capacitively-loaded conducting coils, bearing high-Q electromagnetic resonant modes. See, for example, the following commonly owned U.S. Patent Applications, all of which are incorporated herein by reference: U.S. Application Serial No. 11/481,077, filed on July 5, 2006, and published as U.S. Patent Publication No. US 2007-0222542 A1; U.S. Application Serial No. 12/055,963, filed on March 26, 2008, and published as U.S. Patent Publication No. US 2008-0278264 A1; and U.S. Patent Application Serial No. 12/466,065, filed on May 14, 2009, and published as U.S. Patent Publication No. __________. In general, the energy-transfer efficiency between similar-frequency, strongly coupled resonant objects decreases as the distance between the objects is increased.

In this work, we explore a further scheme of efficient energy transfer between resonant objects that extends the range over which energy may be efficiently transferred. Instead of transferring energy directly between two resonant objects, as has been described in certain embodiments of the cross-referenced patents, in certain embodiments, an intermediate resonant object, with a resonant frequency equal or nearly-equal to that of the two energy-exchanging resonant objects is used to mediate the transfer. The intermediate resonant object may be chosen so that it couples more strongly to each of the resonant objects involved in the energy transfer than those two resonant objects couple to each other. One way to design such an intermediate resonator is to make it larger than either of the resonant objects involved in the energy transfer. However, increasing the size of the intermediate resonant object may lower its quality factor, or Q, by increasing its radiation losses. Surprisingly enough, this new "indirect" energy transfer scheme may be shown to be very efficient and only weakly-radiative by introducing a meticulously chosen time variation of the resonator coupling rates.

The advantage of this method over the prior commonly owned wireless energy transfer techniques is that, in certain embodiments, it can enable energy to be transferred wirelessly between two objects with a larger efficiency and/or with a smaller radiation loss and/or with fewer interactions with extraneous objects.

Accordingly, in certain embodiments, we disclose an efficient wireless energy transfer scheme between two similar resonant objects, strongly coupled to an intermediate resonant object of substantially different properties, but with the same resonance frequency. The transfer mechanism essentially makes use of the adiabatic evolution of an instantaneous (so called 'dark') resonant state of the coupled three-object system. Our analysis is based on temporal coupled mode theory (CMT), and is general. Of particular commercial interest is the application of this technique to strongly-coupled electromagnetic resonators used for mid-range wireless energy transfer applications. We show that in certain parameter regimes of interest, this scheme can be more efficient, and/or less radiative than other wireless energy transfer approaches.

While the technique described herein is primarily directed to tangible resonator structures, the technique shares certain features with a quantum interference phenomenon known in the atomic physics community as Electromagnetically Induced Transparency (EIT). In EIT, three atomic states participate. Two of them, which are non-lossy, are coupled to one that has substantial losses. However, by meticulously controlling the mutual couplings between the states, one can establish a coupled system which is overall non-lossy. This phenomena has been demonstrated using carefully timed optical pulses, referred to as probe laser pulses and Stokes laser pulses, to reduce the opacity of media with the appropriate collection of atomic states. A closely related phenomenon known as Stimulated Raman Adiabatic Passage (STIRAP) may take place in a similar system; namely, the probe and Stokes laser beams may be used to achieve complete coherent population transfer between two molecular states of a medium. Hence, we may refer to the currently proposed scheme as the "EIT-like" energy transfer scheme.

In certain embodiments, we disclose an efficient near-field energy transfer scheme between two similar resonant objects, based on an EIT-like transfer of the energy through a mediating resonant object with the same resonant frequency. In embodiments, this EIT-like energy transfer may be realized using electromagnetic resonators as have been described in the cross-referenced patents, but the scheme is not bound only to wireless energy transfer applications. Rather, this scheme is general and may find applications in various other types of coupling between general resonant objects. In certain embodiments described below, we describe particular examples of electromagnetic resonators, but the nature of the resonators and their coupling mechanisms could be quite different (e.g. acoustic, mechanical, etc.). To the extent that many resonant phenomena can be modeled with nearly identical CMT equations, similar behavior to that described herein would occur.

1. Efficient energy-transfer by two 'strongly coupled' resonances

Fig. 1 shows a schematic that generally describes one example of the invention, in which energy is transferred wirelessly between two resonant objects. Referring to Fig. 1, energy is transferred over a distance D, between a resonant source object having a characteristic size r_1 and a resonant device object of characteristic size r_2. Both objects are resonant objects. The wireless near-field energy transfer is performed using the field (e.g. the electromagnetic field or acoustic field) of the system of two resonant objects.

The characteristic size of an object can be regarded as being equal to the radius of the smallest sphere which can fit around the entire object. The characteristic thickness of an object can be regarded as being, when placed on a flat surface in any arbitrary configuration, the smallest possible height of the highest point of the object above a flat surface. The characteristic width of an object can be regarded as being the radius of the smallest possible circle that the object can pass through while traveling in a straight line. For example, the characteristic width of a cylindrical object is the radius of the cylinder.

Initially, we present a theoretical framework for understanding near-field wireless energy transfer. Note however that it is to be understood that the scope of the invention is not bound by theory.

Different temporal schemes can be employed, depending on the application, to transfer energy between two resonant objects. Here we will consider two particularly simple but important schemes: a one-time finite-amount energy-transfer scheme and a continuous finite-rate energy-transfer (power) scheme.

1.1 Finite-amount energy-transfer efficiency

Let the source and device objects be 1, 2 respectively and their resonance modes, which we will use for the energy exchange, have angular frequencies $\omega_{1,2}$, frequency-widths due to intrinsic (absorption, radiation etc.) losses $\Gamma_{1,2}$ and (generally) vector fields $\mathbf{F}_{1,2}(\mathbf{r})$, normalized to unity energy. Once the two resonant objects are brought in proximity, they can interact and an appropriate analytical framework for modeling this resonant interaction is that of the well-known coupled-mode theory (CMT). This model works well, when the resonances are well defined by having large quality factors and their resonant frequencies are relatively close to each other. In this picture, the field of the system of the two resonant objects 1, 2 can be approximated by $\mathbf{F}(\mathbf{r},t)=a_1(t)\mathbf{F}_1(\mathbf{r})+a_2(t)\mathbf{F}_2(\mathbf{r})$, where $a_{1,2}(t)$ are the field amplitudes, with $|a_{1,2}(t)|^2$ equal to the energy stored inside the object 1, 2 respectively, due to the normalization. Then, using $e^{-i\omega t}$ time dependence, the field amplitudes can be shown to satisfy, to lowest order:

$$\begin{aligned}\frac{d}{dt}a_1(t)&=-i(\omega_1-i\Gamma_1)a_1(t)+i\kappa_{11}a_1(t)+i\kappa_{12}a_2(t)\\ \frac{d}{dt}a_2(t)&=-i(\omega_2-i\Gamma_2)a_2(t)+i\kappa_{21}a_1(t)+i\kappa_{22}a_2(t)\end{aligned} \quad (1)$$

where $\kappa_{11,22}$ are the shifts in each object's frequency due to the presence of the other, which are a second-order correction and can be absorbed into the resonant frequencies (eigenfrequencies) by setting $\omega_{1,2}\to\omega_{1,2}+\kappa_{11,22}$, and $\kappa_{12,21}$ are the coupling coefficients, which from the reciprocity requirement of the system satisfy $\kappa_{21}=\kappa_{12}\equiv\kappa$.

The resonant modes of the combined system are found by substituting $[a_1(t), a_2(t)] = [A_1, A_2]e^{-i\bar{\omega}t}$. They have complex resonant frequencies

$$\bar{\omega}_{\pm} = \omega_{12} \pm \sqrt{(\Delta\omega_{12})^2 + \kappa^2} \tag{2a}$$

where $\omega_{12} = [(\omega_1 + \omega_2) - i(\Gamma_1 + \Gamma_2)]/2$, $\Delta\omega_{12} = [(\omega_1 - \omega_2) - i(\Gamma_1 - \Gamma_2)]/2$ and whose splitting we denote as $\delta_E \equiv \bar{\omega}_+ - \bar{\omega}_-$, and corresponding resonant field amplitudes

$$\vec{V}_{\pm} = \begin{bmatrix} A_1 \\ A_2 \end{bmatrix}_{\pm} = \begin{bmatrix} \kappa \\ \Delta\omega_{12} \mp \sqrt{(\Delta\omega_{12})^2 + \kappa^2} \end{bmatrix}. \tag{2b}$$

Note that, at exact resonance $\omega_1 = \omega_2 = \omega_A$ and for $\Gamma_1 = \Gamma_2 = \Gamma_A$, we get $\Delta\omega_{12} = 0$, $\delta_E = 2\kappa$, and then

$$\bar{\omega}_{\pm} = \omega_A \pm \kappa - i\Gamma_A$$

$$\vec{V}_{\pm} = \begin{bmatrix} A_1 \\ A_2 \end{bmatrix}_{\pm} = \begin{bmatrix} 1 \\ \mp 1 \end{bmatrix},$$

namely we get the known result that the resonant modes split to a lower frequency even mode and a higher frequency odd mode.

Assume now that at time $t = 0$ the source object 1 has finite energy $|a_1(0)|^2$, while the device object has $|a_2(0)|^2 = 0$. Since the objects are coupled, energy will be transferred from 1 to 2. With these initial conditions, Eqs.(1) can be solved, predicting the evolution of the device field-amplitude to be

$$\frac{a_2(t)}{|a_1(0)|} = \frac{2\kappa}{\delta_E} \sin\left(\frac{\delta_E t}{2}\right) e^{-\frac{\Gamma_1+\Gamma_2}{2}t} e^{-i\frac{\omega_1+\omega_2}{2}t}. \tag{3}$$

The energy-transfer efficiency will be $\eta_E \equiv |a_2(t)|^2/|a_1(0)|^2$. The ratio of energy converted to loss due to a specific loss mechanism in resonators 1 and 2, with respective loss rates $\Gamma_{1,\text{loss}}$ and $\Gamma_{2,\text{loss}}$ will be $\eta_{loss,E} = \int_0^t d\tau\left[2\Gamma_{1,\text{loss}}|a_1(\tau)|^2 + 2\Gamma_{2,\text{loss}}|a_2(\tau)|^2\right]/|a_1(0)|^2$. Note that, at exact resonance $\omega_1 = \omega_2 = \omega_A$ (an optimal condition), Eq.(3) can be written as

$$\frac{a_2(T)}{|a_1(0)|} = \frac{\sin\left(\sqrt{1-\Delta^2}T\right)}{\sqrt{1-\Delta^2}} e^{-T/U} e^{-i\omega_A t} \tag{4}$$

where $\equiv \kappa t$, $\Delta^{-1} = 2\kappa/(\Gamma_2 - \Gamma_1)$ and $U = 2\kappa/(\Gamma_1 + \Gamma_2)$.

In some examples, the system designer can adjust the duration of the coupling t at will. In some examples, the duration t can be adjusted to maximize the device energy (and thus efficiency η_E). Then, it can be inferred from Eq.(4) that η_E is maximized for

$$T_* = \frac{\tan^{-1}\left(U\sqrt{1-\Delta^2}\right)}{\sqrt{1-\Delta^2}} \tag{5}$$

resulting in an optimal energy-transfer efficiency

$$\eta_{E*} \equiv \eta_E(T_*) = \frac{U^2}{1+U^2\left(1-\Delta^2\right)} \exp\left(-\frac{2\tan^{-1}\left(U\sqrt{1-\Delta^2}\right)}{U\sqrt{1-\Delta^2}}\right). \tag{6a}$$

which is a monotonically increasing function of the coupling-to-loss ratio $U = 2\kappa/(\Gamma_1 + \Gamma_2)$ and tends to unity when $U \gg 1 \Longrightarrow |\Delta|^{-1} \gg 1$. Therefore, the energy transfer is nearly perfect, when the coupling rate is much faster than all loss rates ($\kappa/\Gamma_{1,2} \gg 1$). In Fig.2(c) we show the optimal energy-transfer efficiency when $\Gamma_1 = \Gamma_2 = \Gamma_A \Leftrightarrow \Delta = 0$:

$$\eta_E(T_*, \Delta = 0) = \frac{U^2}{1+U^2} \exp\left(-\frac{2\tan^{-1}U}{U}\right). \tag{6b}$$

In a real wireless energy-transfer system, the source object can be connected to a power generator (not shown in Fig.1), and the device object can be connected to a power consuming load (e.g. a resistor, a battery, an actual device, not shown in Fig.1). The generator will supply the energy to the source object, the energy will be transferred wirelessly and non-radiatively from the source object to the device object, and the load will consume the energy from the device object. To incorporate such supply and consumption mechanisms into this temporal scheme, in some examples, one can imagine that the generator is very briefly but very strongly coupled to the source at time $t = 0$ to almost instantaneously provide the energy, and the load is similarly very briefly but very strongly coupled to the device at the optimal time $t = t_*$ to almost instantaneously drain the energy. For a constant powering mechanism, at time $t = t_*$ also the generator can again be coupled to the source to feed a new amount of energy, and this process can be repeated periodically with a period t_*.

1.2 Finite-rate energy-transfer (power-transmission) efficiency

Let the generator be continuously supplying energy to the source object 1 at a rate κ_1 and the load continuously draining energy from the device object 2 at a rate κ_2. Field amplitudes $s_{\pm1,2}(t)$ are then defined, so that $\left|s_{\pm1,2}(t)\right|^2$ is equal to the power ingoing to (for the + sign) or outgoing from (for the - sign) the object 1, 2 respectively, and the CMT equations are modified to

$$\begin{aligned}
\frac{d}{dt}a_1(t) &= -i(\omega_1 - i\Gamma_1)a_1(t) + i\kappa_{11}a_1(t) + i\kappa_{12}a_2(t) - \kappa_1 a_1(t) + \sqrt{2\kappa_1}s_{+1}(t)\\
\frac{d}{dt}a_2(t) &= -i(\omega_2 - i\Gamma_2)a_2(t) + i\kappa_{21}a_1(t) + i\kappa_{22}a_2(t) - \kappa_2 a_2(t)\\
s_{-1}(t) &= \sqrt{2\kappa_1}a_1(t) - s_{+1}(t)\\
s_{-2}(t) &= \sqrt{2\kappa_2}a_2(t)
\end{aligned} \tag{7}$$

where again we can set $\omega_{1,2} \to \omega_{1,2} + \kappa_{11,22}$ and $\kappa_{21} = \kappa_{12} \equiv \kappa$.

Assume now that the excitation is at a fixed frequency ω, namely has the form $s_{+1}(t) = S_{+1}e^{-i\omega t}$. Then the response of the linear system will be at the same frequency, namely $a_{1,2}(t) = A_{1,2}e^{-i\omega t}$ and $s_{-1,2}(t) = S_{-1,2}e^{-i\omega t}$. By substituting these into Eqs.(7), using $\delta_{1,2} \equiv \omega - \omega_{1,2}$, and solving the system, we find the field-amplitude transmitted to the load (S_{21} scattering-matrix element)

$$S_{21} \equiv \frac{S_{-2}}{S_{+1}} = \frac{2i\kappa\sqrt{\kappa_1\kappa_2}}{(\Gamma_1+\kappa_1-i\delta_1)(\Gamma_2+\kappa_2-i\delta_2)+\kappa^2} = \frac{2iU\sqrt{U_1U_2}}{(1+U_1-iD_1)(1+U_2-iD_2)+U^2} \tag{8}$$

and the field-amplitude reflected to the generator (S_{11} scattering-matrix element)

$$S_{11} \equiv \frac{S_{-1}}{S_{+1}} = \frac{(\Gamma_1-\kappa_1-i\delta_1)(\Gamma_2+\kappa_2-i\delta_2)+\kappa^2}{(\Gamma_1+\kappa_1-i\delta_1)(\Gamma_2+\kappa_2-i\delta_2)+\kappa^2} = \frac{(1-U_1-iD_1)(1+U_2-iD_2)+U^2}{(1+U_1-iD_1)(1+U_2-iD_2)+U^2} \tag{9}$$

where $D_{1,2} \equiv \delta_{1,2}/\Gamma_{1,2}$, $U_{1,2} \equiv \kappa_{1,2}/\Gamma_{1,2}$ and $U \equiv \kappa/\sqrt{\Gamma_1\Gamma_2}$. Similarly, the scattering-matrix elementsS_{12}, S_{22} are given by interchanging 1 $\leftrightarrow$ 2 in Eqs.(8),(9) and, as expected from reciprocity, $S_{21} = S_{12}$. The coefficients for power transmission (efficiency) and reflection and loss are respectively $\eta_P \equiv |S_{21}|^2 = |S_{-2}|^2/|S_{+1}|^2$ and $|S_{11}|^2 = |S_{-1}|^2/|S_{+1}|^2$ and $1 - |S_{21}|^2 - |S_{11}|^2 = (2\Gamma_1|A_1|^2 + 2\Gamma_2|A_2|^2)/|S_{+1}|^2$.

In some implementations, the parameters $D_{1,2}$, $U_{1,2}$ can be designed (engineered), since one can adjust the resonant frequencies $\omega_{1,2}$ (compared to the desired operating frequency ω) and the generator/load supply/drain rates $\kappa_{1,2}$. Their choice can target the optimization of some system performance-characteristic of interest.

In some examples, a goal can be to maximize the power transmission (efficiency) $\eta_P \equiv |S_{21}|^2$ of the system, so one would require

$$\eta_P'(D_{1,2}) = \eta_P'(U_{1,2}) = 0 \tag{10}$$

Since S_{21} (from Eq.(8)) is symmetric upon interchanging 1 $\leftrightarrow$ 2, the optimal values for $D_{1,2}$ (determined by Eqs.(10)) will be equal, namely $D_1 = D_2 \equiv D_o$, and similarly $U_1 = U_2 \equiv U_o$. Then,

$$S_{21} = \frac{2iUU_o}{(1+U_o-iD_o)^2+U^2} \tag{11}$$

and from the condition $\eta_P'(D_o) = 0$ we get that, for fixed values of U and U_o, the efficiency can be maximized for the following values of the symmetric detuning

$$D_o = \begin{matrix} \pm\sqrt{U^2-(1+U_o)^2}, & \text{if} \quad U > 1+U_o \\ 0, & \text{if} \quad U \le 1+U_o \end{matrix}, \tag{12}$$

which, in the case $U > 1 + U_o$, can be rewritten for the two frequencies at which the efficiency peaks as

$$\omega_\pm = \frac{\omega_1\Gamma_2+\omega_2\Gamma_1}{\Gamma_1+\Gamma_2} \pm \frac{2\sqrt{\Gamma_1\Gamma_2}}{\Gamma_1+\Gamma_2}\sqrt{\kappa^2-(\Gamma_1+\kappa_1)(\Gamma_2+\kappa_2)}, \tag{13}$$

whose splitting we denote as $\delta_P \equiv \bar{\omega}_+ - \bar{\omega}_-$. Note that, at exact resonance $\omega_1 = \omega_2$, and for $\Gamma_1 = \Gamma_2 \equiv \Gamma_o$ and $\kappa_1 = \kappa_2 \equiv \kappa_o$, we get $\delta_P = 2\sqrt{\kappa^2 - (\Gamma_o + \kappa_o)^2} < \delta_E$, namely the transmission-peak splitting is smaller than the normal-mode splitting. Then, by substituting D_o into η_P from Eq.(12), from the condition $\eta'_P(U_o) = 0$ we get that, for fixed value of U, the efficiency can be maximized for

$$U_{o*} = \sqrt{1+U^2} \overset{\text{Eq.(12)}}{\Rightarrow} D_{o*} = 0 \tag{14}$$

which is known as 'critical coupling' condition, whereas for $U_o < U_{o*}$ the system is called 'undercoupled' and for $U_o > U_{o*}$ it is called 'overcoupled'. The dependence of the efficiency on the frequency detuning D_o for different values of U_o (including the 'critical-coupling' condition) are shown in Fig. 2(a,b). The overall optimal power efficiency using Eqs.(14) is

$$\eta_{P*} \equiv \eta_P(D_{o*}, U_{o*}) = \frac{U_{o*}-1}{U_{o*}+1} = \left(\frac{U}{1+\sqrt{1+U^2}}\right)^2, \tag{15}$$

which is again only a function of the coupling-to-loss ratio $U = \kappa/\sqrt{\Gamma_1\Gamma_2}$ and tends to unity when $U \gg 1$, as depicted in Fig. 2(c).

In some examples, a goal can be to minimize the power reflection at the side of the generator $|S_{11}|^2$ and the load $|S_{22}|^2$, so one would then need

$$S_{11,22} = 0 \Rightarrow (1 \mp U_1 - iD_1)(1 \pm U_2 - iD_2) + U^2 = 0, \tag{16}$$

The equations above present ‘impedance matching’ conditions. Again, the set of these conditions is symmetric upon interchanging $1 \leftrightarrow 2$, so, by substituting $D_1 = D_2 \equiv D_o$ and $U_1 = U_2 \equiv U_o$ into Eqs.(16), we get

$$(1-iD_o)^2 - U_o^2 + U^2 = 0, \tag{17}$$

from which we easily find that the values of D_o and U_o that cancel all reflections are again exactly those in Eqs.(14).

It can be seen that, the two goals and their associated sets of conditions (Eqs.(10) and Eqs.(16)) result in the same optimized values of the intra-source and intra-device parameters $D_{1,2}$, $U_{1,2}$. Note that for a lossless system this would be an immediate consequence of power conservation (Hermiticity of the scattering matrix), but this is not apparent for a lossy system.

Accordingly, for any temporal energy-transfer scheme, once the parameters specific only to the source or to the device (such as their resonant frequencies and their excitation or loading rates respectively) have been optimally designed, the efficiency monotonically increases with the ratio of the source-device coupling-rate to their loss rates. Using the definition of a resonance quality factor $Q = \omega/2\Gamma$ and defining by analogy the coupling factor $k \equiv 1/Q_\kappa \equiv 2\kappa/\sqrt{\omega_1\omega_2}$, it is therefore exactly this ratio

$$U = \frac{\kappa}{\sqrt{\Gamma_1\Gamma_2}} = k\sqrt{Q_1Q_2} \tag{18}$$

that has been set as a figure-of-merit for any system under consideration for wireless energy-transfer, along with the distance over which this ratio can be achieved (clearly, U will be a decreasing function of distance). The operating regime $U > 1$ is sometimes called ‘strong-coupling’ regime and is a sufficient condition for efficient energy-transfer. In particular, for $U > 1$ we get, from Eq.(15), $\eta_{P*} > 17\%$, large enough for many practical applications. Note that in some applications, U>0.1 may be sufficient. In applications where it is impossible or impractical to run wires to supply power to a device, U<0.1 may be considered sufficient. One skilled in the art will recognize that the sufficient U is application and specification dependent. The figure-of-merit U may be called the strong-coupling factor. We will further show how to design systems with a large strong-coupling factor.

To achieve a large strong-coupling factor U, in some examples, the energy-transfer application preferably uses resonant modes of high quality factors Q, corresponding to low (i.e. slow) intrinsic-loss rates Γ. This condition can be satisfied by designing resonant modes where all loss mechanisms, typically radiation and absorption, are sufficiently suppressed.

This suggests that the coupling be implemented using, not the lossy radiative far-field, which should rather be suppressed, but the evanescent (non-lossy) stationary near-field. To implement an energy-transfer scheme, usually more appropriate are finite objects, namely ones that are topologically surrounded everywhere by air, into where the near field extends to achieve the coupling. Objects of finite extent do not generally support electromagnetic states that are exponentially decaying in *all* directions in air away from the objects, since Maxwell's Equations in free space imply that $\mathbf{k}^2=\omega^2/c^2$, where $\mathbf{k}$ is the wave vector, ω the angular frequency, and c the speed of light, because of which one can show that such finite objects cannot support states of infinite Q, rather there always is some amount of radiation. However, very long-lived (so-called "high-Q") states can be found, whose tails display the needed exponential or exponential-like decay away from the resonant object over long enough distances before they turn oscillatory (radiative). The limiting surface, where this change in the field behavior happens, is called the "radiation caustic", and, for the wireless energy-transfer scheme to be based on the near field rather than the far/radiation field, the distance between the coupled objects must be such that one lies within the radiation caustic of the other. One typical way of achieving a high radiation-Q (Q_{rad}) is to design subwavelength resonant objects. When the size of an object is much smaller than the wavelength of radiation in free space, its electromagnetic field couples to radiation very weakly. Since the extent of the near-field into the area surrounding a finite-sized resonant object is set typically by the wavelength, in some examples, resonant objects of subwavelength size have significantly longer evanescent field-tails. In other words, the radiation caustic is pushed far away from the object, so the electromagnetic mode enters the radiative regime only with a small amplitude.

Moreover, most realistic materials exhibit some nonzero amount of absorption, which can be frequency dependent, and thus cannot support states of infinite Q, rather there always is some amount of absorption. However, very long-lived ("high-Q") states can be found, where electromagnetic modal energy is only weakly dissipated. Some typical ways of achieving a high absorption-Q (Q_{abs}) is to use materials which exhibit very small absorption at the resonant frequency and/or to shape the field to be localized more inside the least lossy materials.

Furthermore, to achieve a large strong-coupling factor U, in some examples, the energy-transfer application may use systems that achieve a high coupling factor k, corresponding to strong (i.e. fast) coupling rate κ, over distances larger than the characteristic sizes of the objects.

Since finite-sized subwavelength resonant objects can often be designed to have high Q, as was discussed above and will be seen in examples later on, such objects may typically be chosen for the resonant device-object. In these cases, the electromagnetic field is, in some examples, of a quasi-static nature and the distance, up to which sufficient coupling can be achieved, is dictated by the decay-law of this quasi-static field.

Note that in some examples, the resonant source-object may be immobile and thus less restricted in its allowed geometry and size. It can be therefore chosen to be large enough that the near-field extent is not limited by the wavelength, and can thus have nearly infinite radiation-Q. Some objects of nearly infinite extent, such as dielectric waveguides, can support guided modes, whose evanescent tails are decaying exponentially in the direction away from the object, slowly if tuned close to cutoff, therefore a good coupling can also be achieved over distances quite a few times larger than a characteristic size of the source- and/or device-object.

2 'Strongly-coupled' resonances at mid-range distances for realistic systems

In the following, examples of systems suitable for energy transfer of the type described above are described. We will demonstrate how to compute the CMT parameters $\omega_{1,2}$, $Q_{1,2}$ and k described above and how to choose or design these parameters for particular examples in order to produce a desirable figure-of-merit $U = \kappa/\sqrt{\Gamma_1\Gamma_2} = k\sqrt{Q_1 Q_2}$ at a desired distance D. In some examples, this figure-of-merit is maximized when $\omega_{1,2}$ are tuned close to a particular angular frequency ω_U.

2.1 Self-resonant conducting coils

In some examples, one or more of the resonant objects are self-resonant conducting coils. Referring to Fig. 3, a conducting wire of length, l, and cross-sectional radius, a, is wound into a helical coil of radius, r, and height, h, (namely with $N = \sqrt{l^2 - h^2}/2\pi r$ number of turns), surrounded by air. As described below, the wire has distributed inductance and distributed capacitance, and therefore it supports a resonant mode of angular frequency ω. The nature of the resonance lies in the periodic exchange of energy from the electric field within the capacitance of the coil, due to the charge distribution $\rho(\mathbf{x})$ across it, to the magnetic field in free space, due to the current distribution $\mathbf{j}(\mathbf{x})$ in the wire. In particular, the charge conservation equation $\nabla \cdot \mathbf{j} = i\omega\rho$ implies that: (i) this periodic exchange is accompanied by a $\pi/2$ phase-shift between the current and the charge density profiles, namely the energy W contained in the coil is at certain points in time completely due to the current and at other points in time completely due to the charge, and (ii) if $\rho_l(x)$ and $I(x)$ are respectively the linear charge and current densities in the wire, where x runs along the wire, $q_o = \frac{1}{2}\int dx|\rho_l(x)|$ is the maximum amount of positive charge accumulated in one side of the coil (where an equal amount of negative charge always also accumulates in the other side to make the system neutral) and $I_o = \max\{|I(x)|\}$ is the maximum positive value of the linear current distribution, then $I_o = \omega q_o$. Then, one can define an effective total inductance L and an effective total capacitance C of the coil through the amount of energy W inside its resonant mode:

$$W \equiv \frac{1}{2} I_o^2 L \Rightarrow L = \frac{\mu_o}{4\pi I_o^2} \iint d\mathbf{x} d\mathbf{x}' \frac{\mathbf{j}(\mathbf{x}) \cdot \mathbf{j}(\mathbf{x}')}{|\mathbf{x}-\mathbf{x}'|}, \tag{19}$$

$$W \equiv \frac{1}{2} q_o^2 \frac{1}{C} \Rightarrow \frac{1}{C} = \frac{1}{4\pi\varepsilon_o q_o^2} \iint d\mathbf{x} d\mathbf{x}' \frac{\rho(\mathbf{x})\rho(\mathbf{x}')}{|\mathbf{x}-\mathbf{x}'|}, \tag{20}$$

where μ_o and ε_o are the magnetic permeability and electric permittivity of free space.

With these definitions, the resonant angular frequency and the effective impedance can be given by the formulas $\omega = 1/\sqrt{LC}$ and $Z = \sqrt{L/C}$ respectively.

Losses in this resonant system consist of ohmic (material absorption) loss inside the wire and radiative loss into free space. One can again define a total absorption resistance R_{abs} from the amount of power absorbed inside the wire and a total radiation resistance R_{rad} from the amount of power radiated due to electric- and magnetic-dipole radiation:

$$P_{abs} \equiv \tfrac{1}{2} I_o^2 R_{abs} \Rightarrow R_{abs} \approx \zeta_c \frac{l}{2\pi a} \cdot \frac{I_{rms}^2}{I_o^2} \tag{21}$$

$$P_{rad} \equiv \tfrac{1}{2} I_o^2 R_{rad} \Rightarrow R_{rad} \approx \frac{\zeta_o}{6\pi}\left[\left(\frac{\omega|\mathbf{p}|}{c}\right)^2 + \left(\frac{\omega\sqrt{|\mathbf{m}|}}{c}\right)^4\right], \tag{22}$$

where $c = 1/\sqrt{\mu_o \varepsilon_o}$ and $\zeta_o = \sqrt{\mu_o/\varepsilon_o}$ are the light velocity and light impedance in free space, the impedance ζ_c is $\zeta_c = 1/\sigma\delta = \sqrt{\mu_o \omega/2\sigma}$ with σ the conductivity of the conductor and δ the skin depth at the frequency ω, $I_{rms}^2 = \frac{1}{l}\int dx |I(x)|^2$, $\mathbf{p} = \int dx\, \mathbf{r}\rho_l(x)$ is the electric-dipole moment of the coil and $\mathbf{m} = \frac{1}{2}\int dx\, \mathbf{r} \times \mathbf{j}(x)$ is the magnetic-dipole moment of the coil. For the radiation resistance formula Eq.(22), the assumption of operation in the quasi-static regime ($h, r \ll \lambda = 2\pi c/\omega$) has been used, which is the desired regime of a subwavelength resonance. With these definitions, the absorption and radiation quality factors of the resonance may be given by $Q_{abs} = Z/R_{abs}$ and $Q_{rad} = Z/R_{rad}$ respectively.

From Eq.(19)-(22) it follows that to determine the resonance parameters one simply needs to know the current distribution $\mathbf{j}$ in the resonant coil. Solving Maxwell's equations to rigorously find the current distribution of the resonant electromagnetic eigenmode of a conducting-wire coil is more involved than, for example, of a standard LC circuit, and we can find no exact solutions in the literature for coils of finite length, making an exact solution difficult. One could in principle write down an elaborate transmission-line-like model, and solve it by brute force. We instead present a model that is (as described below) in good agreement (~5%) with experiment. Observing that the finite extent of the conductor forming each coil imposes the boundary condition that the current has to be zero at the ends of the coil, since no current can leave the wire, we assume that the resonant mode of each coil is well approximated by a sinusoidal current profile along the length of the conducting wire. We shall be interested in the lowest mode, so if we denote by x the coordinate along the conductor, such that it runs from $-l/2$ to $+l/2$, then the current amplitude profile would have the form $I(x)=I_o\cos(\pi x/l)$, where we have assumed that the current does not vary significantly along the wire circumference for a particular x, a valid assumption provided $a \ll r$. It immediately follows from the continuity equation for charge that the linear charge density profile should be of the form $\rho_l(x)=\rho_o\sin(\pi x/l)$, and thus $q_o=\int_0^{l/2}dx\rho_o\left|\sin(\pi x/l)\right|=\rho_o l/\pi$. Using these sinusoidal profiles we find the so-called "self-inductance" L_s and "self-capacitance" C_s of the coil by computing numerically the integrals Eq.(19) and (20); the associated frequency and effective impedance are ω_s and Z_s respectively. The "self-resistances" R_s are given analytically by Eq.(21) and (22) using $I_{rms}^2=\frac{1}{l}\int_{-l/2}^{l/2}dx\left|I_o\cos(\pi x/l)\right|^2=\frac{1}{2}I_o^2$, $|\mathbf{p}|=q_o\sqrt{\left(\frac{2}{\pi}h\right)^2+\left(\frac{4N\cos(\pi N)}{(4N^2-1)\pi}r\right)^2}$ and $|\mathbf{m}|=I_o\sqrt{\left(\frac{2}{\pi}N\pi r^2\right)^2+\left(\frac{\cos(\pi N)(12N^2-1)-\sin(\pi N)\pi N(4N^2-1)}{(16N^4-8N^2+1)\pi}hr\right)^2}$, and therefore the associated Q_s factors can be calculated.

The results for two examples of resonant coils with subwavelength modes of $\lambda_s / r \geq 70$ (i.e. those highly suitable for near-field coupling and well within the quasi-static limit) are presented in Table 1. Numerical results are shown for the wavelength and absorption, radiation and total loss rates, for the two different cases of subwavelength-coil resonant modes. Note that, for conducting material, copper (σ=5.998•10^-7 S/m) was used. It can be seen that expected quality factors at microwave frequencies are $Q_{s,abs} \geq 1000$ and $Q_{s,rad} \geq 5000$.

Table 1

single coil	λ_s / r	f (MHz)	$Q_{s,rad}$	$Q_{s,abs}$	Q_s
r=30cm, h=20cm, a=1cm, N=4	74.7	13.39	4164	8170	2758
r=10cm, h=3cm, a=2mm, N=6	140	21.38	43919	3968	3639

Referring to Fig. 4, in some examples, energy is transferred between two self-resonant conducting-wire coils. The electric and magnetic fields are used to couple the different resonant conducting-wire coils at a distance D between their centers. Usually, the electric coupling highly dominates over the magnetic coupling in the system under consideration for coils with $h \gg 2r$ and, oppositely, the magnetic coupling highly dominates over the electric coupling for coils with $h \ll 2r$. Defining the charge and current distributions of two coils 1,2 respectively as $\rho_{1,2}(\mathbf{x})$ and $\mathbf{j}_{1,2}(\mathbf{x})$, total charges and peak currents respectively as $q_{1,2}$ and $I_{1,2}$, and capacitances and inductances respectively as $C_{1,2}$ and $L_{1,2}$, which are the analogs of $\rho(\mathbf{x})$, $\mathbf{j}(\mathbf{x})$, q_o, I_o, C and L for the single-coil case and are therefore well defined, we can *define* their mutual capacitance and inductance through the total energy:

$$W \equiv W_1 + W_2 + \frac{1}{2}\left(q_1^* q_2 + q_2^* q_1\right) / M_C + \frac{1}{2}\left(I_1^* I_2 + I_2^* I_1\right) M_L$$

$$\Rightarrow 1/M_C = \frac{1}{4\pi\varepsilon_o q_1 q_2}\iint d\mathbf{x}d\mathbf{x}'\frac{\rho_1(\mathbf{x})\rho_2(\mathbf{x}')}{|\mathbf{x}-\mathbf{x}'|}u, \quad M_L = \frac{\mu_o}{4\pi I_1 I_2}\iint d\mathbf{x}d\mathbf{x}'\frac{\mathbf{j}_1(\mathbf{x})\cdot\mathbf{j}_2(\mathbf{x}')}{|\mathbf{x}-\mathbf{x}'|}u, \quad (23)$$

where $W_1 = \frac{1}{2}q_1^2/C_1 = \frac{1}{2}I_1^2 L_1$, $W_2 = \frac{1}{2}q_2^2/C_2 = \frac{1}{2}I_2^2 L_2$ and the retardation factor of $u = \exp(i\omega|\mathbf{x}-\mathbf{x}'|/c)$ inside the integral can been ignored in the quasi-static regime $D \ll \lambda$ of interest, where each coil is within the near field of the other. With this definition, the coupling factor is given by $k = \sqrt{C_1C_2}/M_C + M_L/\sqrt{L_1L_2}$.

Therefore, to calculate the coupling rate between two self-resonant coils, again the current profiles are needed and, by using again the assumed sinusoidal current profiles, we compute numerically from Eq.(23) the mutual capacitance $M_{C,s}$ and inductance $M_{L,s}$ between two self-resonant coils at a distance D between their centers, and thus $k = 1/Q_\kappa$ is also determined.

Table 2

pair of coils	D/r	Q	$Q_\kappa = 1/k$	U
r=30cm, h=20cm, a=1cm, N=4, $\lambda/r \approx 75$, $Q_s^{abs} \approx 8170$, $Q_s^{rad} \approx 4164$	3	2758	38.9	70.9
	5	2758	139.4	19.8
	7	2758	333.0	8.3
	10	2758	818.9	3.4
r=10cm, h=3cm, a=2mm, N=6, $\lambda/r \approx 140$, $Q_s^{abs} \approx 3968$, $Q_s^{rad} \approx 43919$	3	3639	61.4	59.3
	5	3639	232.5	15.7
	7	3639	587.5	6.2
	10	3639	1580	2.3

Referring to Table 2, relevant parameters are shown for exemplary examples featuring pairs or identical self resonant coils. Numerical results are presented for the average wavelength and loss rates of the two normal modes (individual values not shown), and also the coupling rate and figure-of-merit as a function of the coupling distance D, for the two cases of modes presented in Table1. It can be seen that for medium distances $D/r = 10 - 3$ the expected coupling-to-loss ratios are in the range $U \sim 2 - 70$.

2.1.1 Experimental Results

An experimental realization of an example of the above described system for wireless energy transfer consists of two self-resonant coils, one of which (the source coil) is coupled inductively to an oscillating circuit, and the second (the device coil) is coupled inductively to a resistive load, as shown schematically in Fig. 5. Referring to Fig. 5, A is a single copper loop of radius 25cm that is part of the driving circuit, which outputs a sine wave with frequency 9.9MHz. s and d are respectively the source and device coils referred to in the text. B is a loop of wire attached to the load ("light-bulb"). The various κ's represent direct couplings between the objects. The angle between coil d and the loop A is adjusted so that their direct coupling is zero, while coils s and d are aligned coaxially. The direct coupling between B and A and between B and s is negligible.

The parameters for the two identical helical coils built for the experimental validation of the power transfer scheme were $h = 20\,\text{cm}$, $a = 3\,\text{mm}$, $r = 30$ cm and $N = 5.25$. Both coils are made of copper. Due to imperfections in the construction, the spacing between loops of the helix is not uniform, and we have encapsulated the uncertainty about their uniformity by attributing a 10% ($2\,\text{cm}$) uncertainty to h. The expected resonant frequency given these dimensions is $f_0 = 10.56 \pm 0.3\,\text{MHz}$, which is approximately 5% off from the measured resonance at around $9.90\,\text{MHz}$.

The theoretical Q for the loops is estimated to be ~ 2500 (assuming perfect copper of resistivity $\rho = 1/\sigma = 1.7 \times 10^{-8}\,\Omega\text{m}$) but the measured value is 950 ± 50. We believe the discrepancy is mostly due to the effect of the layer of poorly conducting copper oxide on the surface of the copper wire, to which the current is confined by the short skin depth ($\sim 20\mu\text{m}$) at this frequency. We have therefore used the experimentally observed Q (and $\Gamma_1 = \Gamma_2 = \Gamma = \omega/(2Q)$ derived from it) in all subsequent computations.

The coupling coefficient κ can be found experimentally by placing the two self-resonant coils (fine-tuned, by slightly adjusting h, to the same resonant frequency when isolated) a distance D apart and measuring the splitting in the frequencies of the two resonant modes in the transmission spectrum. According to Eq.(13) derived by coupled-mode theory, the splitting in the transmission spectrum should be $\delta_P = 2\sqrt{\kappa^2 - \Gamma^2}$, when $\kappa_{A,B}$ are kept very small by keeping A and B at a relatively large distance. The comparison between experimental and theoretical results as a function of distance when the two the coils are aligned coaxially is shown in Fig. 6.

Fig. 7 shows a comparison of experimental and theoretical values for the strong-coupling factor $U = \kappa / \Gamma$ as a function of the separation between the two coils. The theory values are obtained by using the theoretically obtained κ and the experimentally measured Γ. The shaded area represents the spread in the theoretical U due to the $\sim 5\%$ uncertainty in Q. As noted above, the maximum theoretical efficiency depends only on the parameter U, which is plotted as a function of distance in Fig. 7. U is greater than 1 even for $D = 2.4\,\mathrm{m}$ (eight times the radius of the coils), thus the sytem is in the strongly-coupled regime throughout the entire range of distances probed.

The power-generator circuit was a standard Colpitts oscillator coupled inductively to the source coil by means of a single loop of copper wire 25cm in radius (see Fig. 5). The load consisted of a previously calibrated light-bulb, and was attached to its own loop of insulated wire, which was in turn placed in proximity of the device coil and inductively coupled to it. Thus, by varying the distance between the light-bulb and the device coil, the parameter $U_B = \kappa_B / \Gamma$ was adjusted so that it matched its optimal value, given theoretically by Eq.(14) as $U_{B*} = \sqrt{1+U^2}$. Because of its inductive nature, the loop connected to the light-bulb added a small reactive component to κ_B which was compensated for by slightly retuning the coil. The work extracted was determined by adjusting the power going into the Colpitts oscillator until the light-bulb at the load was at its full nominal brightness.

In order to isolate the efficiency of the transfer taking place specifically between the source coil and the load, we measured the current at the mid-point of each of the self-resonant coils with a current-probe (which was not found to lower the Q of the coils noticeably.) This gave a measurement of the current parameters I_1 and I_2 defined above.

The power dissipated in each coil was then computed from $P_{1,2} = \Gamma L \mid I_{1,2} \mid^2$, and the efficiency was directly obtained from $\eta = P_B / (P_1 + P_2 + P_B)$. To ensure that the experimental setup was well described by a two-object coupled-mode theory model, we positioned the device coil such that its direct coupling to the copper loop attached to the Colpitts oscillator was zero. The experimental results are shown in Fig. 8, along with the theoretical prediction for maximum efficiency, given by Eq.(15).

Using this example, we were able to transmit significant amounts of power using this setup from the source coil to the device coil, fully lighting up a 60W light-bulb from distances more than 2m away, for example. As an additional test, we also measured the total power going into the driving circuit. The efficiency of the wireless power-transmission itself was hard to estimate in this way, however, as the efficiency of the Colpitts oscillator itself is not precisely known, although it is expected to be far from 100%. Nevertheless, this gave an overly conservative lower bound on the efficiency. When transmitting 60W to the load over a distance of 2m, for example, the power flowing into the driving circuit was 400W. This yields an overall wall-to-load efficiency of $\sim 15\%$, which is reasonable given the expected $\sim 40\%$ efficiency for the wireless power transmission at that distance and the low efficiency of the driving circuit.

From the theoretical treatment above, we see that in typical examples it is important that the coils be on resonance for the power transmission to be practical. We found experimentally that the power transmitted to the load dropped sharply as one of the coils was detuned from resonance. For a fractional detuning $\Delta f/f_0$ of a few times the inverse loaded Q, the induced current in the device coil was indistinguishable from noise.

The power transmission was not found to be visibly affected as humans and various everyday objects, such as metallic and wooden furniture, as well as electronic devices large and small, were placed between the two coils, even when they drastically obstructed the line of sight between source and device. External objects were found to have an effect only when they were closer than 10cm from either one of the coils. While some materials (such as aluminum foil, styrofoam and humans) mostly just shifted the resonant frequency, which could in principle be easily corrected with a feedback circuit of the type described earlier, others (cardboard, wood, and PVC) lowered Q when placed closer than a few centimeters from the coil, thereby lowering the efficiency of the transfer.

This method of power transmission is believed safe for humans. When transmitting 60W (more than enough to power a laptop computer) across 2m, we estimated that the magnitude of the magnetic field generated is much weaker than the Earth's magnetic field for all distances except for less than about 1cm away from the wires in the coil, an indication of the safety of the scheme even after long-term use. The power radiated for these parameters was ~ 5 W, which is roughly an order of magnitude higher than cell phones but could be drastically reduced, as discussed below.

Although the two coils are currently of identical dimensions, it is possible to make the device coil small enough to fit into portable devices without decreasing the efficiency. One could, for instance, maintain the product of the characteristic sizes of the source and device coils constant.

These experiments demonstrated experimentally a system for power transmission over medium range distances, and found that the experimental results match theory well in multiple independent and mutually consistent tests.

The efficiency of the scheme and the distances covered can be improved by silver-plating the coils, which may increase their Q, or by working with more elaborate geometries for the resonant objects. Nevertheless, the performance characteristics of the system presented here are already at levels where they could be useful in practical applications.

<u>2.2 Capacitively-loaded conducting loops or coils</u>

In some examples, one or more of the resonant objects are capacitively-loaded conducting loops or coils. Referring to Fig. 9 a helical coil with N turns of conducting wire, as described above, is connected to a pair of conducting parallel plates of area A spaced by distance d via a dielectric material of relative permittivity ε, and everything is surrounded by air (as shown, $N=1$ and $h=0$). The plates have a capacitance

$C_p = \varepsilon_o \varepsilon A / d$, which is added to the distributed capacitance of the coil and thus modifies its resonance. Note however, that the presence of the loading capacitor may modify the current distribution inside the wire and therefore the total effective inductance L and total effective capacitance C of the coil may be different respectively from L_s and C_s, which are calculated for a self-resonant coil of the same geometry using a sinusoidal current profile. Since some charge may accumulate at the plates of the external loading capacitor, the charge distribution ρ inside the wire may be reduced, so $C < C_s$, and thus, from the charge conservation equation, the current distribution $\mathbf{j}$ may flatten out, so $L > L_s$. The resonant frequency for this system may be $\omega = 1/\sqrt{L(C + C_p)} < \omega_s = 1/\sqrt{L_s C_s}$, and $I(x) \to I_o \cos(\pi x / l) \Rightarrow C \to C_s \Rightarrow \omega \to \omega_s$, as $C_p \to 0$.

In general, the desired CMT parameters can be found for this system, but again a very complicated solution of Maxwell's Equations is required. Instead, we will analyze only a special case, where a reasonable guess for the current distribution can be made. When $C_p \gg C_s > C$, then $\omega \approx 1/\sqrt{LC_p} \ll \omega_s$ and $Z \approx \sqrt{L/C_p} \ll Z_s$, while all the charge is on the plates of the loading capacitor and thus the current distribution is constant along the wire. This allows us now to compute numerically L from Eq.(19). In the case $h = 0$ and N integer, the integral in Eq.(19) can actually be computed analytically, giving the formula $L = \mu_o r\left[\ln(8r/a) - 2\right]N^2$. Explicit analytical formulas are again available for R from Eq.(21) and (22), since $I_{rms} = I_o$, $|\mathbf{p}| \approx 0$ and $|\mathbf{m}| = I_o N \pi r^2$ (namely only the magnetic-dipole term is contributing to radiation), so we can determine also $Q_{abs} = \omega L / R_{abs}$ and $Q_{rad} = \omega L / R_{rad}$. At the end of the calculations, the validity of the assumption of constant current profile is confirmed by checking that indeed the condition $C_p \gg C_s \Leftrightarrow \omega \ll \omega_s$ is satisfied. To satisfy this condition, one could use a large external capacitance, however, this would usually shift the operational frequency lower than the optimal frequency, which we will determine shortly; instead, in typical examples, one often prefers coils with very small self-capacitance C_s to begin with, which usually holds, for the types of coils under consideration, when $N=1$, so that the self-capacitance comes from the charge distribution across the single turn, which is almost always very small, or when $N>1$ and $h \gg 2Na$, so that the dominant self-capacitance comes from the charge distribution across adjacent turns, which is small if the separation between adjacent turns is large.

The external loading capacitance C_p provides the freedom to tune the resonant frequency (for example by tuning A or d). Then, for the particular simple case $h=0$, for which we have analytical formulas, the total $Q=\omega L/(R_{abs}+R_{rad})$ becomes highest at the optimal frequency

$$\omega_Q = \left[\frac{c^4}{\pi}\sqrt{\frac{\varepsilon_o}{2\sigma}}\cdot\frac{1}{aNr^3}\right]^{2/7}, \tag{24}$$

reaching the value

$$Q_{\max} = \frac{6}{7\pi}\left(2\pi^2\eta_o\frac{\sigma a^2 N^2}{r}\right)^{3/7}\cdot\left[\ln\left(\frac{8r}{a}\right)-2\right]. \tag{25}$$

At lower frequencies Q is dominated by ohmic loss and at higher frequencies by radiation. Note, however, that the formulas above are accurate as long as $\omega_Q \ll \omega_s$ and, as explained above, this holds almost always when $N=1$, and is usually less accurate when $N>1$, since $h=0$ usually implies a large self-capacitance. A coil with large h can be used, if the self-capacitance needs to be reduced compared to the external capacitance, but then the formulas for L and ω_Q, $Q_{\max}$ are again less accurate. Similar qualitative behavior is expected, but a more complicated theoretical model is needed for making quantitative predictions in that case.

The results of the above analysis for two examples of subwavelength modes of $\lambda/r \geq 70$ (namely highly suitable for near-field coupling and well within the quasi-static limit) of coils with $N=1$ and $h=0$ at the optimal frequency Eq.(24) are presented in Table 3. To confirm the validity of constant-current assumption and the resulting analytical formulas, mode-solving calculations were also performed using another completely independent method: computational 3D finite-element frequency-domain (FEFD) simulations (which solve Maxwell's Equations in frequency domain exactly apart for spatial discretization) were conducted, in which the boundaries of the conductor were modeled using a complex impedance $\zeta_c = \sqrt{\mu_o \omega / 2\sigma}$ boundary condition, valid as long as $\zeta_c / \zeta_o \ll 1$ ($<10^{-5}$ for copper in the microwave). Table 3 shows Numerical FEFD (and in parentheses analytical) results for the wavelength and absorption, radiation and total loss rates, for two different cases of subwavelength-loop resonant modes. Note that copper was used for the conducting material ($\sigma = 5.998 \cdot 10^7 S/m$). Specific parameters of the plot in Fig. 4 are highlighted in bold in the table. The two methods (analytical and computational) are in good agreement and show that, in some examples, the optimal frequency is in the low-MHz microwave range and the expected quality factors are $Q_{abs} \geq 1000$ and $Q_{rad} \geq 10000$.

Table 3

single coil	λ / r	f	Q_{rad}	Q_{abs}	Q
r=30cm, a=2cm ε=10, A=138cm^2, d=4mm	**111.4 (112.4)**	**8.976 (8.897)**	**29546 (30512)**	**4886 (5117)**	**4193 (4381)**
r=10cm, a=2mm ε=10, A=3.14cm^2, d=1mm	69.7 (70.4)	43.04 (42.61)	10702 (10727)	1545 (1604)	1350 (1395)

Referring to Fig. 10, in some examples, energy is transferred between two capacitively-loaded coils. For the rate of energy transfer between two capacitively-loaded coils 1 and 2 at distance D between their centers, the mutual inductance M_L can be evaluated numerically from Eq.(23) by using constant current distributions in the case $\omega \ll \omega_s$. In the case $h = 0$, the coupling may be only magnetic and again we have an analytical formula, which, in the quasi-static limit $r \ll D \ll \lambda$ and for the relative orientation shown in Fig. 10, is $M_L \approx \pi\mu_o / 2 \cdot (r_1 r_2)^2 N_1 N_2 / D^3$, which means that $k \propto \left(\sqrt{r_1 r_2} / D\right)^3$ may be independent of the frequency ω and the number of turns N_1, N_2. Consequently, the resultant coupling figure-of-merit of interest is

$$U = k\sqrt{Q_1 Q_2} \approx \left(\frac{\sqrt{r_1 r_2}}{D}\right)^3 \cdot \frac{\pi^2 \eta_o \frac{\sqrt{r_1 r_2}}{\lambda} \cdot N_1 N_2}{\prod_{j=1,2}\left(\sqrt{\frac{\pi\eta_o}{\lambda\sigma}} \cdot \frac{r_j}{a_j} N_j + \frac{8}{3}\pi^5 \eta_o \left(\frac{r_j}{\lambda}\right)^4 N_j^2\right)^{1/2}}, \tag{26}$$

which again is more accurate for $N_1 = N_2 = 1$.

From Eq.(26) it can be seen that the optimal frequency ω_U, where the figure-of-merit is maximized to the value $U_{\max}$, is close to the frequency $\omega_{Q_1 Q_2}$ at which $Q_1 Q_2$ is maximized, since k does not depend much on frequency (at least for the distances $D \ll \lambda$ of interest for which the quasi-static approximation is still valid). Therefore, the optimal frequency $\omega_U \approx \omega_{Q_1 Q_2}$ may be mostly independent of the distance D between the two coils and may lie between the two frequencies ω_{Q_1} and ω_{Q_2} at which the single-coil Q_1 and Q_2 respectively peak. For same coils, this optimal frequency is given by Eq.(24) and then the strong-coupling factor from Eq.(26) becomes

$$U_{\max} = kQ_{\max} \approx \left(\frac{r}{D}\right)^3 \cdot \frac{3}{7}\left(2\pi^2 \eta_o \frac{\sigma a^2 N^2}{r}\right)^{3/7}. \tag{27}$$

In some examples, one can tune the capacitively-loaded conducting loops or coils, so that their angular resonant frequencies are close to ω_U within Γ_U, which is half the angular frequency width for which $U > U_{\max}/2$.

Referring to Table 4, numerical FEFD and, in parentheses, analytical results based on the above are shown for two systems each composed of a matched pair of the loaded coils described in Table 3. The average wavelength and loss rates are shown along with the coupling rate and coupling to loss ratio figure-of-merit $U = \kappa / \Gamma$ as a function of the coupling distance D, for the two cases. Note that the average numerical Γ_{rad} shown are slightly different from the single-loop value of Figure 3. Analytical results for Γ_{rad} are not shown but the single-loop value is used. (The specific parameters corresponding to the plot in Fig. 10 are highlighted with bold in the table.) Again we chose $N = 1$ to make the constant-current assumption a good one and computed M_L numerically from Eq.(23). Indeed the accuracy can be confirmed by their agreement with the computational FEFD mode-solver simulations, which give κ through the frequency splitting of the two normal modes of the combined system ($\delta_E = 2\kappa$ from Eq.(4)). The results show that for medium distances $D/r = 10 - 3$ the expected coupling-to-loss ratios are in the range $U \sim 0.5 - 50$.

Table 4

pair of coils	D/r	Q^{rad}	$Q = \omega/2\Gamma$	$Q_\kappa = \omega/2\kappa$	κ/Γ
r=30cm, a=2cm **ε=10, A=138cm², d=4mm** $\lambda/r \approx 112$ $Q^{abs} \approx 4886$	3	30729	4216	62.6 (63.7)	67.4 (68.7)
	5	29577	4194	235 (248)	17.8 (17.6)
	7	**29128**	**4185**	**589 (646)**	**7.1 (6.8)**
	10	28833	4177	1539 (1828)	2.7 (2.4)
r=10cm, a=2mm ε=10, A=3.14cm², d=1mm $\lambda/r \approx 70$ $Q^{abs} \approx 1546$	3	10955	1355	85.4 (91.3)	15.9 (15.3)
	5	10740	1351	313 (356)	4.32 (3.92)
	7	10759	1351	754 (925)	1.79 (1.51)
	10	10756	1351	1895 (2617)	0.71 (0.53)

<u>*2.2.1 Derivation of optimal power-transmission efficiency*</u>

Referring to Fig. 11, to rederive and express Eq.(15) in terms of the parameters which are more directly accessible from particular resonant objects, such as the capacitively-loaded conducting loops, one can consider the following circuit-model of the system, where the inductances L_s, L_d represent the source and device loops respectively, R_s, R_d their respective losses, and $C\ , C_d$ are the required corresponding capacitances to achieve for both resonance at frequency ω. A voltage generator V_g is considered to be connected to the source and a load resistance R_l to the device. The mutual inductance is denoted by M.

Then from the source circuit at resonance ($\omega L\ = 1/\omega C\ $):

$$V_g = I_s R_s - j\omega M I_d \Rightarrow \frac{1}{2} V_g^* I_s = \frac{1}{2} \left|I_s\right|^2 R_s + \frac{1}{2} j\omega M I_d^* I_s, \tag{28}$$

and from the device circuit at resonance ($\omega L_d = 1/\omega C_d$):

$$0 = I_d\left(R_d + R_l\right) - j\omega M I_s \Rightarrow j\omega M I_s = I_d\left(R_d + R_l\right) \tag{29}$$

So by substituting Eq.(29) to Eq.(28) and taking the real part (for time-averaged power) we get:

$$P_g = \mathrm{Re}\left\{\frac{1}{2} V_g^* I_s\right\} = \frac{1}{2}\left|I_s\right|^2 R_s + \frac{1}{2}\left|I_d\right|^2\left(R_d + R_l\right) = P_s + P_d + P_l, \tag{30}$$

where we identified the power delivered by the generator $P_g = \mathrm{Re}\left\{V_g^* I_s / 2\right.$,the power lost inside the source $P_s = \left|I_s\right|^2 R_s / 2$, the power lost inside the device $P_d = \left|I_d\right|^2 R_d / 2$ and the power delivered to the load $P_l = \left|I_d\right|^2 R_l / 2$. Then, the power transmission efficiency is:

$$\eta_{\mathrm{P}} \equiv \frac{P_l}{P_g} = \frac{R_l}{\left|\frac{I_s}{I_d}\right|^2 R_s + \left(R_d + R_l\right.} \overset{(29)}{=} \frac{R_l}{\frac{\left(R_d + R_l\right.^2}{\left(\omega M\right.^2} R_s + \left(R_d + R_l\right.}. \tag{31}$$

If we now choose the load impedance R_l to optimize the efficiency by $\eta_{\mathrm{P}}^{'}\left(R_l\right) = 0$, we get the optimal load impedance

$$\frac{R_{l*}}{R_d} = \sqrt{1 + \frac{\left(\omega M\right)^2}{R_s R_d}} \tag{32}$$

and the maximum possible efficiency

$$\eta_{\mathrm{P}*} = \frac{R_{l*} / R_d - 1}{R_{l*} / R_d + 1} = \left[\frac{\omega M / \sqrt{R_s R_d}}{1 + \sqrt{1 + \left(\omega M / \sqrt{R_s R_d}\right)^2}}\right]^2. \tag{33}$$

To check now the correspondence with the CMT model, note that $\kappa_l = R_l / 2L_d$, $\Gamma_d = R_d / 2L_d$, $\Gamma_s = R_s / 2L_s$, and $\kappa \quad \omega M / 2\sqrt{L_s L_d}$, so then $U_l = \kappa_l / \Gamma_d = R_l / R_d$ and $U = \kappa / \sqrt{\Gamma_s \Gamma_d} = \omega M / \sqrt{R_s R_d}$. Therefore, the condition Eq.(32) is identical to the condition Eq.(14) and the optimal efficiency Eq.(33) is identical to the general Eq.(15). Indeed, as the CMT analysis predicted, to get a large efficiency, we need to design a system that has a large strong-coupling factor U.

2.2.2 Optimization of U

The results above can be used to increase or optimize the performance of a wireless energy transfer system, which employs capacitively-loaded coils. For example, from the scaling of Eq.(27) with the different system parameters, one sees that to maximize the system figure-of-merit U, in some examples, one can:

-- Decrease the resistivity of the conducting material. This can be achieved, for example, by using good conductors (such as copper or silver) and/or lowering the temperature. At very low temperatures one could use also superconducting materials to achieve extremely good performance.

-- Increase the wire radius a. In typical examples, this action can be limited by physical size considerations. The purpose of this action is mainly to reduce the resistive losses in the wire by increasing the cross-sectional area through which the electric current is flowing, so one could alternatively use also a Litz wire, or ribbon, or any low AC-resistance structure, instead of a circular wire.

-- For fixed desired distance D of energy transfer, increase the radius of the loop r. In typical examples, this action can be limited by physical size considerations.

-- For fixed desired distance vs. loop-size ratio D/r, decrease the radius of the loop r. In typical examples, this action can be limited by physical size considerations.

-- Increase the number of turns N. (Even though Eq.(27) is expected to be less accurate for $N > 1$, qualitatively it still provides a good indication that we expect an improvement in the coupling-to-loss ratio with increased N.) In typical examples, this action can be limited by physical size and possible voltage considerations, as will be discussed in following paragraphs.

-- Adjust the alignment and orientation between the two coils. The figure-of-merit is optimized when both cylindrical coils have exactly the same axis of cylindrical symmetry (namely they are "facing" each other). In some examples, particular mutual coil angles and orientations that lead to zero mutual inductance (such as the orientation where the axes of the two coils are perpendicular and the centers of the two coils are on one of the two axes) should be avoided.

-- Finally, note that the height of the coil h is another available design parameter, which can have an impact on the performance similar to that of its radius *r*, and thus the design rules can be similar.

The above analysis technique can be used to design systems with desired parameters. For example, as listed below, the above described techniques can be used to determine the cross sectional radius *a* of the wire used to design a system including two same single-turn loops with a given radius in order to achieve a specific performance in terms of $U = \kappa / \Gamma$ at a given D / r between them, when the loop material is copper ($\sigma=5.998\cdot 10^7 S/m$):

$$\begin{aligned}
&D / r = 5,\ U \geq 10,\ r = 30cm \Rightarrow a \geq 9mm \\
&D / r = 5,\ U \geq 10,\ r = 5cm \Rightarrow a \geq 3.7mm \\
&D / r = 5,\ U \geq 20,\ r = 30cm \Rightarrow a \geq 20mm \\
&D / r = 5,\ U \geq 20,\ r = 5cm \Rightarrow a \geq 8.3mm \\
&D / r = 10,\ U \geq 1,\ r = 30cm \Rightarrow a \geq 7mm \\
&D / r = 10,\ U \geq 1,\ r = 5cm \Rightarrow a \geq 2.8mm \\
&D / r = 10,\ U \geq 3,\ r = 30cm \Rightarrow a \geq 25mm \\
&D / r = 10,\ U \geq 3,\ r = 5cm \Rightarrow a \geq 10mm
\end{aligned}$$

Similar analysis can be done for the case of two dissimilar loops. For example, in some examples, the device under consideration may be identified specifically (e.g. a laptop or a cell phone), so the dimensions of the device object (r_d, h_d, a_d, N_d) may be restricted. However, in some such examples, the restrictions on the source object (r_s, h_s, a_s, N_s) may be much less, since the source can, for example, be placed under the floor or on the ceiling. In such cases, the desired distance between the source and device may be fixed; in other cases it may be variable. Listed below are examples (simplified to the case $N_s = N_d = 1$ and $h_s = h_d = 0$) of how one can vary the dimensions of the source object to achieve a desired system performance in terms of $U_{sd} = \kappa / \sqrt{\Gamma_s \Gamma_d}$, when the material is again copper (σ=*5.998·10*^{7}*S/m*):

$$
\begin{aligned}
&D = 1.5m,\ U_{sd} \geq 15,\ r_d = 30cm,\ a_d = 6mm \Rightarrow r_s = 1.158m,\ a_s \geq 5mm \\
&D = 1.5m,\ U_{sd} \geq 30,\ r_d = 30cm,\ a_d = 6mm \Rightarrow r_s = 1.15m,\ a_s \geq 33mm \\
&D = 1.5m, U_{sd} \geq 1,\ r_d = 5cm,\ a_d = 4mm \Rightarrow r_s = 1.119m,\ a_s \geq 7mm \\
&D = 1.5m,\ U_{sd} \geq 2,\ r_d = 5cm,\ a_d = 4mm \Rightarrow r_s = 1.119m,\ a_s \geq 52mm \\
&D = 2m,\ U_{sd} \geq 10,\ r_d = 30cm,\ a_d = 6mm \Rightarrow r_s = 1.518m,\ a_s \geq 7mm \\
&D = 2m,\ U_{sd} \geq 20,\ r_d = 30cm,\ a_d = 6mm \Rightarrow r_s = 1.514m,\ a_s \geq 50mm \\
&D = 2m,\ U_{sd} \geq 0.5,\ r_d = 5cm,\ a_d = 4mm \Rightarrow r_s = 1.491m,\ a_s \geq 5mm \\
&D = 2m,\ U_{sd} \geq 1,\ r_d = 5cm,\ a_d = 4mm \Rightarrow r_s = 1.491m,\ a_s \geq 36mm
\end{aligned}
$$

2.2.3 Optimization of k

As described below, in some examples, the quality factor Q of the resonant objects may be limited from external perturbations and thus varying the coil parameters may not lead to significant improvements in Q. In such cases, one can opt to increase the strong-coupling factor U by increasing the coupling factor k. The coupling does not depend on the frequency and may weakly depend on the number of turns. Therefore, in some examples, one can:

-- Increase the wire radii a_1 and a_2. In typical examples, this action can be limited by physical size considerations.

-- For fixed desired distance D of energy transfer, increase the radii of the coils r_1 and r_2. In typical examples, this action can be limited by physical size considerations.

-- For fixed desired distance vs. coil-sizes ratio $D/\sqrt{r_1 r_2}$, only the weak (logarithmic) dependence of the inductance remains, which suggests that one should decrease the radii of the coils r_1 and r_2. In typical examples, this action can be limited by physical size considerations.

-- Adjust the alignment and orientation between the two coils. In typical examples, the coupling is optimized when both cylindrical coils have exactly the same axis of cylindrical symmetry (namely they are "facing" each other). Particular mutual coil angles and orientations that lead to zero mutual inductance (such as the orientation where the axes of the two coils are perpendicular and the centers of the two coils are on one of the two axes) should obviously be avoided.

-- Finally, note that the heights of the coils h_1 and h_2 are other available design parameters, which can have an impact to the coupling similar to that of their radii r_1 and r_2, and thus the design rules can be similar.

Further practical considerations apart from efficiency, e.g. physical size limitations, will be discussed in detail below.

2.2.4 Optimization of overall system performance

In embodiments, the dimensions of the resonant objects may be determined by the particular application. For example, when the application is powering a laptop or a cell-phone, the device resonant object cannot have dimensions that exceed those of the laptop or cell-phone respectively. For a system of two loops of specified dimensions, in terms of loop radii $r_{s,d}$ and wire radii $a_{s,d}$, the independent parameters left to adjust for the system optimization are: the number of turns $N_{s,d}$, the frequency *f*, the power-load consumption rate $\kappa_l = R_l / 2L_d$ and the power-generator feeding rate $\kappa_g = R_g / 2L_s$, where R_g is the internal (characteristic) impedance of the generator.

In general, in various examples, the dependent variable that one may want to increase or optimize may be the overall efficiency η. However, other important variables may need to be taken into consideration upon system design. For example, in examples featuring capacitively-loaded coils, the designs can be constrained by, the currents flowing inside the wires $I_{s,d}$ and other components and the voltages across the capacitors $V_{s,d}$. These limitations can be important because for ~*Watt* power applications the values for these parameters can be too large for the wires or the capacitors respectively to handle. Furthermore, the total loaded (by the load) quality factor of the device $Q_{d[l]} = \omega / 2(\Gamma_d + \Gamma_l) = \omega L_d / (R_d + R_l)$ and the total loaded (by the generator) quality factor of the source $Q_{s[g]} = \omega / 2(\Gamma_s + \Gamma_g) = \omega L_s / (R_s + R_g)$ are quantities that should be preferably small, because to match the source and device resonant frequencies to within their Q's, when those are very large, can be challenging experimentally and more sensitive to slight variations. Lastly, the radiated powers $P_{s,rad}$ and $P_{d,rad}$ may need to be minimized for concerns about far-field interference and safety, even though, in general, for a magnetic, non-radiative scheme they are already typically small. In the following, we examine then the effects of each one of the independent variables on the dependent ones.

We define a new variable wp to express the power-load consumption rate for some particular value of U through $U_l = \kappa_l / \Gamma_d = \sqrt{1 + wp \cdot U^2}$. Then, in some examples, values which may impact the choice of this rate are: $U_l = 1 \Leftrightarrow wp = 0$ to minimize the required energy stored in the source (and therefore I_s and V_s), $U_l = \sqrt{1+U^2} > 1 \Leftrightarrow wp = 1$ to maximize the efficiency, as seen earlier, or $U_l \gg 1 \Leftrightarrow wp \gg 1$ to decrease the required energy stored in the device (and therefore I_d and V_d) and to decrease or minimize $Q_{d[l]}$. Similar is the impact of the choice of the power-generator feeding rate $U_g = \kappa_g / \Gamma_s$, with the roles of the source/device and generator/load reversed.

In some examples, increasing N_s and N_d may increase Q_s and Q_d, and thus U and the efficiency significantly. It also may decrease the currents I_s and I_d, because the inductance of the loops may increase, and thus the energy $W_{s,d} = L_{s,d} \left| I_{s,d} \right|^2 / 2$ required for given output power P_l can be achieved with smaller currents. However, in some examples, increasing N_d and thus Q_d can increase $Q_{d[l]}$, $P_{d,rad}$ and the voltage across the device capacitance V_d. Similar can be the impact of increasing N_s on $Q_{s[g]}$, $P_{s,rad}$ and V_s. As a conclusion, in some examples, the number of turns N_s and N_d may be chosen large enough (for high efficiency) but such that they allow for reasonable voltages, loaded Q's and/or powers radiated.

With respect to the resonant frequency, again, there may be an optimal one for efficiency. Close to that optimal frequency $Q_{d[l]}$ and/or $Q_{s[g]}$ can be approximately maximum. In some examples, for lower frequencies the currents may typically get larger but the voltages and radiated powers may get smaller. In some examples, one may choose the resonant frequency to maximize any of a number of system parameters or performance specifications, such as efficiency.

One way to decide on an operating regime for the system may be based on a graphical method. Consider two loops of $r_s = 25cm,\ r_d = 15cm,\ h_s = h_d = 0,\ a_s = a_d = 3mm$ and distance $D = 2m$ between them. In Fig. 12, we plot some of the above dependent variables (currents, voltages and radiated powers normalized to *1Watt* of output power) in terms of frequency f and N_d, given some choice for wp and N_s. Fig. 12 depicts the trend of system performance explained above. In Fig. 13, we make a contour plot of the dependent variables as functions of both frequency and *wp* but for both N_s and N_d fixed. For example, in embodiments, a reasonable choice of parameters for the system of two loops with the dimensions given above may be: $N_s = 2,\ N_d = 6$, *f=10MHz* and *wp=10*, which gives the following performance characteristics: $\eta = 20.6\%$, $Q_{d[l]} = 1264$, $I_s = 7.2A,\ I_d = 1.4A,$ $V_s = 2.55kV,\ V_d = 2.30kV,\ P_{s,rad} = 0.15W,\ P_{d,rad} = 0.006W$. Note that the results in Figs. 12, 13 and the calculated performance characteristics are made using the analytical formulas provided above, so they are expected to be less accurate for large values of N_s, N_d, but still may give a good estimate of the scalings and the orders of magnitude.

Finally, in embodiments, one could additionally optimize for the source dimensions, if, for example, only the device dimensions are limited. Namely, one can add r_s and a_s in the set of independent variables and optimize with respect to these all the dependent variables of the system. In embodiments, such an optimization may lead to improved results.

2.3 Inductively-loaded conducting rods

A straight conducting rod of length $2h$ and cross-sectional radius a has distributed capacitance and distributed inductance, and therefore can support a resonant mode of angular frequency ω. Using the same procedure as in the case of self-resonant coils, one can define an effective total inductance L and an effective total capacitance C of the rod through formulas Eqs.(19) and (20). With these definitions, the resonant angular frequency and the effective impedance may be given again by the common formulas $\omega = 1/\sqrt{LC}$ and $Z = \sqrt{L/C}$ respectively. To calculate the total inductance and capacitance, one can assume again a sinusoidal current profile along the length of the conducting wire. When interested in the lowest mode, if we denote by x the coordinate along the conductor, such that it runs from $-h$ to $+h$, then the current amplitude profile may have the form $I(x) = I_o \cos(\pi x / 2h)$, since it has to be zero at the open ends of the rod. This is the well-known half-wavelength electric dipole resonant mode.

In some examples, one or more of the resonant objects may be inductively-loaded conducting rods. Referring to Fig. 14, a straight conducting rod of length $2h$ and cross-sectional radius a, as in the previous paragraph, is cut into two equal pieces of length h, which may be connected via a coil wrapped around a magnetic material of relative permeability μ, and everything is surrounded by air. The coil has an inductance L_C, which is added to the distributed inductance of the rod and thus modifies its resonance. Note however, that the presence of the center-loading inductor may modify the current distribution inside the wire and therefore the total effective inductance L and total effective capacitance C of the rod may be different respectively from L_s and C_s, which are calculated for a self-resonant rod of the same total length using a sinusoidal current profile, as in the previous paragraph. Since some current may be running inside the coil of the external loading inductor, the current distribution $\mathbf{j}$ inside the rod may be reduced, so $L < L_s$, and thus, from the charge conservation equation, the linear charge distribution ρ_l may flatten out towards the center (being positive in one side of the rod and negative in the other side of the rod, changing abruptly through the inductor), so $C > C_s$. The resonant frequency for this system may be $\omega = 1/\sqrt{(L + L_c\ C} < \omega_s = 1/\sqrt{L_s C_s}$, and $I(x) \to I_o \cos(\pi x / 2h) \Rightarrow\ L \to L_s \Rightarrow\ \omega \to \omega_s$, as $L_c \to 0$.

In general, the desired CMT parameters can be found for this system, but again a very complicated solution of Maxwell's Equations is generally required. In a special case, a reasonable estimate for the current distribution can be made. When $L_c \gg L_s > L$, then $\omega \approx 1/\sqrt{L_c C} \ll \omega_s$ and $Z \approx \sqrt{L_c / C} \gg Z_s$, while the current distribution is triangular along the rod (with maximum at the center-loading inductor and zero at the ends) and thus the charge distribution may be positive constant on one half of the rod and equally negative constant on the other side of the rod. This allows us to compute numerically C from Eq.(20). In this case, the integral in Eq.(20) can actually be computed analytically, giving the formula $1/C = 1/(\pi\varepsilon_o h)\left[\ln(h/a) - 1\right]$. Explicit analytical formulas are again available for R from Eq.(21) and (22), since $I_{rms} = I_o$, $|\mathbf{p}| = q_o h$ and $|\mathbf{m}| = 0$ (namely only the electric-dipole term is contributing to radiation), so we can determine also $Q_{abs} = 1/\omega C R_{abs}$ and $Q_{rad} = 1/\omega C R_{rad}$. At the end of the calculations, the validity of the assumption of triangular current profile may be confirmed by checking that indeed the condition $L_c \gg L_s \Leftrightarrow \omega \ll \omega_s$ is satisfied. This condition may be relatively easily satisfied, since typically a conducting rod has very small self-inductance L_s to begin with.

Another important loss factor in this case is the resistive loss inside the coil of the external loading inductor L_c and it may depend on the particular design of the inductor. In some examples, the inductor may be made of a Brooks coil, which is the coil geometry which, for fixed wire length, may demonstrate the highest inductance and thus quality factor. The Brooks coil geometry has N_{Bc} turns of conducting wire of cross-sectional radius a_{Bc} wrapped around a cylindrically symmetric coil former, which forms a coil with a square cross-section of side r_{Bc}, where the inner side of the square is also at radius r_{Bc} (and thus the outer side of the square is at radius $2r_{Bc}$), therefore $N_{Bc} \approx (r_{Bc}/2a_{Bc})^2$. The inductance of the coil is then $L_c = 2.0285\mu_o r_{Bc} N_{Bc}^2 \approx 2.0285\mu_o r_{Bc}^5 / 8a_{Bc}^4$ and its resistance $R_c \approx \frac{1}{\sigma}\frac{l_{Bc}}{\pi a_{Bc}^2}\sqrt{1+\frac{\mu_o\omega\sigma}{2}\left(\frac{a_{Bc}}{2}\right)^2}$, where the total wire length is $l_{Bc} \approx 2\pi(3r_{Bc}/2)N_{Bc} \approx 3\pi r_{Bc}^3/4a_{Bc}^2$ and we have used an approximate square-root law for the transition of the resistance from the dc to the ac limit as the skin depth varies with frequency.

The external loading inductance L_c provides the freedom to tune the resonant frequency. For example, for a Brooks coil with a fixed size r_{Bc}, the resonant frequency can be reduced by increasing the number of turns N_{Bc} by decreasing the wire cross-sectional radius a_{Bc}. Then the desired resonant angular frequency $\omega = 1/\sqrt{L_c C}$ may be achieved for $a_{Bc} \approx \left(2.0285\mu_o r_{Bc}^5 \omega^2 C\right)^{1/4}$ and the resulting coil quality factor may be $Q_c \approx 0.169\mu_o \sigma r_{Bc}^2 \omega / \sqrt{1+\omega^2\mu_o\sigma\sqrt{2.0285\mu_o (r_{Bc}/4)^5 C}}$. Then, for the particular simple case $L_c \gg L_s$, for which we have analytical formulas, the total $Q = 1/\omega C(R_c + R_{abs} + R_{rad})$ becomes highest at some optimal frequency ω_Q, reaching the value $Q_{\max}$, that may be determined by the loading-inductor specific design. For example, for the Brooks-coil procedure described above, at the optimal frequency $Q_{\max} \approx Q_c \approx 0.8\left(\mu_o \sigma^2 r_{Bc}^3 / C\right)^{1/4}$. At lower frequencies it is dominated by ohmic loss inside the inductor coil and at higher frequencies by radiation. Note, again, that the above formulas are accurate as long as $\omega_Q \ll \omega_s$ and, as explained above, this may be easy to design for by using a large inductance.

The results of the above analysis for two examples, using Brooks coils, of subwavelength modes of $\lambda / h \geq 200$ (namely highly suitable for near-field coupling and well within the quasi-static limit) at the optimal frequency ω_Q are presented in Table 5.

Table 5 shows in parentheses (for similarity to previous tables) analytical results for the wavelength and absorption, radiation and total loss rates, for two different cases of subwavelength-rod resonant modes. Note that copper was used for the conducting material ($\sigma=5.998\cdot 10^7 S/m$). The results show that, in some examples, the optimal frequency may be in the low-MHz microwave range and the expected quality factors may be $Q_{abs} \geq 1000$ and $Q_{rad} \geq 100000$.

Table 5

single rod	λ / h	f (MHz)	Q_{rad}	Q_{abs}	Q
h=30cm, a=2cm μ=1, r_{Bc}=2cm, a_{Bc}=0.88mm, N_{Bc}=129	(403.8)	(2.477)	(2.72*10^6)	(7400)	(7380)
h=10cm, a=2mm μ=1, r_{Bc}=5mm, a_{Bc}=0.25mm, N_{Bc}=103	(214.2)	(14.010)	(6.92*10^5)	(3908)	(3886)

In some examples, energy may be transferred between two inductively-loaded rods. For the rate of energy transfer between two inductively-loaded rods 1 and 2 at distance D between their centers, the mutual capacitance M_C can be evaluated numerically from Eq.(23) by using triangular current distributions in the case $\omega \ll \omega_s$. In this case, the coupling may be only electric and again we have an analytical formula, which, in the quasi-static limit $h \ll D \ll \lambda$ and for the relative orientation such that the two rods are aligned on the same axis, is $1/M_C \approx 1/2\pi\varepsilon_o \cdot (h_1 h_2)^2 / D^3$, which means that $k \propto \left(\sqrt{h_1 h_2} / D\right)^3$ is independent of the frequency ω. One can then get the resultant strong-coupling factor U.

It can be seen that the frequency ω_U, where the figure-of-merit is maximized to the value U_{max}, is close to the frequency $\omega_{Q_1Q_2}$, where Q_1Q_2 is maximized, since k does not depend much on frequency (at least for the distances $D \ll \lambda$ of interest for which the quasi-static approximation is still valid). Therefore, the optimal frequency $\omega_U \approx \omega_{Q_1Q_2}$ may be mostly independent of the distance D between the two rods and may lie between the two frequencies ω_{Q_1} and ω_{Q_2} at which the single-rod Q_1 and Q_2 respectively peak. In some typical examples, one can tune the inductively-loaded conducting rods, so that their angular eigenfrequencies are close to ω_U within Γ_U, which is half the angular frequency width for which $U > U_{max}/2$.

Referring to Table 6, in parentheses (for similarity to previous tables) analytical results based on the above are shown for two systems each composed of a matched pair of the loaded rods described in Table 5. The average wavelength and loss rates are shown along with the coupling rate and coupling to loss ratio figure-of-merit $U = \kappa / \Gamma$ as a function of the coupling distance D, for the two cases. Note that for Γ_{rad} the single-rod value is used. Again we chose $L_c \gg L_s$ to make the triangular-current assumption a good one and computed M_C numerically from Eq.(23). The results show that for medium distances $D/h = 10-3$ the expected coupling-to-loss ratios are in the range $U \sim 0.5-100$.

Table 6

pair of rods	D/h	$Q_\kappa = 1/k$	U
h=30cm, a=2cm μ=1, r_{Bc}=2cm, a_{Bc}=0.88mm, N_{Bc}=129 $\lambda/h \approx 404$ $Q \approx 7380$	3	(70.3)	(105.0)
	5	(389)	(19.0)
	7	(1115)	(6.62)
	10	(3321)	(2.22)
h=10cm, a=2mm μ=1, r_{Bc}=5mm, a_{Bc}=0.25mm, N_{Bc}=103 $\lambda/h \approx 214$ $Q \approx 3886$	3	(120)	(32.4)
	5	(664)	(5.85)
	7	(1900)	(2.05)
	10	(5656)	(0.69)

2.4 Dielectric disks

In some examples, one or more of the resonant objects are dielectric objects, such as disks. Consider a two dimensional dielectric disk object, as shown in Fig. 15(a), of radius r and relative permittivity ε surrounded by air that supports high-Q "whispering-gallery" resonant modes. The loss mechanisms for the energy stored inside such a resonant system are radiation into free space and absorption inside the disk material. High-Q_{rad} and long-tailed subwavelength resonances can be achieved when the dielectric permittivity ε is large and the azimuthal field variations are slow (namely of small principal number m). Material absorption is related to the material loss tangent: $Q_{abs} \sim \mathrm{Re}\{\varepsilon\}/\mathrm{Im}\{\varepsilon\}$. Mode-solving calculations for this type of disk resonances were performed using two independent methods: numerically, 2D finite-difference frequency-domain (FDFD) simulations (which solve Maxwell's Equations in frequency domain exactly apart for spatial discretization) were conducted with a resolution of *30pts/r*; analytically, standard separation of variables (SV) in polar coordinates was used.

Table 7

single disk	λ / r	Q^{abs}	Q^{rad}	Q
Re{ε}=147.7, m=2	**20.01 (20.00)**	**10103 (10075)**	**1988 (1992)**	**1661 (1663)**
Re{ε}=65.6, m=3	9.952 (9.950)	10098 (10087)	9078 (9168)	4780 (4802)

The results for two TE-polarized dielectric-disk subwavelength modes of $\lambda / r \geq 10$ are presented in Table 7. Table 7 shows numerical FDFD (and in parentheses analytical SV) results for the wavelength and absorption, radiation and total loss rates, for two different cases of subwavelength-disk resonant modes. Note that disk-material loss-tangent $\mathrm{Im}\{\varepsilon\}/\mathrm{Re}\{\varepsilon\}=10^{-4}$ was used. (The specific parameters corresponding to the plot in Fig. 15(a) are highlighted with bold in the table.) The two methods have excellent agreement and imply that for a properly designed resonant low-loss-dielectric object values of $Q_{rad} \geq 2000$ and $Q_{abs} \sim 10000$ are achievable. Note that for the 3D case the computational complexity would be immensely increased, while the physics would not be significantly different. For example, a spherical object of $\varepsilon=147.7$ has a whispering gallery mode with $m=2$, $Q_{rad}=13962$, and $\lambda/r=17$.

The required values of ε, shown in Table 7, might at first seem unrealistically large. However, not only are there in the microwave regime (appropriate for approximately meter-range coupling applications) many materials that have both reasonably high enough dielectric constants and low losses (e.g. Titania, Barium tetratitanate, Lithium tantalite etc.), but also ε could signify instead the effective index of other known subwavelength surface-wave systems, such as surface modes on surfaces of metallic materials or plasmonic (metal-like, negative-ε) materials or metallo-dielectric photonic crystals or plasmono-dielectric photonic crystals.

To calculate now the achievable rate of energy transfer between two disks 1 and 2, as shown in Fig. 15(b) we place them at distance D between their centers. Numerically, the FDFD mode-solver simulations give κ through the frequency splitting of the normal modes of the combined system ($\delta_E = 2\kappa$ from Eq.(4)), which are even and odd superpositions of the initial single-disk modes; analytically, using the expressions for the separation-of-variables eigenfields $\mathbf{E}_{1,2}(\mathbf{r})$ CMT gives κ through

$$\kappa = \omega_1 / 2 \cdot \int d^3 r \varepsilon_2(r) E_2^*(r) E_1(r) / \int d^3 r \varepsilon(r) \left|E_1(r)\right|^2,$$

where $\varepsilon_j(\mathbf{r})$ and $\varepsilon(\mathbf{r})$ are the dielectric functions that describe only the disk j (minus the constant ε_o background) and the whole space respectively. Then, for medium distances $D/r = 10-3$ and for non-radiative coupling such that $D<2r_C$, where $r_C = m\lambda/2\pi$ is the radius of the radiation caustic, the two methods agree very well, and we finally find , as shown in Table 8, strong-coupling factors in the range $U \sim 1-50$. Thus, for the analyzed examples, the achieved figure-of-merit values are large enough to be useful for typical applications, as discussed below.

Table 8

two disks	D/r	Q^{rad}	$Q=\omega/2\Gamma$	$\omega/2\kappa$	κ/Γ
Re{ε}=147.7, m=2	3	2478	1989	46.9 (47.5)	42.4 (35.0)
$\lambda/r \approx 20$	5	2411	1946	298.0 (298.0)	6.5 (5.6)
$Q^{abs} \approx 10093$	7	2196	1804	769.7 (770.2)	2.3 (2.2)
	10	**2017**	**1681**	**1714 (1601)**	**0.98 (1.04)**
Re{ε}=65.6, m=3	3	7972	4455	144 (140)	30.9 (34.3)
$\lambda/r \approx 10$	5	9240	4824	2242 (2083)	2.2 (2.3)
$Q^{abs} \approx 10096$	7	9187	4810	7485 (7417)	0.64 (0.65)

Note that even though particular examples are presented and analyzed above as examples of systems that use resonant electromagnetic coupling for wireless energy transfer, those of self-resonant conducting coils, capacitively-loaded resonant conducting coils, inductively-loaded resonant conducting rods and resonant dielectric disks, any system that supports an electromagnetic mode with its electromagnetic energy extending much further than its size can be used for transferring energy. For example, there can be many abstract geometries with distributed capacitances and inductances that support the desired kind of resonances. In some examples, the resonant structure can be a dielectric sphere. In any one of these geometries, one can choose certain parameters to increase and/or optimize U or, if the Q's are limited by external factors, to increase and/or optimize for k or, if other system performance parameters are of importance, to optimize those.

Illustrative example

In one example, consider a case of wireless energy transfer between two identical resonant conducting loops, labeled by L_1 and L_2. The loops are capacitively-loaded and couple inductively via their mutual inductance. Let r_A denote the loops' radii, N_A their numbers of turns, and b_A the radii of the wires making the loops. We also denote by D_{12} the center-to-center separation between the loops.

Resonant objects of this type have two main loss mechanisms: ohmic absorption and far-field radiation. Using the same theoretical method as in previous sections, we find that for r_A=7cm, b_A=6mm, and N_A=15 turns, the quality factors for absorption, radiation and total loss are respectively, $Q_{A,abs} = \pi f / \Gamma_{A,abs} = 3.19 \times 10^4$, $Q_{A,rad} = \pi f / \Gamma_{A,rad} = 2.6 \times 10^8$ and $Q_A = \pi f / \Gamma_A = 2.84 \times 10^4$ at a resonant frequency $f = 1.8 \times 10^7$ Hz (remember that L_1 and L_2 are identical and have the same properties). $\Gamma_{A,abs}$, $\Gamma_{A,rad}$ are respectively the rates of absorptive and radiative loss of L_1 and L_2, and the rate of coupling between L_1 and L_2 is denoted by κ_{12}.

When the loops are in fixed distinct parallel planes separated by D_{12} =1.4m and have their centers on an axis (C) perpendicular to their planes, as shown in Fig. 17a(Left), the coupling factor for inductive coupling (ignoring retardation effects) is $k_{12} \equiv \kappa_{12} / \pi f = 7.68 \times 10^{-5}$, independent of time, and thus the strong-coupling factor is $U_{12} \equiv k_{12} Q_A = 2.18$. This configuration of parallel loops corresponds to the largest possible coupling rate κ_{12} at the particular separation D_{12}.

We find that the energy transferred to L_2 is maximum at time $\boldsymbol{T_* = \kappa t_* = \tan^{-1}(2.18) = 1.14 \Rightarrow t_* = 4775(1/f)}$ from Eq.(5), and constitutes $\boldsymbol{\eta_{E*} = 29}$% of the initial total energy from Eq.(6a), as shown in Fig. 17a(Right). The energies radiated and absorbed are respectively $\boldsymbol{\eta_{rad,E}(t_*) = 7.2}$% and $\boldsymbol{\eta_{abs,E}(t_*) = 58.1}$% of the initial total energy, with $\boldsymbol{5.8}$% of the energy remaining in L_1.

We would like to be able to further increase the efficiency of energy transfer between these two resonant objects at their distance D_{12}. In some examples, one can use an intermediate resonator between the source and device resonators, so that energy is transferred more efficiently from the source to the device resonator indirectly through the intermediate resonator.

3. Efficient energy-transfer by a chain of three 'strongly coupled' resonances

Fig. 16 shows a schematic that generally describes one example of the invention, in which energy is transferred wirelessly among three resonant objects. Referring to Fig. 16, energy is transferred, over a distance D_{1B}, from a resonant source object R_1 of characteristic size r_1 to a resonant intermediate object R_B of characteristic size r_B, and then, over an additional distance D_{B2}, from the resonant intermediate object R_B to a resonant device object R_2 of characteristic size r_2, where $D_{1B}+D_{B2}=D$. As described above, the source object R_1 can be supplied power from a power generator that is coupled to the source object R_1. In some examples, the power generator can be wirelessly, e.g., inductively, coupled to the source object R_1. In some examples, the power generator can be connected to the source object R_1 by means of a wire or cable. Also, the device object R_2 can be connected to a power load that consumes energy transferred to the device object R_2. For example, the device object can be connected to e.g. a resistor, a battery, or other device. All objects are resonant objects. The wireless near-field energy transfer is performed using the field (e.g. the electromagnetic field or acoustic field) of the system of three resonant objects.

Different temporal schemes can be employed, depending on the application, to transfer energy among three resonant objects. Here we will consider a particularly simple but important scheme: a one-time finite-amount energy-transfer scheme

3.1 Finite-amount energy-transfer efficiency

Let again the source, intermediate and device objects be 1, B, 2 respectively and their resonance modes, which we will use for the energy exchange, have angular frequencies $\omega_{1,B,2}$, frequency-widths due to intrinsic (absorption, radiation etc.) losses $\Gamma_{1,B,2}$ and (generally) vector fields $\mathbf{F}_{1,B,2}(\mathbf{r})$, normalized to unity energy. Once the three resonant objects are brought in proximity, they can interact and an appropriate analytical framework for modeling this resonant interaction is again that of the well-known coupled-mode theory (CMT), which can give a good description of the system for resonances having quality factors of at least, for example, 10. Then, using $e^{-i\omega t}$ time dependence, for the chain arrangement shown in Fig.16, the field amplitudes can be shown to satisfy, to lowest order:

$$\begin{aligned}
\frac{d}{dt}a_1(t) &= -i(\omega_1 - i\Gamma_1)a_1(t) + i\kappa_{11}a_1(t) + i\kappa_{1B}a_B(t) \\
\frac{d}{dt}a_B(t) &= -i(\omega_B - i\Gamma_B)a_B(t) + i\kappa_{BB}a_B(t) + i\kappa_{B1}a_1(t) + i\kappa_{B2}a_2(t) \\
\frac{d}{dt}a_2(t) &= -i(\omega_2 - i\Gamma_2)a_2(t) + i\kappa_{22}a_2(t) + i\kappa_{2B}a_B(t)
\end{aligned} \tag{34}$$

where $\kappa_{11,BB,22}$ are the shifts in each object's frequency due to the presence of the other, which are a second-order correction and can be absorbed into the eigenfrequencies by setting $\omega_{1,B,2} \to \omega_{1,B,2} + \kappa_{11,BB,22}$, and κ_{ij} are the coupling coefficients, which from the reciprocity requirement of the system must satisfy $\kappa_{ij} = \kappa_{ji}$. Note that, in some examples, the direct coupling coefficient κ_{12} between the resonant objects 1 and 2 may be much smaller than the coupling coefficients κ_{1B} and κ_{B2} between these two resonant objects with the intermediate object B, implying that the direct energy transfer between 1 and 2 is substantially dominated by the indirect energy transfer through the intermediate object. In some examples, if the coupling rates κ_{1B} and κ_{B2} are at least 5 times larger than the direct coupling rate κ_{12}, using an intermediate resonator may lead to an improvement in terms of energy transfer efficiency, as the indirect transfer may dominate the direct transfer. Therefore, in the CMT Eqs.(34) above, the direct coupling coefficient κ_{12} has been ignored, to analyze those particular examples.

The three resonant modes of the combined system are found by substituting $[a_1(t)\ a_B(t), a_2(t)] = [A_1, A_B, A_2]e^{-i\bar{\omega}t}$. When the resonators 1 and 2 are at exact resonance $\omega_1 = \omega_2 = \omega_A$ and for $\Gamma_1 = \Gamma_2 = \Gamma_A$, the resonant modes have complex resonant frequencies

$$\bar{\omega}_\pm = \omega_{AB} \pm \sqrt{(\Delta\omega_{AB})^2 + \kappa_{1B}^2 + \kappa_{B2}^2} \text{ and } \bar{\omega}_{ds} = \omega_A - i\Gamma_A \tag{35a}$$

where $\omega_{AB}=\left[\left(\omega_A+\omega_B\right)-i\left(\Gamma_A+\Gamma_B\right)\right]/2$, $\Delta\omega_{AB}=\left[\left(\omega_A-\omega_B\right)-i\left(\Gamma_A-\Gamma_B\right)\right]/2$ and whose splitting we denote as $\tilde{\delta}_E \equiv \overline{\omega}_+ - \overline{\omega}_-$, and corresponding resonant field amplitudes

$$\vec{V}_\pm=\begin{bmatrix}A_1\\A_B\\A_2\end{bmatrix}_\pm=\begin{bmatrix}\kappa_{1B}\\ \Delta\omega_{AB}\mp\sqrt{\left(\Delta\omega_{AB}\right)^2+\kappa_{1B}^2+\kappa_{B2}^2}\\ \kappa_{B2}\end{bmatrix} \text{ and } \vec{V}_{ds}=\begin{bmatrix}A_1\\A_B\\A_2\end{bmatrix}_{ds}=\begin{bmatrix}-\kappa_{B2}\\0\\ \kappa_{1B}\end{bmatrix} \quad (35b)$$

Note that, when all resonators are at exact resonance $\omega_1=\omega_2(=\omega_A)=\omega_B$ and for $\Gamma_1=\Gamma_2(=\Gamma_\mathrm{A})=\Gamma_\mathrm{B}$, we get $\Delta\omega_\mathrm{AB}=0$, $\tilde{\delta}_E=2\sqrt{\kappa_{1B}^2+\kappa_{B2}^2}$, and then

$$\overline{\omega}_\pm=\omega_A\pm\sqrt{\kappa_{1B}^2+\kappa_{B2}^2}-i\Gamma_A \text{ and } \overline{\omega}_{ds}=\omega_A-i\Gamma_A \quad (36a)$$

$$\vec{V}_\pm=\begin{bmatrix}A_1\\A_B\\A_2\end{bmatrix}_\pm=\begin{bmatrix}\kappa_{1B}\\ \mp\sqrt{\kappa_{1B}^2+\kappa_{B2}^2}\\ \kappa_{B2}\end{bmatrix} \text{ and } \vec{V}_{ds}=\begin{bmatrix}A_1\\A_B\\A_2\end{bmatrix}_{ds}=\begin{bmatrix}-\kappa_{B2}\\0\\ \kappa_{1B}\end{bmatrix}, \quad (36b)$$

namely we get that the resonant modes split to a lower frequency, a higher frequency and a same frequency mode.

Assume now that at time $t=0$ the source object 1 has finite energy $|a_1(0)|^2$, while the intermediate and device objects have $|a_B(0)|^2=|a_2(0)|^2=0$. Since the objects are coupled, energy will be transferred from 1 to B and from B to 2. With these initial conditions, Eqs.(34) can be solved, predicting the evolution of the field-amplitudes. The energy-transfer efficiency will be $\tilde{\eta}_E \equiv |a_2(t)|^2/|a_1(0)|^2$. The ratio of energy converted to loss due to a specific loss mechanism in resonators 1, B and 2, with respective loss rates $\Gamma_{1,\mathrm{loss}}$, $\Gamma_{\mathrm{B,loss}}$ and $\Gamma_{2,\mathrm{loss}}$ will be $\tilde{\eta}_{loss,E}=\int_0^t d\tau\left[2\Gamma_{1,\mathrm{loss}}|a_1(\tau)|^2+2\Gamma_{\mathrm{B,loss}}|a_B(\tau)|^2+2\Gamma_{2,\mathrm{loss}}|a_2(\tau)|^2\right]/|a_1(0)|^2$. At exact resonance $\omega_1 = \omega_2 (= \omega_A) = \omega_B$ (an optimal condition) and in the special symmetric case $\Gamma_1 = \Gamma_2 = \Gamma_A$ and $\kappa_{1B} = \kappa_{B2} = \kappa$, the field amplitudes are

$$a_1(\tilde{T}) = \frac{1}{2} e^{-i\omega_A t} e^{-\tilde{T}/\tilde{U}} \left[\tilde{\Delta} \frac{\sin(\sqrt{1-\tilde{\Delta}^2}\tilde{T})}{\sqrt{1-\tilde{\Delta}^2}} + \cos(\sqrt{1-\tilde{\Delta}^2}\tilde{T}) + e^{-\tilde{\Delta}\tilde{T}} \right] \quad (37a)$$

$$a_B(\tilde{T}) = \frac{1}{2} e^{-i\omega_A t} e^{-\tilde{T}/\tilde{U}} \frac{\sin(\sqrt{1-\tilde{\Delta}^2}\tilde{T})}{\sqrt{1-\tilde{\Delta}^2}} \quad (37b)$$

$$a_2(\tilde{T}) = \frac{1}{2} e^{-i\omega_A t} e^{-\tilde{T}/\tilde{U}} \left[\tilde{\Delta} \frac{\sin(\sqrt{1-\tilde{\Delta}^2}\tilde{T})}{\sqrt{1-\tilde{\Delta}^2}} + \cos(\sqrt{1-\tilde{\Delta}^2}\tilde{T}) - e^{-\tilde{\Delta}\tilde{T}} \right] \quad (37c)$$

where $\tilde{T} \equiv \sqrt{2}\kappa t$, $\tilde{\Delta}^{-1} = 2\sqrt{2}\kappa/(\Gamma_B - \Gamma_A)$ and $\tilde{U} = 2\sqrt{2}\kappa/(\Gamma_A + \Gamma_B)$.

In some examples, the system designer can adjust the duration of the coupling t at will. In some examples, the duration t can be adjusted to maximize the device energy (and thus efficiency $\tilde{\eta}_E$). Then, in the special case above, it can be inferred from Eq.(37c) that $\tilde{\eta}_E$ is maximized for the $\tilde{T} = \tilde{T}_*$ that satisfies

$$\left[\tilde{\Delta} - \tilde{U}\left(1-\tilde{\Delta}^2\right)\right] \frac{\sin(\sqrt{1-\tilde{\Delta}^2}\tilde{T})}{\sqrt{1-\tilde{\Delta}^2}} + \left(1-\tilde{\Delta}\tilde{U}\right)\left[\cos(\sqrt{1-\tilde{\Delta}^2}\tilde{T}) - e^{\tilde{\Delta}\tilde{T}}\right] = 0 . \quad (38)$$

In general, this equation may not have an obvious analytical solution, but it does admit a simple solution in the following two cases:

When $\Gamma_A = \Gamma_B \Leftrightarrow \tilde{\Delta} = 0, \tilde{U} = \sqrt{2}\kappa/\Gamma_B$, the energy transfer from resonator 1 to resonator 2 is maximized at

$$\tilde{T}_*\left(\tilde{\Delta} = 0\right) = 2\tan^{-1}\tilde{U} \quad (39)$$

resulting in an energy-transfer efficiency

$$\tilde{\eta}_E\left(\tilde{T}_*, \tilde{\Delta} = 0\right) = \left[\frac{\tilde{U}^2}{1+\tilde{U}^2} \exp\left(-\frac{2\tan^{-1}\tilde{U}}{\tilde{U}}\right)\right]^2, \quad (40)$$

which has the form of the two-object energy-transfer efficiency of Eq.(6b) but squared.

When $\Gamma_A = 0 \Leftrightarrow \tilde{\Delta}^{-1} = \tilde{U} = 2\sqrt{2}\kappa/\Gamma_B$, the energy transfer from resonator 1 to resonator 2 is maximized at

$$\tilde{T}_*\left(\tilde{\Delta}^{-1}=\tilde{U}\right)=\begin{cases}\pi\tilde{U}\Big/\sqrt{\tilde{U}^2-1}, \tilde{U}>1\\ \infty, \tilde{U}\le 1\end{cases} \tag{41}$$

resulting in an energy-transfer efficiency

$$\eta_E\left(\tilde{T}_*, \tilde{\Delta}^{-1}=\tilde{U}\right)=\frac{1}{4}\cdot\begin{cases}\left[1+\exp\left(-\pi\Big/\sqrt{\tilde{U}^2-1}\right)\right]^2, \tilde{U}>1\\ 1, \tilde{U}\le 1\end{cases}. \tag{42}$$

In both cases, and in general for any $\tilde{\Delta}$, the efficiency is an increasing function of $\tilde{U}$. Therefore, once more resonators that have high quality factors are preferred. In some examples, one may design resonators with Q>50. In some examples, one may design resonators with Q>100.

Illustrative example

Returning to our illustrative example, in order to improve the ~29% efficiency of the energy transfer from L_1 to L_2, while keeping the distance D_{12} separating them fixed, we propose to introduce an intermediate resonant object that couples strongly to both L_1 and L_2, while having the same resonant frequency as both of them. In one example, we take that mediator to also be a capacitively-loaded conducting-wire loop, which we label by L_B. We place L_B at equal distance (D_{1B}=D_{B2}=D_{12}/2=0.7m) from L_1 and L_2 such that its axis also lies on the same axis (C), and we orient it such that its plane is parallel to the planes of L_1 and L_2, which is the optimal orientation in terms of coupling. A schematic diagram of the three-loops configuration is depicted in Fig. 17b(Left).

In order for L_B to couple strongly to L_1 and L_2, its size needs to be substantially larger than the size of L_1 and L_2. However, this increase in the size of L_B has considerable drawback in the sense that it may also be accompanied by significant decrease in its radiation quality factor. This feature may often occur for the resonant systems of this type: stronger coupling can often be enabled by increasing the objects' size, but it may imply stronger radiation from the object in question. Large radiation may often be undesirable, because it could lead to far-field interference with other RF systems, and in some systems also because of safety concerns. For r_B=70cm, b_B=1.5cm, and N_B=1 turn, we get $Q_{B,abs} = \pi f / \Gamma_{B,abs} = 7706$, $Q_{B,rad} = \pi f / \Gamma_{B,rad} = 400$, so $Q_B = \pi f / \Gamma_B = 380$, and $k_{1B} \equiv \kappa_{1B} / \pi f = k_{B2} = 0.0056$, so $\tilde{U} = 2\sqrt{2}\kappa / (\Gamma_A + \Gamma_B) = 5.94$ and $\tilde{\Delta}^{-1} = 2\sqrt{2}\kappa / (\Gamma_B - \Gamma_A) = 6.1$, at $f = 1.8 \times 10^7$ Hz. Note that since the coupling rates κ_{1B} and κ_{B2} are $\approx$70 times larger than κ_{12}, indeed we can ignore the direct coupling between L_1 and L_2, and focus only on the indirect energy transfer through the intermediate loop L_B, therefore the analysis of the previous section can be used.

The optimum in energy transferred to L_2 occurs at time $\tilde{T}_* = \sqrt{2}\kappa t_* = 3.21 \Rightarrow t_* = 129(1/f)$, calculated from Eq.(38), and is equal to $\tilde{\eta}_{E*} = 61.5\%$ of the initial energy from Eq.(37c). [Note that, since $Q_A \gg Q_B$, we could have used the analytical conclusions of the case $\tilde{\Delta}^{-1} = \tilde{U}$ and then we would have gotten a very close approximation of $\tilde{T}_* = 3.19$ from Eq.(41).] The energy radiated is $\tilde{\eta}_{rad,E}(t_*) = 31.1\%$, while the energy absorbed is $\tilde{\eta}_{abs,E}(t_*) = 3.3\%$, and 4.1% of the initial energy is left in L_1. Thus, the energy transferred, now indirectly, from L_1 to L_2 has increased by factor of 2 relative to the two-loops direct transfer case. Furthermore, the transfer time in the three-loops case is now $\approx$35 times shorter than in the two-loops direct transfer, because of the stronger coupling rate. The dynamics of the energy transfer in the three-loops case is shown in Fig. 17b(Right).

Note that the energy radiated in the three-loop system has undesirably increased by factor of 4 compared to the two-loop system. We would like to be able to achieve a similar improvement in energy-transfer efficiency, while not allowing the radiated energy to increase. In this specification, we disclose that, in some examples, this can be achieved by appropriately varying the values of the coupling strengths between the three resonators.

4. Efficient energy-transfer by a chain of three resonances with adiabatically varying coupling strengths

Consider again the system of a source resonator R_1, a device resonator R_2 and an intermediate resonator R_B. For the purposes of the present analysis, R_1 and R_2 will be assumed to have negligible mutual interactions with each other, while each of them can be strongly coupled to R_B, with coupling rates κ_{1B} and κ_{B2} respectively. Note that, in some examples of wireless energy transfer systems, for the resonant object R_B to be able to have strong coupling with other resonant objects, it may be accompanied with inferior loss properties compared to R_1 and R_2, usually in terms of substantially larger radiation losses.

It was seen in a previous section that, when the resonators 1 and 2 are at exact resonance $\omega_1 = \omega_2 = \omega_A$ and for $\Gamma_1 = \Gamma_2 = \Gamma_A$, the system supports a resonant state (eigenstate) with resonant frequency (eigenfrequency) $\bar{\omega}_{ds} = \omega_A - i\Gamma_A$ and resonant field amplitudes $\vec{V}_{ds} = \begin{bmatrix}-\kappa_{B2} & 0 & \kappa_{1B}\end{bmatrix}^T \Big/ \sqrt{\kappa_{1B}^2 + \kappa_{B2}^2}$, where here we normalized it. This eigenstate $\vec{V}_{ds}$ we call the "dark state" in analogy with atomic systems and the related phenomenon of Electromagnetically Induced Transparency (EIT), wherein complete population transfer between two quantum states through a third lossy state, coupled to each of the other two states, is enabled. The dark state is the most essential building block of our proposed efficient weakly-radiative energy-transfer scheme, because it has no energy at all in the intermediate (lossy) resonator R_B, i.e. $a_B(t)=0$ $\forall$ t whenever the three-object system is in state $\vec{V}_{ds}$. In fact, if $\Gamma_A \rightarrow 0$, then this state is completely lossless, or if $\Gamma_{A,rad} \rightarrow 0$, then this state is completely non-radiative. Therefore, we disclose using predominantly this state to implement the wireless energy transfer. By doing that, the proposed energy transfer scheme can be made completely lossless, in the limit $\Gamma_A \rightarrow 0$, no matter how large is the loss rate Γ_B, as shown in Fig. 20, or completely non-radiative, in the limit $\Gamma_{A,rad} \rightarrow 0$, no matter how large is the radiative loss rate $\Gamma_{B,rad}$. Since the energy transfer efficiency increases as the quality factors of the first (source) and second (device) resonances increase, one may design these resonators to have a high quality factor. In some examples, one may design resonators with Q_A>50. In some examples, one may design resonators with Q_A>100.

Let us demonstrate how it is possible to use the dark state for energy transfer with minimal loss. From the expression of $\vec{V}_{ds}$ one can see that, if the three-object system is in the state $\vec{V}_{ds}$, then, in general, there is energy in the source resonator with field amplitude proportional to the coupling rate κ_{B2} between the device resonator and the intermediate resonator, and there is also energy in the device resonator with field amplitude proportional to the coupling rate κ_{1B} between the source resonator and the intermediate resonator. Then, $\kappa_{1B}=0$ corresponds to all the system's energy being in R_1, while $\kappa_{B2}=0$ corresponds to all the system's energy being in R_2.

So, the important considerations necessary to achieve efficient weakly-radiative energy transfer, consist of preparing the system initially in state $\vec{V}_{ds}$ and varying the coupling rates in time appropriately to evolve this state $\vec{V}_{ds}$ in a way that will cause energy transfer. Thus, if at t=0 all the energy is in R_1, then one should have $\kappa_{1B}(t=0) \ll \kappa_{B2}(t=0)$, for example $\kappa_{1B}(t=0)=0$ and $\kappa_{B2}(t=0)\neq 0$. In order for the total energy of the system to end up in R_2, at a time t_{EIT} when the full variation of the coupling rates has been completed, we should have $\kappa_{1B}(t=t_{EIT}) \gg \kappa_{B2}(t=t_{EIT})$, for example $\kappa_{1B}(t=t_{EIT})\neq 0$ and $\kappa_{B2}(t=t_{EIT})=0$. This ensures that the initial and final states of the three-object system are parallel to $\vec{V}_{ds}$. However, a second important consideration is to keep the three-object system at all times in $\vec{V}_{ds}(t)$, even as $\kappa_{1B}(t)$ and $\kappa_{B2}(t)$ are varied in time. This is crucial in order to prevent the system's energy from getting into any of the two other eigenstates $\vec{V}_{\pm}$ and thus getting into the intermediate object R_B, which may be highly radiative or lossy in general, as in the example of Fig. 17. This consideration requires changing $\kappa_{1B}(t)$ and $\kappa_{B2}(t)$ slowly enough so as to make the entire three-object system adiabatically follow the time evolution of $\vec{V}_{ds}(t)$. The criterion for adiabatic following can be expressed, in analogy to the population transfer case as

$$\left|\left\langle \vec{V}_{\pm} \left| \frac{d\vec{V}_{ds}}{dt} \right.\right\rangle\right| \ll \left|\bar{\omega}_{\pm} - \bar{\omega}_{ds}\right| \qquad (43)$$

where $\vec{V}_{\pm}$ are the remaining two eigenstates of the system, with corresponding eigenvalues $\bar{\omega}_{\pm}$, as shown earlier. Note that any functional dependence of the coupling rates with time (with duration parameter t_{EIT}) will work, provided it satisfies the adiabaticity criterion Eq.(43) above. The time functional can be linear, sinusoidal (as in the illustrative example to follow) or the temporal analog of a Butterworth (maximally flat) taper, a Chebyshev taper, an exponential taper and the like.

Referring to Fig. 18, an example coupling rate adjustment system 300 for adjusting coupling rates for the one or more of the resonator structures R_1, R_2, or R_B is shown. As described, the coupling rates between the first and intermediate resonator structures and the intermediate and second resonator structures are characterized by κ_{1B} and κ_{B2} respectively. These coupling rates, κ_{1B} and κ_{B2}, are several times (e.g., 70 times) greater than the coupling rate κ_{12} between the first and second resonator structure. In some examples, the coupling rate adjustment system can be a mechanical, electrical, or electro-mechanical system for dynamically adjusting, e.g., rotating, or effecting a translational movement, of the one or more resonator structures with respect to each other.

In some examples, the coupling rate κ_{1B} is much smaller than the coupling rate the coupling rate κ_{B2} at the beginning of the energy transfer. By the end, i.e., when a substantial amount of energy has been transferred from the first resonator structure R_1 to the second resonator structure, R_2, the coupling rate κ_{1B} is much greater than the coupling rate κ_{B2}. In some examples, the coupling rate κ_{1B} can be set to a fixed value while the coupling rate κ_{B2} is being varied from its maximum to its minimum value. In some examples, the coupling rate κ_{B2} can be set to a fixed value while the coupling rate κ_{1B} is being varied from its minimum to its maximum value. In some examples, the coupling rate κ_{1B} can be varied from a minimum to a maximum value while the coupling rate κ_{B2} is being varied from its maximum to minimum value.

Referring now to Fig. 19, a graph for implementing an example coupling rate adjustment system 300 is shown. As shown, in some examples, the coupling rate κ_{1B} is set at its minimum value at time, t=0, and increased as a function of time (see, for example, equation 44), while the coupling rate κ_{B2} is at its maximum value at t=0 and decreased as a function of time (see, for example, equation 45). Accordingly, at the beginning (t=0), the value of κ_{1B} is much smaller than the value of κ_{B2}. In some examples, the value of κ_{1B} can be selected to be any value much smaller than the value of κ_{B2}. During the wireless energy transfer, the value of κ_{1B} is increased, while the value of κ_{B2} is decreased. After a predetermined amount of time t_{EIT} has elapsed (e.g., after a substantial amount of energy has been transferred to the second resonator), the value of κ_{1B} becomes much greater than the value of κ_{B2}.

In some implementations, the coupling rate adjustment system 300 can effect an adjustment of coupling rates between the resonator structures by changing a relative orientation of one or more of the resonator structures with respect to each other. For example, referring again to Fig. 18, the first and second resonator structures, R_1 and R_2, can be rotated about their respective axes (e.g., varying angles θ_1 and θ_2), with respect to the intermediate resonator structure R_B to simultaneously change κ_{1B} and κ_{B2}. Alternatively, the orientation of the intermediate resonator structure, R_B, can be adjusted, e.g., rotated about an axis, with respect to the first and second resonator structures to simultaneously change κ_{1B} and κ_{B2}. Alternatively, the orientation of only the first resonator structure R_1 can be rotated about its respective axis to change κ_{1B}, while the orientations of R_2 and R_B are fixed and thus κ_{B2} is fixed to a value intermediate between the minimum and maximum values of κ_{1B}. Alternatively, the orientation of only the second resonator structure R_2 can be rotated about its respective axis to change κ_{B2}, while the orientations of R_1 and R_B are fixed and thus κ_{1B} is fixed to a value intermediate between the minimum and maximum values of κ_{B2}.

In some implementations, the coupling rate adjustment system 300 can effect an adjustment of coupling rates between the resonator structures by translationally moving one or more of the resonator structures with respect to each other. For example, the positions of the first and second resonator structures, R_1 and R_2, can be adjusted, e.g., moved along an axis, with respect to the intermediate resonator structure R_B to simultaneously change κ_{1B} and κ_{B2}. Alternatively, a position of the intermediate resonator structure, R_B, can be adjusted, e.g., moved along an axis, with respect to the first and second resonator structures to simultaneously change κ_{1B} and κ_{B2}. Alternatively, a position of only the first resonator structure, R_1, can be adjusted, e.g., moved along an axis, with respect to the intermediate R_B resonator structure to change κ_{1B}, while the positions of R_2 and R_B are fixed and thus κ_{B2} is fixed to a value intermediate between the minimum and maximum values of κ_{1B}. Alternatively, a position of only the second resonator structure, R_2, can be adjusted, e.g., moved along an axis, with respect to the intermediate R_B resonator structure to change κ_{B2}, while the positions of R_1 and R_B are fixed and thus κ_{1B} is fixed to a value intermediate between the minimum and maximum values of κ_{B2}.

In some examples, the coupling rate adjustment system 300 can dynamically adjust an effective size of the resonator objects to effect adjustments in the coupling rates κ_{1B} and κ_{B2} similar to that described above. The effective size can be adjusted by changing a physical size of the resonator objects. For example, the physical size can be adjusted by effecting mechanical changes in area, length, or other physical aspect of one or more of the resonator structures. The effective size can also be adjusted through non-mechanical changes, such as, but not limited to, applying a magnetic field to change the permeability of the one or more of the resonator structures.

In principle, one would think of making the transfer time t_{EIT} as long as possible to ensure adiabaticity. However there is limitation on how slow the transfer process can optimally be, imposed by the losses in R_1 and R_2. Such a limitation may not be a strong concern in typical atomic EIT case, because the initial and final states there can be chosen to be non-lossy ground states. However, in our case, losses in R_1 and R_2 are not avoidable, and can be detrimental to the energy transfer process whenever the transfer time t_{EIT} is not less than $1/\Gamma_A$. This is because, even if the three-object system is carefully kept in $\vec{V}_{ds}$ at all times, the total energy of the system will decrease from its initial value as a consequence of losses in R_1 and R_2. Thus the duration of the transfer may be a compromise between these two limits: the desire to keep t_{EIT} long enough to ensure near-adiabaticity, but short enough not to suffer from losses in R_1 and R_2.

Given a particular functional variation of the coupling rates with time with variation duration parameter t_{EIT}, one may calculate the optimal energy transfer efficiency in the following way: First, for each t_{EIT}, determine the optimal time t_*, at which the energy at R_2 is maximized and the transfer process may be be terminated. Then find the optimal variation time t_{EIT*} based on the counteracting mechanisms discussed above. The optimal efficiency of energy transfer $\tilde{\eta}_{EIT,E*}$ can then be calculated. In most cases, this procedure may need to be done numerically using the CMT Eqs.(34) as analytical solutions may not be possible. With respect to optimizing the functional dependence of the coupling rates with time, one may choose one that minimizes the coupling of energy to the eigenstates $\vec{V}_\pm$ for a given t_{EIT}, which may lead to the temporal analog of a Chebyshev taper.

In some examples, the optimal t_{EIT} may not be long enough for the adiabadicity criterion of Eq.(43) to be always satisfied. In those cases, some energy may get into at least one of the lossy states $\vec{V}_\pm$. Still significant improvement in efficiency and radiation loss may be achieved by the mode of operation where the coupling rates are variable, compared to the mode of operation where the coupling rates are constant, provided the maximum energy that enters the states $\vec{V}_\pm$ is much less in the variable rate scenario than in the constant rate scenario. In examples, using the proposed scheme of time-varying coupling rates may be advantageous as long as the maximum energy stored in the intermediate resonator is substantially small. In some examples, substantially small may be at most 5% of the peak total energy of the system. In some examples, substantially small may be at most 10% of the peak total energy of the system.

We can now also see why the mode of operation of the system where the coupling rates are kept constant in time may cause a considerable amount of lost (and especially radiated) energy, compared to the proposed mode of operation where the coupling rates are varied adiabatically in time. The reason is that, when $\kappa_{1B}=\kappa_{B2}$=const, the energies in R_1 and R_2 will always be equal to each other if the three-object system is to stay in $\vec{V}_{ds}$. So one cannot transfer energy from R_1 to R_2 by keeping the system purely in state $\vec{V}_{ds}$; note that even the initial state of the system, in which all the energy is in R_1 and there is no energy in R_3, cannot be solely in $\vec{V}_{ds}$ for fixed nonzero κ_{1B} and κ_{B2}, and has nonzero components along the eigenstates $\vec{V}_{\pm}$ which implies a finite energy will build up in R_B, and consequently result in an increased radiation, especially if $\Gamma_{B,rad} \gg \Gamma_{A,rad}$, which may be the case if the resonator R_B is chosen large enough to couple strongly to R_1 and R_2, as explained earlier.

Illustrative example

The previous analysis explains why a considerable amount of energy was radiated when the inductive coupling rates of the loops were kept constant in time, like in Fig. 17b. Let us now consider the modifications necessary for an adiabatically-varied-κ three-loops indirect transfer scheme, as suggested in the previous section, aiming to reduce the total radiated energy back to its reasonable value in the two-loops direct transfer case (Fig.17a), while maintaining the total energy transfer at level comparable to the constant-κ three-loops indirect transfer case (Fig. 17b). In one example, shown in Fig. 17c(Left and Right), we will keep the orientation of L_B fixed, and start initially (t=0) with L_1 perpendicular to L_B for $\kappa_{1B}=0$ and L_2 parallel to L_B for κ_{B2}=max, then uniformly rotate L_1 and L_2, at the same rates, until finally, at ($t=t_{EIT}$), L_1 becomes parallel to L_B for κ_{1B}=max and L_2 perpendicular to L_B for $\kappa_{B2}=0$, where we stop the rotation process. In this example, we choose a sinusoidal temporal variation of the coupling rates:

$$\kappa_{1B}(t)=\kappa\sin(\pi t/2t_{EIT}) \tag{44}$$

$$\kappa_{B2}(t)=\kappa\cos(\pi t/2t_{EIT}) \tag{45}$$

for $0<t<t_{EIT}$, and $k_{1B}\equiv\kappa_{1B}/\pi f=k_{B2}=0.0056$ as before. By using the same CMT analysis as in Eq. (34), we find, in Fig. 17c(Center), that for an optimal $t_{EIT*}=1989(1/f)$, an optimum transfer of $\tilde{\eta}_{EIT,E*}=61.2\%$ can be achieved at $t_*=1796(1/f)$, with only 8.2% of the initial energy being radiated, 28.6% absorbed, and 2% left in L_1. This is quite remarkable: by simply rotating the loops during the transfer, the energy radiated has dropped by factor of 4, while keeping the same 61% level of the energy transferred.

This considerable decrease in radiation may seem surprising, because the intermediate resonator L_B, which mediates all the energy transfer, is highly radiative ($\approx$ 650 times more radiative than L_1 and L_2), and there is much more time to radiate, since the whole process lasts 14 times longer than in Fig. 17b. Again, the clue to the physical mechanism behind this surprising result can be obtained by observing the differences between the curves describing the energy in R_B in Fig. 17b and Fig. 17c. Unlike the case of constant coupling rates, depicted in Fig. 17b, where the amount of energy ultimately transferred to L_2 goes first through the intermediate loop L_B, with peak energy storage in L_B as much as 30% of the peak total energy of the system, in the case of time-varying coupling rates, shown in Fig. 17c, there is almost little or no energy in L_B at all times during the transfer. In other words, the energy is transferred quite efficiently from L_1 to L_2, mediated by L_B without considerable amount of energy ever being in the highly radiative intermediate loop L_B. (Note that direct transfer from L_1 to L_2 is identically zero here since L_1 is always perpendicular to L_2, so all the energy transfer is indeed mediated through L_B). In some examples, improvement in efficiency and/or radiated energy can still have been accomplished if the energy transfer had been designed with a time t_{EIT} smaller than its optimal value (perhaps to speed up the process), if the maximum energy accumulated inside the intermediate resonator was less than 30%. For example, improvement can have been achieved for maximum energy accumulation inside the intermediate resonator of 5% or even 10%.

An example implementation of the coupling rate adjustment system 300 is described below, where the resonators are capacitively-loaded loops, which couple to each other inductively. At the beginning (t=0), the coupling rate adjustment system 300 sets the relative orientation of the first resonator structure L_1 to be perpendicular to the intermediate resonator structure L_B. At this orientation, the value of the coupling rate κ_{1B} between the first and intermediate resonator structure is at its minimum value. Also, the coupling rate adjustment system 300 can set the relative orientation of the second resonator structure L_2 to be parallel to the intermediate resonator structure L_B. At this orientation, the value of the coupling rate κ_{B2} is at a maximum value. During wireless energy transfer, the coupling rate adjustment system 300 can effect the rotation of the first resonator structure L_1 about its axis so that the value of κ_{1B} is increased. In some examples, the coupling rate adjustment system 300 can also effect the rotation of the second resonator structure, L_2, about its axis so that the value of κ_{B2} is decreased. In some examples, a similar effect can be achieved by fixing L_1 and L_2 to be perpendicular to each other and rotating only L_B to be parallel to L_2 and perpendicular to L_1 at t=0 and parallel to L_1 and perpendicular to L_2 at $t=t_{EIT}$. In some examples, a similar effect can be achieved by fixing L_B and one of L_1 and L_2 (e.g., L_1) at a predetermined orientation (e.g. at 45 degrees with respect to the intermediate resonator L_B) and rotating only the other of L_1 and L_2 (e.g., L_2 from parallel to L_B at t=0 to perpendicular to L_B at $t=t_{EIT}$).

Similarly, in some implementations, at the beginning (t=0), the coupling rate adjustment system 300 can set the position of the first resonator structure L_1 at a first large predetermined distance from the intermediate resonator structure L_B so that the value of the coupling rate κ_{1B} is at its minimum value. Correspondingly, the coupling rate adjustment system 300 can set the position of the second resonator structure L_2 at a second small predetermined distance from the intermediate resonator structure L_B so that the value of the coupling rate κ_{B2} between the first and intermediate resonator structure is at its maximum value. During wireless energy transfer, the coupling rate adjustment system 300 can affect the position of the first resonator structure L_1 to bring it closer to

L_B so that the value of κ_{1B} is increased. In some examples, the coupling rate adjustment system 300 can also effect the position of the second resonator structure, L_2, to take it away from L_B so that the value of κ_{B2} is decreased. In some examples, a similar effect can be achieved by fixing L_1 and L_2 to be at a fixed distance to each other and effecting the position of only L_B to be close to L_2 and away from L_1 at t=0 and close to L_1 and away from L_2 at $t=t_{EIT}$. In some examples, a similar effect can be achieved by fixing L_B and one of L_1 and L_2 (e.g., L_1) at a predetermined (not too large but not too small) distance and effecting the position only the other of L_1 and L_2 (e.g., L_2 from close to L_B at t=0 to away from L_B at $t=t_{EIT}$).

<u>5. Comparison of static and adiabatically dynamic systems</u>

In the abstract case of energy transfer from R_1 to R_2, where no constraints are imposed on the relative magnitude of κ, Γ_{rad}^{A}, Γ_{rad}^{B}, Γ_{abs}^{A}, and Γ_{abs}^{B}, it is not certain that the adiabatic-κ (EIT-like) system will always perform better than the constant-κ one, in terms of the transferred and radiated energies. In fact, there could exist some range of the parameters (κ, Γ_{rad}^{A}, Γ_{rad}^{B}, Γ_{abs}^{A}, Γ_{abs}^{B}), for which the energy radiated in the constant-κ transfer case is less than that radiated in the EIT-like case. For this reason, we investigate both the adiabatic-κ and constant-κ transfer schemes, as we vary some of the crucial parameters of the system. The percentage of energies transferred and lost (radiated+absorbed) depends only on the relative values of κ, $\Gamma_A = \Gamma_{rad}^{A} + \Gamma_{abs}^{A}$ and $\Gamma_B = \Gamma_{rad}^{B} + \Gamma_{abs}^{B}$. Hence we first calculate and visualize the dependence of these energies on the relevant parameters κ/Γ_B and Γ_B/Γ_A, in the contour plots shown in Fig. 21.

The way the contour plots are calculated is as follows. For each value of (κ/Γ_B, Γ_B/Γ_A) in the adiabatic case, where $\kappa_{1B}(t)$ and $\kappa_{B2}(t)$ are given by Eq. (44)-(45), one tries range of values of t_{EIT}. For each t_{EIT}, the maximum energy transferred E_2(%) over $0< t < t_{EIT}$, denoted by max(E_2, t_{EIT}), is calculated together with the total energy lost at that maximum transfer. Next the maximum of max(E_2, t_{EIT}) over all values of t_{EIT} is selected and plotted as single point on the contour plot in Fig. 21a. We refer to this point as the optimum energy transfer (%) in the adiabatic-κ case for the particular (κ/Γ_B, Γ_B/Γ_A) under consideration. We also plot in Fig. 21d the corresponding value of the total energy lost (%) at the optimum of E_2. We repeat these calculations for all pairs (κ/Γ_B, Γ_B/Γ_A) shown in the contour plots. In the constant-κ transfer case, for each (κ/Γ_B, Γ_B/Γ_A), the time evolution of E_2(%) and E_{lost} are calculated for $0<t<2/\kappa$, and optimum transfer, shown in Fig. 21b, refers to the maximum of $E_3(t)$ over $0<t<2/\kappa$. The corresponding total energy lost at optimum constant-transfer is shown in Fig. 21e. Now that we calculated the energies of interest as functions of (κ/Γ_B, Γ_B/Γ_A), we look for ranges of the relevant parameters in which the adiabatic-κ transfer has advantages over the constant-κ one. So, we plot the ratio of $(E_2)_{adiabatic-\kappa}/(E_2)_{constant-\kappa}$ in Fig. 21c, and $(E_{lost})_{constant-\kappa}$ / $(E_{lost})_{adiabatic-\kappa}$ in Fig. 21f. We find that, for $\Gamma_B/\Gamma_A>50$, the optimum energy transferred in the adiabatic-κ case exceeds that in the constant-κ case, and the improvement factor can be larger than 2. From Fig. 21f, one sees that the adiabatic-κ scheme can reduce the total energy lost by factor of 3 compared to the constant-κ scheme, also in the range Γ_B/Γ_A >50. As in the constant-κ case, also in the adiabatic-κ case the efficiency increases as the ratio of the maximum value, κ, of the coupling rates to the loss rate of the intermediate object (and thus also the first and second objects for $\Gamma_B/\Gamma_A>1$) increases. In some examples, one may design a system so that κ is larger than each of Γ_B and Γ_A. In some examples, one may design a system so that κ is at least 2 times larger than each of Γ_B and Γ_A. In some examples, one may design a system so that κ is at least 4 times larger than each of Γ_B and Γ_A.

Although one may be interested in reducing the total energy lost (radiated + absorbed) as much as possible in order to make the transfer more efficient, the undersirable nature of the radiated energy may make it important to consider reducing the energy radiated. For this purpose, we calculate the energy radiated at optimum transfer in both the adiabatic-κ and constant-κ schemes, and compare them. The relevant parameters in this case are κ/Γ_B, Γ_B/Γ_A, Γ_{rad}^A/Γ_A, and Γ_{rad}^B/Γ_B. The problem is more complex because the parameter space is now 4-dimensional. So we focus on those particular cross sections that can best reveal the most important differences between the two schemes. From Fig. 21c and 21f, one can guess that the best improvement in both E_2 and E_{lost} occurs for $\Gamma_B / \Gamma_A \geq 500$. Moreover, knowing that it is the intermediate object R_B that makes the main difference between the adiabatic-κ and constant-κ schemes, being "energy-empty" in the adiabatic-κ case and "energy-full" in the constant-κ one, we first look at the special situation where $\Gamma_{rad}^{A}=0$. In Fig. 22a and Fig. 22b, we show contour plots of the energy radiated at optimum transfer, in the constant-κ and adiabatic-κ schemes respectively, for the particular cross section having $\Gamma_B / \Gamma_A=500$ and $\Gamma_{rad}^{A}=0$. Comparing these two figures, one can see that, by using the adiabatic-κ scheme, one can reduce the energy radiated by factor of 6.3 or more.

To get quantitative estimate of the radiation reduction factor in the general case where $\Gamma_{A,rad} \neq 0$, we calculate the ratio of energies radiated at optimum transfers in both schemes, namely,

$$\frac{\left(E_{rad}\right)_{\text{constant}-\kappa}}{\left(E_{rad}\right)_{\text{adiabatic}-\kappa}}=\frac{2\int_0^{t_*^{\text{constant}-\kappa}}\left\{\frac{\Gamma_{rad}^{B}}{\Gamma_{rad}^{A}}\left|a_B^{\text{constant}-\kappa}(t)\right|^2+\left|a_1^{\text{constant}-\kappa}(t)\right|^2+\left|a_2^{\text{constant}-\kappa}(t)\right|^2\right\}}{2\int_0^{t_*^{\text{adiabatic}-\kappa}}\left\{\frac{\Gamma_{rad}^{B}}{\Gamma_{rad}^{A}}\left|a_B^{\text{adiabatic}-\kappa}(t)\right|^2+\left|a_1^{\text{adiabatic}-\kappa}(t)\right|^2+\left|a_2^{\text{adiabatic}-\kappa}(t)\right|^2\right\}} \quad (46)$$

which depends only on $\Gamma_{rad}^{B} / \Gamma_{rad}^{A}$, the time-dependent mode amplitudes, and the optimum transfer times in both schemes. The latter two quantities are completely determined by κ/Γ_B and Γ_B / Γ_A. Hence the only parameters relevant to the calculations of the ratio of radiated energies are $\Gamma_{rad}^{B} / \Gamma_{rad}^{A}$, κ/Γ_B and Γ_B / Γ_A, thus reducing the dimensionality of the investigated parameter space from down to 3. For convenience, we multiply the first relevant parameter $\Gamma_{rad}^{B} / \Gamma_{rad}^{A}$ by Γ_B/Γ_A which becomes $(\Gamma_{rad}^{B}/\Gamma_B)/(\Gamma_{rad}^{A}/\Gamma_A)$, i.e. the ratio of quantities that specify what percentage of each object's loss is radiated. Next, we calculate the ratio of energies radiated as function of ($\Gamma_{rad}^{B}/\Gamma_B$)/($\Gamma_{rad}^{A}/\Gamma_A$) and κ/Γ_B in the two special cases Γ_B/Γ_A=50, and Γ_B/Γ_A=500, and plot them in Fig. 22c and Fig. 22d, respectively. We also show, in Fig. 22e, the dependence of $(E_{rad})_{constant\text{-}\kappa}/(E_{rad})_{EIT\text{-}like}$ on κ/Γ_B and Γ_B/Γ_A, for the special case $\Gamma_{rad}^{A}=0$. As can be seen from Fig. 22c-22d, the adiabatic-κ scheme is less radiative than the constant-κ scheme whenever $\Gamma_{rad}^{B}/\Gamma_B$ is larger than $\Gamma_{rad}^{A}/\Gamma_A$, and the radiation reduction ratio increases as Γ_B/Γ_A and κ/Γ_B are increased (see fig. 22e). In some examples, the adiabatic-κ scheme is less radiative than the constant-κ scheme whenever $\Gamma_{rad}^{B}/\Gamma_{rad}^{A}$ is larger than about 20. In some examples, the adiabatic-κ scheme is less radiative than the constant-κ scheme whenever $\Gamma_{rad}^{B}/\Gamma_{rad}^{A}$ is larger than about 50.

It is to be understood that while three resonant objects are shown in the previous examples, other examples can feature four or more resonant objects. For example, in some examples, a single source object can transfer energy to multiple device objects through one intermediate object. In some examples, energy can be transferred from a source resonant object to a device resonant object, through two or more intermediate resonant objects, and so forth.

6. System Sensitivity to Extraneous Objects

In general, the overall performance of an example of the resonance-based wireless energy-transfer scheme depends strongly on the robustness of the resonant objects' resonances. Therefore, it is desirable to analyze the resonant objects' sensitivity to the near presence of random non-resonant extraneous objects. One appropriate analytical model is that of "perturbation theory" (PT), which suggests that in the presence of an extraneous perturbing object p the field amplitude $a_1(t)$ inside the resonant object 1 satisfies, to first order:

$$\frac{da_1}{dt} = -i(\omega_1 - i\Gamma_1)a_1 + i\left(\delta\kappa_{11(p)} + i\delta\Gamma_{1(p)}\right)a_1 \tag{47}$$

where again ω_1 is the frequency and Γ_1 the intrinsic (absorption, radiation etc.) loss rate, while $\delta\kappa_{11(\mathrm{p})}$ is the frequency shift induced onto 1 due to the presence of p and $\delta\Gamma_{1(p)}$ is the extrinsic due to p (absorption inside p, scattering from p etc.) loss rate. $\delta\Gamma_{1(p)}$ is defined as $\delta\Gamma_{1(p)} \equiv \Gamma_{1(p)} - \Gamma_1$, where $\Gamma_{1(p)}$ is the total perturbed loss rate in the presence of p. The first-order PT model is valid only for small perturbations. Nevertheless, the parameters $\delta\kappa_{11(\mathrm{p})}$, $\delta\Gamma_{1(p)}$ are well defined, even outside that regime, if a_1 is taken to be the amplitude of the exact perturbed mode. Note also that interference effects between the radiation field of the initial resonant-object mode and the field scattered off the extraneous object can for strong scattering (e.g. off metallic objects) result in total $\Gamma_{1,\mathrm{rad(p)}}$ that are smaller than the initial $\Gamma_{1,\mathrm{rad}}$ (namely $\delta\Gamma_{1,\mathrm{rad(p)}}$ is negative).

It has been shown that a specific relation is desired between the resonant frequencies of the source and device-objects and the driving frequency. In some examples, all resonant objects must have the same eigenfrequency and this must be equal to the driving frequency. In some implementations, this frequency-shift can be "fixed" by applying to one or more resonant objects and the driving generator a feedback mechanism that corrects their frequencies. In some examples, the driving frequency from the generator can be fixed and only the resonant frequencies of the objects can be tuned with respect to this driving frequency.

The resonant frequency of an object can be tuned by, for example, adjusting the geometric properties of the object (e.g. the height of a self-resonant coil, the capacitor plate spacing of a capacitively-loaded loop or coil, the dimensions of the inductor of an inductively-loaded rod, the shape of a dielectric disc, etc.) or changing the position of a non-resonant object in the vicinity of the resonant object.

In some examples, referring to Fig. 23a, each resonant object is provided with an oscillator at fixed frequency and a monitor which determines the eigenfrequency of the object. At least one of the oscillator and the monitor is coupled to a frequency adjuster which can adjust the frequency of the resonant object. The frequency adjuster determines the difference between the fixed driving frequency and the object frequency and acts, as described above, to bring the object frequency into the required relation with respect to the fixed frequency. This technique assures that all resonant objects operate at the same fixed frequency, even in the presence of extraneous objects.

In some examples, referring to Fig. 23(b), during energy transfer from a source object to a device object, the device object provides energy or power to a load, and an efficiency monitor measures the efficiency of the energy-transfer or power-transmission. A frequency adjuster coupled to the load and the efficiency monitor acts, as described above, to adjust the frequency of the object to maximize the efficiency.

In other examples, the frequency adjusting scheme can rely on information exchange between the resonant objects. For example, the frequency of a source object can be monitored and transmitted to a device object, which is in turn synched to this frequency using frequency adjusters, as described above. In other embodiments the frequency of a single clock can be transmitted to multiple devices, and each device then synched to that frequency using frequency adjusters, as described above.

Unlike the frequency shift, the extrinsic perturbing loss due to the presence of extraneous perturbing objects can be detrimental to the functionality of the energy-transfer scheme, because it is difficult to remedy. Therefore, the total perturbed quality factors $Q_{(p)}$ (and the corresponding perturbed strong-coupling factor $U_{(p)}$ should be quantified.

In some examples, a system for wireless energy-transfer uses primarily magnetic resonances, wherein the energy stored in the near field in the air region surrounding the resonator is predominantly magnetic, while the electric energy is stored primarily inside the resonator. Such resonances can exist in the quasi-static regime of operation ($r \ll \lambda$) that we are considering: for example, for coils with $h \ll 2r$, most of the electric field is localized within the self-capacitance of the coil or the externally loading capacitor and, for dielectric disks, with $\epsilon \gg 1$ the electric field is preferentially localized inside the disk. In some examples, the influence of extraneous objects on magnetic resonances is nearly absent. The reason is that extraneous non-conducting objects p that could interact with the magnetic field in the air region surrounding the resonator and act as a perturbation to the resonance are those having significant magnetic properties (magnetic permeability $Re\{\mu\}>1$ or magnetic loss $Im\{\mu\}>0$). Since almost all every-day non-conducting materials are non-magnetic but just dielectric, they respond to magnetic fields in the same way as free space, and thus will not disturb the resonance of the resonator. Extraneous conducting materials can however lead to some extrinsic losses due to the eddy currents induced inside them or on their surface (depending on their conductivity). However, even for such conducting materials, their presence will not be detrimental to the resonances, as long as they are not in very close proximity to the resonant objects.

The interaction between extraneous objects and resonant objects is reciprocal, namely, if an extraneous object does not influence a resonant object, then also the resonant object does not influence the extraneous object. This fact can be viewed in light of safety considerations for human beings. Humans are also non-magnetic and can sustain strong magnetic fields without undergoing any risk. A typical example, where magnetic fields $B\sim 1T$ are safely used on humans, is the Magnetic Resonance Imaging (MRI) technique for medical testing. In contrast, the magnetic near-field required in typical embodiments in order to provide a few Watts of power to devices is only $B\sim 10^{-4}T$, which is actually comparable to the magnitude of the Earth's magnetic field. Since, as explained above, a strong electric near-field is also not present and the radiation produced from this non-radiative scheme is minimal, the energy-transfer apparatus, methods and systems described herein is believed safe for living organisms.

An advantage of the presently proposed technique using adiabatic variations of the coupling rates between the first and intermediate resonators and between the intermediate and second resonators compared to a mode of operation where these coupling rates are not varied but are constant is that the interactions of the intermediate resonator with extraneous objects can be greatly reduced with the presently proposed scheme. The reason is once more the fact that there is always a substantially small amount of energy in the intermediate resonator in the adiabatic-κ scheme, therefore there is little energy that can be induced from the intermediate object to an extraneous object in its vicinity. Furthermore, since the losses of the intermediate resonator are substantially avoided in the adiabatic-κ case, the system is less immune to potential reductions of the quality factor of the intermediate resonator due to extraneous objects in its vicinity.

6.1 Capacitively-loaded conducting loops or coils

In some examples, one can estimate the degree to which the resonant system of a capacitively-loaded conducting-wire coil has mostly magnetic energy stored in the space surrounding it. If one ignores the fringing electric field from the capacitor, the electric and magnetic energy densities in the space surrounding the coil come just from the electric and magnetic field produced by the current in the wire; note that in the far field, these two energy densities must be equal, as is always the case for radiative fields. By using the results for the fields produced by a subwavelength ($r \ll \lambda$) current loop (magnetic dipole) with $h = 0$, we can calculate the ratio of electric to magnetic energy densities, as a function of distance D_p from the center of the loop (in the limit $r \ll D_p$) and the angle θ with respect to the loop axis:

$$\begin{aligned} \frac{w_e(x)}{w_m(x)} &= \frac{\varepsilon_o |E(x)|^2}{\mu_o |H(x)|^2} = \frac{\left(1+\frac{1}{x^2}\right)\sin^2\theta}{\left(\frac{1}{x^2}+\frac{1}{x^4}\right)4\cos^2\theta + \left(1-\frac{1}{x^2}+\frac{1}{x^4}\right)\sin^2\theta};\ x = 2\pi\frac{D_p}{\lambda} \\ \Rightarrow \frac{\oint\!\!\oint_{S_p} w_e(x)dS}{\oint\!\!\oint_{S_p} w_m(x)dS} &= \frac{1+\frac{1}{x^2}}{1+\frac{1}{x^2}+\frac{3}{x^4}};\ x = 2\pi\frac{D_p}{\lambda} \end{aligned}, \qquad (48)$$

where the second line is the ratio of averages over all angles by integrating the electric and magnetic energy densities over the surface of a sphere of radius D_p. From Eq.(48) it is obvious that indeed for all angles in the near field ($x \ll 1$) the magnetic energy density is dominant, while in the far field ($x \gg 1$) they are equal as they should be. Also, the preferred positioning of the loop is such that objects which can interfere with its resonance lie close to its axis ($\theta = 0$), where there is no electric field. For example, using the systems described in Table 4, we can estimate from Eq.(48) that for the loop of $r = 30cm$ at a distance $D_p = 10r = 3m$ the ratio of average electric to average magnetic energy density would be $\sim 12\%$ and at $D_p = 3r = 90cm$ it would be $\sim 1\%$, and for the loop of $r = 10cm$ at a distance $D_p = 10r = 1m$ the ratio would be $\sim 33\%$ and at $D_p = 3r = 30cm$ it would be $\sim 2.5\%$. At closer distances this ratio is even smaller and thus the energy is predominantly magnetic in the near field, while in the radiative far field, where they are necessarily of the same order (ratio$\rightarrow 1$), both are very small, because the fields have significantly decayed, as capacitively-loaded coil systems are designed to radiate very little. Therefore, this is the criterion that qualifies this class of resonant system as a magnetic resonant system.

To provide an estimate of the effect of extraneous objects on the resonance of a capacitively-loaded loop including the capacitor fringing electric field, we use the perturbation theory formula, stated earlier, $\delta\Gamma_{1,abs(p)} = \omega_1 / 4 \cdot \int d^3\mathbf{r}\,\mathrm{Im}\{\varepsilon_p(\mathbf{r})\}|\mathbf{E}_1(\mathbf{r})|^2 / W$ with the computational FEFD results for the field of an example like the one shown in the plot of Fig. 5 and with a rectangular object of dimensions *30cm x 30cm x 1.5m* and permittivity *ε=49+16i* (consistent with human muscles) residing between the loops and almost standing on top of one capacitor (*~3cm* away from it) and find $\delta Q_{abs(human)} \sim 10^5$ and for *~10cm* away $\delta Q_{abs(human)} \sim 5 \cdot 10^5$. Thus, for ordinary distances (*~1m*) and placements (not immediately on top of the capacitor) or for most ordinary extraneous objects p of much smaller loss-tangent, we conclude that it is indeed fair to say that $\delta Q_{abs(p)} \rightarrow \infty$. The only perturbation that is expected to affect these resonances is a close proximity of large metallic structures.

Self-resonant coils can be more sensitive than capacitively-loaded coils, since for the former the electric field extends over a much larger region in space (the entire coil) rather than for the latter (just inside the capacitor). On the other hand, self-resonant coils can be simple to make and can withstand much larger voltages than most lumped capacitors. Inductively-loaded conducting rods can also be more sensitive than capacitively-loaded coils, since they rely on the electric field to achieve the coupling.

6.2 Dielectric disks

For dielectric disks, small, low-index, low-material-loss or far-away stray objects will induce small scattering and absorption. In such cases of small perturbations these extrinsic loss mechanisms can be quantified using respectively the analytical first-order perturbation theory formulas

$$\left[\delta Q_{1,rad(p)}\right]^{-1} \equiv 2\delta\Gamma_{1,rad(p)} / \omega_1 \propto \int d^3\mathbf{r}\left[\mathrm{Re}\left\{\varepsilon_p(\mathbf{r})\right\}\left|\mathbf{E}_1(\mathbf{r})\right|\right]^2 \Big/ W$$

$$\left[\delta Q_{1,abs(p)}\right]^{-1} \equiv 2\delta\Gamma_{1,abs(p)} / \omega_1 = \int d^3\mathbf{r}\,\mathrm{Im}\left\{\varepsilon_p(\mathbf{r})\right\}\left|\mathbf{E}_1(\mathbf{r})\right|^2 \Big/ 2W$$

where $W = \int d^3\mathbf{r}\varepsilon(\mathbf{r})\left|\mathbf{E}_1(\mathbf{r})\right|^2 / 2$ is the total resonant electromagnetic energy of the unperturbed mode. As one can see, both of these losses depend on the square of the resonant electric field tails $\mathbf{E}1$ at the site of the extraneous object. In contrast, the coupling factor from object 1 to another resonant object 2 is, as stated earlier,

$$k_{12} = 2\kappa_{12} / \sqrt{\omega_1\omega_2} \approx \int d^3\mathrm{r}\varepsilon_2(\mathrm{r})\mathrm{E}_2^*(\mathrm{r})\mathrm{E}_1(\mathrm{r}) / \int d^3\mathrm{r}\varepsilon(\mathrm{r})\left|\mathrm{E}_1(\mathrm{r})\right|^2$$

and depends *linearly* on the field tails $\mathbf{E}_1$ of 1 inside 2. This difference in scaling gives us confidence that, for, for example, exponentially small field tails, coupling to other resonant objects should be much faster than all extrinsic loss rates ($\kappa_{12} \gg \delta\Gamma_{1,2(p)}$), at least for small perturbations, and thus the energy-transfer scheme is expected to be sturdy for this class of resonant dielectric disks.

However, we also want to examine certain possible situations where extraneous objects cause perturbations too strong to analyze using the above first-order perturbation theory approach. For example, we place a dielectric disk close to another off-resonance object of large $\mathrm{Re}\{\varepsilon\}$, $\mathrm{Im}\{\varepsilon\}$ and of same size but different shape (such as a human being *h*), as shown in Fig. 24a, and a roughened surface of large extent but of small $\mathrm{Re}\{\varepsilon\}$, $\mathrm{Im}\{\varepsilon\}$ (such as a wall *w*), as shown in Fig. 24b. For distances $D_{h,w} / r = 10-3$ between the disk-center and the "human"-center or "wall", the numerical FDFD simulation results presented in Figs. 24a and 24b suggest that, the disk resonance seems to be fairly robust, since it is not detrimentally disturbed by the presence of extraneous objects, with the exception of the *very* close proximity of high-loss objects. To examine the influence of large perturbations on an entire energy-transfer system we consider two resonant disks in the close presence of both a "human" and a "wall". Comparing Table 8 to the table in Figure 24c, the numerical FDFD simulations show that the system performance deteriorates from $U \sim 1-50$ to $U_{(hw)} \sim 0.5-10$, i.e. only by acceptably small amounts.

In general, different examples of resonant systems have different degree of sensitivity to external perturbations, and the resonant system of choice depends on the particular application at hand, and how important matters of sensitivity or safety are for that application. For example, for a medical implantable device (such as a wirelessly powered artificial heart) the electric field extent must be minimized to the highest degree possible to protect the tissue surrounding the device. In such cases where sensitivity to external objects or safety is important, one should design the resonant systems so that the ratio of electric to magnetic energy density w_e / w_m is reduced or minimized at most of the desired (according to the application) points in the surrounding space.

7 Applications

The non-radiative wireless energy transfer techniques described above can enable efficient wireless energy-exchange between resonant objects, while suffering only modest transfer and dissipation of energy into other extraneous off-resonant objects. The technique is general, and can be applied to a variety of resonant systems in nature. In this Section, we identify a variety of applications that can benefit from or be designed to utilize wireless power transmission.

Remote devices can be powered directly, using the wirelessly supplied power or energy to operate or run the devices, or the devices can be powered by or through or in addition to a battery or energy storage unit, where the battery is occasionally being charged or re-charged wirelessly. The devices can be powered by hybrid battery/energy storage devices such as batteries with integrated storage capacitors and the like. Furthermore, novel battery and energy storage devices can be designed to take advantage of the operational improvements enabled by wireless power transmission systems.

Devices can be turned off and the wirelessly supplied power or energy used to charge or recharge a battery or energy storage unit. The battery or energy storage unit charging or recharging rate can be high or low. The battery or energy storage units can be trickle charged or float charged. It would be understood by one of ordinary skill in the art that there are a variety of ways to power and/or charge devices, and the variety of ways could be applied to the list of applications that follows.

Some wireless energy transfer examples that can have a variety of possible applications include for example, placing a source (e.g. one connected to the wired electricity network) on the ceiling of a room, while devices such as robots, vehicles, computers, PDAs or similar are placed or move freely within the room. Other applications can include powering or recharging electric-engine buses and/or hybrid cars and medical implantable devices. Additional example applications include the ability to power or recharge autonomous electronics (e.g. laptops, cell-phones, portable music players, house-hold robots, GPS navigation systems, displays, etc), sensors, industrial and manufacturing equipment, medical devices and monitors, home appliances (e.g. lights, fans, heaters, displays, televisions, counter-top appliances, etc.), military devices, heated or illuminated clothing, communications and navigation equipment, including equipment built into vehicles, clothing and protective-wear such as helmets, body armor and vests, and the like, and the ability to transmit power to physically isolated devices such as to implanted medical devices, to hidden, buried, implanted or embedded sensors or tags, to and/or from roof-top solar panels to indoor distribution panels, and the like.

A number of examples of the invention have been described. Nevertheless, it will be understood that various modifications can be made without departing from the spirit and scope of the invention.

WHAT IS CLAIMED IS:

1. A method for transferring energy wirelessly, the method comprising:
transferring energy wirelessly from a first resonator structure to an intermediate resonator structure, wherein the coupling rate between the first resonator structure and the intermediate resonator structure is κ_{1B};
transferring energy wirelessly from the intermediate resonator structure to a second resonator structure, wherein the coupling rate between the intermediate resonator structure and the second resonator structure is κ_{B2}; and
during the wireless energy transfers, adjusting at least one of the coupling rates κ_{1B} and κ_{B2} to reduce energy accumulation in the intermediate resonator structure and improve wireless energy transfer from the first resonator structure to the second resonator structure through the intermediate resonator structure.

2. The method of claim 1, wherein the adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} minimizes energy accumulation in the intermediate resonator structure and causes wireless energy transfer from the first resonator structure to the second resonator structure.

3. The method of claims 1 or 2, wherein the adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} maintains energy distribution in the field of the three-resonator system in an eigenstate having substantially no energy in the intermediate resonator structure.

4. The method of claim 3, wherein the adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} further causes the eigenstate to evolve substantially adiabatically from an initial state with substantially all energy in the resonator structures in the first resonator structure to a final state with substantially all of the energy in the resonator structures in the second resonator structure.

5. The method of any of claims 1 to 4, wherein the adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} comprises adjustments of both coupling rates κ_{1B} and κ_{B2} during wireless energy transfer.

6. The method of claim 1, wherein the resonator structures each have a quality factor larger than 10.

7. The method of any of the preceding claims, wherein resonant energy in each of the resonator structures comprises electromagnetic fields.

8. The method of claim 7, wherein the maximum value of the coupling rate κ_{1B} and the maximum value of the coupling rate κ_{B2} for inductive coupling between the intermediate resonator structure and each of the first and second resonator structures are each larger than twice the loss rate Γ for each of the first and second resonators.

9. The method of claim 8, wherein the maximum value of the coupling rate κ_{1B} and the maximum value of the coupling rate κ_{B2} for inductive coupling between the intermediate resonator structure and each of the first and second resonator structures are each larger than four (4) times the loss rate Γ for each of the first and second resonators.

10. The method of claim 7, wherein each resonator structure has a resonant frequency between 50 KHz and 500 MHz.

11. The method of any of the preceding claims, wherein the maximum value of the coupling rate κ_{1B} and the maximum value of the coupling rate κ_{B2} are each at least five (5) times greater than the coupling rate between the first resonator structure and the second resonator structure.

12. The method of any of the preceding claims, wherein the intermediate resonator structure has a rate of radiative energy loss that is at least twenty (20) times greater than that for either the first resonator structure or the second resonator structure.

13. The method of claim 1, wherein the first and second resonator structures are substantially identical.

14. The method of any of the preceding claims, wherein the adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} causes peak energy accumulation in the intermediate resonator structure to be less than five percent (5%) of the peak total energy in the three resonator structures.

15. The method of any of the preceding claims, wherein adjusting at least one of the coupling rates κ_{1B} and κ_{B2} comprises adjusting a relative position and/or orientation between one or more pairs of the resonator structures.

16. The method of any of the preceding claims, wherein adjusting at least one of the coupling rates κ_{1B} and κ_{B2} comprises adjusting a resonator property of one or more of the resonator structures.

17. The method of claim 16, wherein the resonator property comprises mutual inductance.

18. The method of any of the preceding claims, wherein at least one of the resonator structures comprises a capacitively loaded loop or coil of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

19. The method of any of the preceding claims, wherein at least one of the resonator structures comprises an inductively loaded rod of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

20. An apparatus comprising:
first, intermediate, and second resonator structures, wherein a coupling rate between the first resonator structure and the intermediate resonator structure is κ_{1B} and a coupling rate between the intermediate resonator structure and the second resonator structure is κ_{B2}; and
means for adjusting at least one of the coupling rates κ_{1B} and κ_{B2} during wireless energy transfers among the resonator structures to reduce energy accumulation in the intermediate resonator structure and improve wireless energy transfer from the first resonator structure to the second resonator structure through the intermediate resonator structure.

21. The apparatus of claim 20, wherein the means for adjusting comprises a rotation stage for adjusting the relative orientation of the intermediate resonator structure with respect to the first and second resonator structures.

22. The apparatus of claim 20, wherein the means for adjusting comprises a translation stage for moving the first and/or second resonator structures relative to the intermediate resonator structure.

23. The apparatus of claim 20, wherein the means for adjusting comprises a mechanical, electro-mechanical, or electrical staging system for dynamically adjusting the effective size of one or more of the resonator structures.

24. The method of claim 4, wherein the adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} causes peak energy accumulation in the intermediate resonator

structure during the wireless energy transfers to be less than ten percent (10%) of the peak total energy in the three resonator structures.

25. The method of any of claims 1-19 and 24, wherein the wireless energy transfers are non-radiative energy transfers mediated by a coupling of a resonant field evanescent tail of the first resonator structure and a resonant field evanescent tail of the intermediate resonator structure and a coupling of the resonant field evanescent tail of the intermediate resonator structure and a resonant field evanescent tail of the second resonator structure.

26. The method of claim 25, wherein the first and second resonator structures each have a quality factor greater than 50.

27. The method of claim 25, wherein the first and second resonator structures each have a quality factor greater than 100.

28. A method for transferring energy wirelessly, the method comprising:
transferring energy wirelessly from a first resonator structure to a intermediate resonator structure, wherein the coupling rate between the first resonator structure and the intermediate resonator structure is κ_{1B};
transferring energy wirelessly from the intermediate resonator structure to a second resonator, wherein the coupling rate between the intermediate resonator structure and the second resonator structure is κ_{B2}; and
during the wireless energy transfers, adjusting at least one of the coupling rates κ_{1B} and κ_{B2} to cause an energy distribution in the field of the three-resonator system to have substantially no energy in the intermediate resonator structure while wirelessly transferring energy from the first resonator structure to the second resonator structure through the intermediate resonator structure.

29. The method of claim 28, wherein having substantially no energy in the intermediate resonator structure means that peak energy accumulation in the intermediate resonator structure is less than ten percent (10%) of the peak total energy in the three resonator structures throughout the wireless energy transfer.

30. The method of claim 28, wherein having substantially no energy in the intermediate resonator structure means that peak energy accumulation in the intermediate resonator structure is less than five percent (5%) of the peak total energy in the three resonator structures throughout the wireless energy transfer.

31. The method of any of claims 28 to 30, wherein the adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} maintains the energy distribution in the field of the three-resonator system in an eigenstate having the substantially no energy in the intermediate resonator structure.

32. The method of claim 31, wherein the adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} further causes the eigenstate to evolve substantially adiabatically from an initial state with substantially all energy in the resonator structures in the first resonator structure to a final state with substantially all of the energy in the resonator structures in the second resonator structure.

33. The method of any of claims 28 to 32, wherein the adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} comprises adjustments of both coupling rates κ_{1B} and κ_{B2} during wireless energy transfers.

34. The method of any of claims 28 to 33, wherein resonant energy in each of the resonator structures comprises electromagnetic fields.

35. The method of claim 34, wherein the maximum value of the coupling rate κ_{1B} and the maximum value of the coupling rate κ_{B2} for inductive coupling between the intermediate resonator structure and each of the first and second resonator structures are each larger than twice the loss rate Γ for each of the first and second resonators.

36. The method of claim 34, wherein the maximum value of the coupling rate κ_{1B} and the maximum value of the coupling rate κ_{B2} for inductive coupling between the intermediate resonator structure and each of the first and second resonator structures are each larger than four (4) times the loss rate Γ for each of the first and second resonators.

37. The method of any of claims 34 to 36, wherein each resonator structure has a resonant frequency between 50 KHz and 500 MHz.

38. The method of any of claims 28 to 37, wherein the maximum value of the coupling rate κ_{1B} and the maximum value of the coupling rate κ_{B2} are each at least five (5) times greater than the coupling rate between the first resonator structure and the second resonator structure.

39. The method of any of claims 28 to 38, wherein the intermediate resonator structure has a rate of radiative energy loss that is at least twenty (20) times greater than that for either the first resonator structure or the second resonator structure.

40. The method of any of claims 28 to 39, wherein the first and second resonator structures are substantially identical.

41. The method of any of claims 28 to 40, wherein adjusting at least one of the coupling rates κ_{1B} and κ_{B2} comprises adjusting a relative position and/or orientation between one or more pairs of the resonator structures.

42. The method of any of claims 28-41, wherein adjusting at least one of the coupling rates κ_{1B} and κ_{B2} comprises adjusting a resonator property of one or more of the resonator structures.

43. The method of claim 42, wherein the resonator property comprises mutual inductance.

44. The method of any of claims 28 to 43, wherein at least one of the resonator structures comprises a capacitively loaded loop or coil of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

45. The method of any of claims 28 to 44, wherein at least one of the resonator structures comprises an inductively loaded rod of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

46. The method of any of claims 28 to 45, wherein the wireless energy transfers are non-radiative energy transfers mediated by a coupling of a resonant field evanescent tail of the first resonator structure and a resonant field evanescent tail of the intermediate resonator structure and a coupling of the resonant field evanescent tail of the intermediate resonator structure and a resonant field evanescent tail of the second resonator structure.

47. The method of claim 46, wherein the first and second resonator structures each have a quality factor greater than 50.

48. The method of claim 47, wherein the first and second resonator structures each have a quality factor greater than 100.

49. The method of any of claims 28 to 48, wherein the adjustment of at least one of the coupling rates κ_{1B} and κ_{B2} to cause the energy distribution in the field of the three-

resonator system to have substantially no energy in the intermediate resonator structure improves wireless energy transfer between the first and second resonator structures.

50. An apparatus comprising:
first, intermediate, and second resonator structures, wherein a coupling rate between the first resonator structure and the intermediate resonator structure is κ_{1B} and a coupling rate between the intermediate resonator structure and the second resonator structure is κ_{B2}; and
means for adjusting at least one of the coupling rates κ_{1B} and κ_{B2} during wireless energy transfers among the resonator structures to cause an energy distribution in the field of the three-resonator system to have substantially no energy in the intermediate resonator structure while wirelessly transferring energy from the first resonator structure to the second resonator structure through the intermediate resonator structure.

51. The apparatus of claim 50, wherein having substantially no energy in the intermediate resonator structure means that peak energy accumulation in the intermediate resonator structure is less than ten percent (10%) of the peak total energy in the three resonator structures throughout the wireless energy transfers.

52. The apparatus of claim 50, wherein having substantially no energy in the intermediate resonator structure means that peak energy accumulation in the intermediate resonator structure is less than five percent (5%) of the peak total energy in the three resonator structures throughout the wireless energy transfers.

53. The apparatus of any of claims 50 to 52, wherein the means for adjusting is configured to maintain the energy distribution in the field of the three-resonator system in an eigenstate having the substantially no energy in the intermediate resonator structure.

54. The apparatus of any of claims claim 50 to 53, wherein the means for adjusting comprises a rotation stage for adjusting the relative orientation of the intermediate resonator structure with respect to the first and second resonator structures.

55. The apparatus of any of claims 50 to 54, wherein the means for adjusting comprises a translation stage for moving the first and/or second resonator structures relative to the intermediate resonator structure.

56. The apparatus of any of claims 50 to 55, wherein the means for adjusting comprises a mechanical, electro-mechanical, or electrical staging system for dynamically adjusting the effective size of one or more of the resonator structures.

57. The apparatus of any of claims 50 to 56, wherein at least one of the resonator structures comprises a capacitively loaded loop or coil of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

58. The apparatus of any of claims 50 to 57, wherein at least one of the resonator structures comprises an inductively loaded rod of at least one of a conducting wire, a conducting Litz wire, and a conducting ribbon.

59. The apparatus of any of claims 50 to 58, further comprising a source coupled to the first resonator structure and a load coupled to the second resonator structure.

60. The method of any of claims 1-19 and 24-27, wherein the adjustment of the at least one of the coupling rates defines a first mode of operation,
wherein the reduction in the energy accumulation in the intermediate resonator structure is relative to energy accumulation in the intermediate resonator structure for a second mode of operation of wireless energy transfer among the three resonator structures having a coupling rate κ'_{1B} for wireless energy transfer from the first resonator structure to the intermediate resonator structure and a coupling rate κ'_{B2} for wireless energy transfer

from the intermediate resonator structure to the second resonator structure with κ'_{1B} and κ'_{B2} each being substantially constant during the second mode of wireless energy transfer, and

wherein the adjustment of the coupling rates κ_{1B} and κ_{2B} in the first mode of operation satisfies $\kappa_{1B}, \kappa_{B2} < \sqrt{\left(\kappa'^{2}_{1B} + \kappa'^{2}_{B2}\right)/2}$.

61. The method of claim 60, wherein the first mode of operation has a greater efficiency of energy transferred from the first resonator to the second resonator compared to that for the second mode of operation.

62. The method of claim 61, wherein the first and second resonator structures are substantially identical and each one has a loss rate Γ_A, the intermediate resonator structure has a loss rate Γ_B, and wherein Γ_B/Γ_A is greater than 50.

63. The method of claim 60, wherein a ratio of energy lost to radiation and total energy wirelessly transferred between the first and second resonator structures in the first mode of operation is less than that for the second mode of operation.

64. The method of claim 63, wherein the first and second resonator structures are substantially identical and each one has a loss rate Γ_A and a loss rate only due to radiation $\Gamma_{A,rad}$, the intermediate resonator structure has a loss rate Γ_B and a loss rate only due to radiation $\Gamma_{B,rad}$ and wherein $\Gamma_{B,rad}/\Gamma_B > \Gamma_{A,rad}/\Gamma_A$.

65. The method of claim 60, wherein in the first mode of operation the intermediate resonator structure interacts less with extraneous objects than it does in the second mode of operation.

66. The method of any of claims 1-19, 24-27, and 60-65, wherein during the wireless energy transfer from the first resonator structure to the second resonator structure at least one of the coupling rates is adjusted so that $\kappa_{1B} << \kappa_{B2}$ at a start of the energy transfer and $\kappa_{1B} >> \kappa_{B2}$ by a time a substantial portion of the energy has been transferred from the first resonator structure to the second resonator structure.

67. The method of claim 66, wherein the coupling rate κ_{B2} is maintained at a fixed value and the coupling rate κ_{1B} is increased during the wireless energy transfer from the first resonator structure to second resonator structure.

68. The method of claim 66, wherein the coupling rate κ_{1B} is maintained at a fixed value and the coupling rate κ_{B2} is decreased during the wireless energy transfer from the first resonator structure to second resonator structure.

69. The method of claim 66, wherein, during the wireless energy transfer from the first resonator structure to second resonator structure, the coupling rate κ_{1B} is increased and the coupling rate κ_{B2} is decreased.

70. The method of any of claims 28-49, wherein the adjustment of the at least one of the coupling rates defines a first mode of operation,

wherein energy accumulation in the intermediate resonator structure during the wireless energy transfer from the first resonator structure to second resonator structure is smaller than that for a second mode of operation of wireless energy transfer among the three resonator structures having a coupling rate κ'_{1B} for wireless energy transfer from the first resonator structure to the intermediate resonator structure and a coupling rate κ'_{B2} for wireless energy transfer from the intermediate resonator structure to the second resonator structure with κ'_{1B} and κ'_{B2} each being substantially constant during the second mode of wireless energy transfer, and wherein the adjustment of the coupling rates κ_{1B} and κ_{B2} in the first mode of operation satisfies $\kappa_{1B}, \kappa_{B2} < \sqrt{\left(\kappa'^{2}_{1B} + \kappa'^{2}_{B2}\right)/2}$.

71. The method of claim 70, wherein the first mode of operation has a greater efficiency of energy transferred from the first resonator to the second resonator compared to that for the second mode of operation.

72. The method of claim 71, wherein the first and second resonator structures are substantially identical and each one has a loss rate Γ_A, the intermediate resonator structure has a loss rate Γ_B, and wherein Γ_B/Γ_A is greater than 50.

73. The method of claim 70, wherein a ratio of energy lost to radiation and total energy wirelessly transferred between the first and second resonator structures in the first mode of operation is less than that for the second mode of operation.

74. The method of claim 73, wherein the first and second resonator structures are substantially identical and each one has a loss rate Γ_A and a loss rate only due to radiation $\Gamma_{A,rad}$, the intermediate resonator structure has a loss rate Γ_B and a loss rate only due to radiation $\Gamma_{B,rad}$ and wherein $\Gamma_{B,rad}/\Gamma_B > \Gamma_{A,rad}/\Gamma_A$.

75. The method of claim 70, wherein in the first mode of operation the intermediate resonator structure interacts less with extraneous objects than it does in the second mode of operation.

76. The method of any of claims 28-49 and 70-75, wherein during the wireless energy transfer from the first resonator structure to the second resonator structure at least one of the coupling rates is adjusted so that $\kappa_{1B} << \kappa_{B2}$ at a start of the energy transfer and

$\kappa_{1B} >> \kappa_{B2}$ by a time a substantial portion of the energy has been transferred from the first resonator structure to the second resonator structure.

77. The method of claim 76, wherein the coupling rate κ_{B2} is maintained at a fixed value and the coupling rate κ_{1B} is increased during the wireless energy transfer from the first resonator structure to second resonator structure.

78. The method of claim 76, wherein the coupling rate κ_{1B} is maintained at a fixed value and the coupling rate κ_{B2} is decreased during the wireless energy transfer from the first resonator structure to second resonator structure.

79. The method of claim 76, wherein, during the wireless energy transfer from the first resonator structure to second resonator structure, the coupling rate κ_{1B} is increased and the coupling rate κ_{B2} is decreased.

80. A method for transferring energy wirelessly, the method comprising:
transferring energy wirelessly from a first resonator structure to a intermediate resonator structure, wherein the coupling rate between the first resonator structure and the intermediate resonator structure is κ_{1B};
transferring energy wirelessly from the intermediate resonator structure to a second resonator, wherein the coupling rate between the intermediate resonator structure and the second resonator structure with a coupling rate is κ_{B2}; and
during the wireless energy transfers, adjusting at least one of the coupling rates κ_{1B} and κ_{B2} to define a first mode of operation in which energy accumulation in the intermediate resonator structure is reduced relative to that for a second mode of operation of wireless energy transfer among the three resonator structures having a coupling rate κ'_{1B} for wireless energy transfer from the first resonator structure to the intermediate resonator

structure and a coupling rate κ'_{B2} for wireless energy transfer from the intermediate resonator structure to the second resonator structure with κ'_{1B} and κ'_{B2} each being substantially constant during the second mode of wireless energy transfer, and
wherein the adjustment of the coupling rates κ_{1B} and κ_{B2} in the first mode of operation satisfies $\kappa_{1B},\kappa_{B2}<\sqrt{\left(\kappa'^{2}_{1B}+\kappa'^{2}_{B2}\right)/2}$.

81. An apparatus comprising:
first, intermediate, and second resonator structures, wherein a coupling rate between the first resonator structure and the intermediate resonator structure is κ_{1B} and a coupling rate between the intermediate resonator structure and the second resonator structure is κ_{B2}; and
means for adjusting at least one of the coupling rates κ_{1B} and κ_{B2} during wireless energy transfers among the resonator structures to define a first mode of operation in which energy accumulation in the intermediate resonator structure is reduced relative to that for a second mode of operation for wireless energy transfer among the three resonator structures having a coupling rate κ'_{1B} for wireless energy transfer from the first resonator structure to the intermediate resonator structure and a coupling rate κ'_{B2} for wireless energy transfer from the intermediate resonator structure to the second resonator structure with κ'_{1B} and κ'_{B2} each being substantially constant during the second mode of wireless energy transfer, and
wherein the adjustment of the coupling rates κ_{12} and κ_{B2} in the first mode of operation satisfies $\kappa_{1B},\kappa_{B2}<\sqrt{\left(\kappa'^{2}_{1B}+\kappa'^{2}_{B2}\right)/2}$.

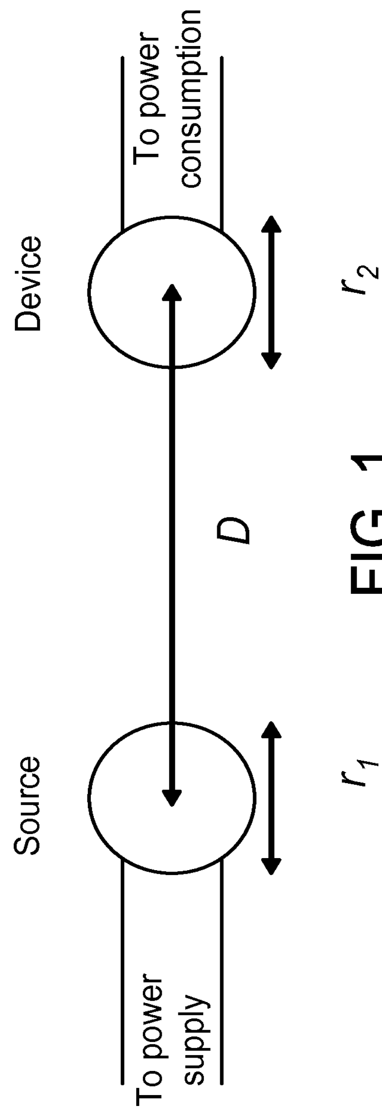


FIG. 1

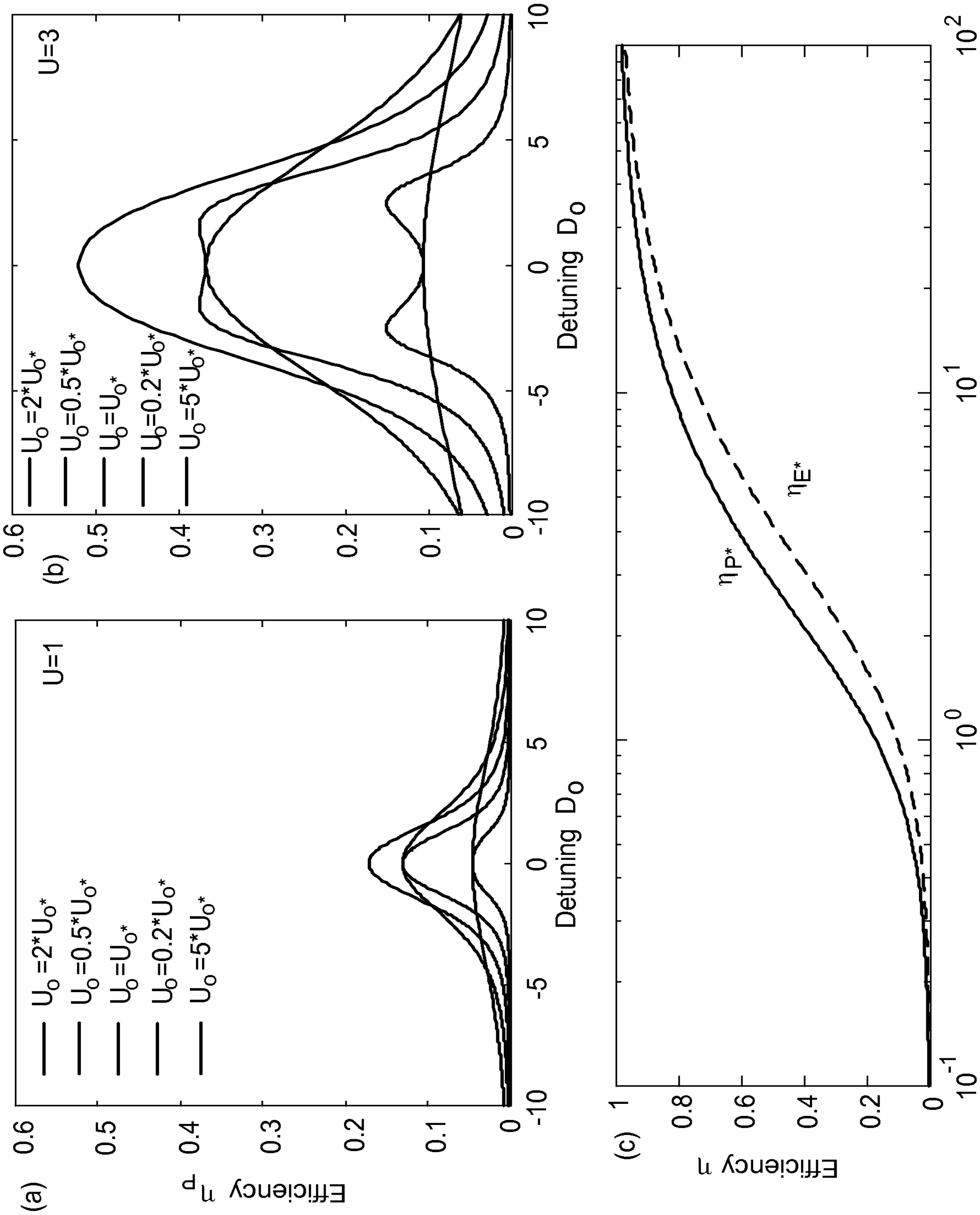

(a)
U=1
$U_o=2*U_o*$
$U_o=0.5*U_o*$
$U_o=U_o*$
$U_o=0.2*U_o*$
$U_o=5*U_o*$
Efficiency η_P
Detuning D_o
(b)
U=3
(c)
Efficiency η
η_{P*}
η_{E*}
strong-coupling factor U=kQ

FIG. 2

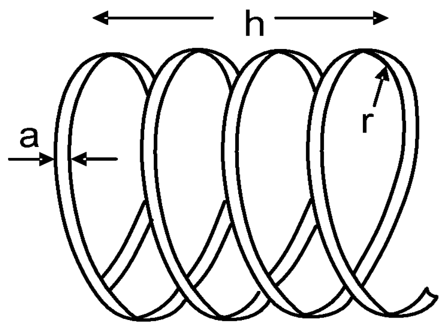


FIG. 3

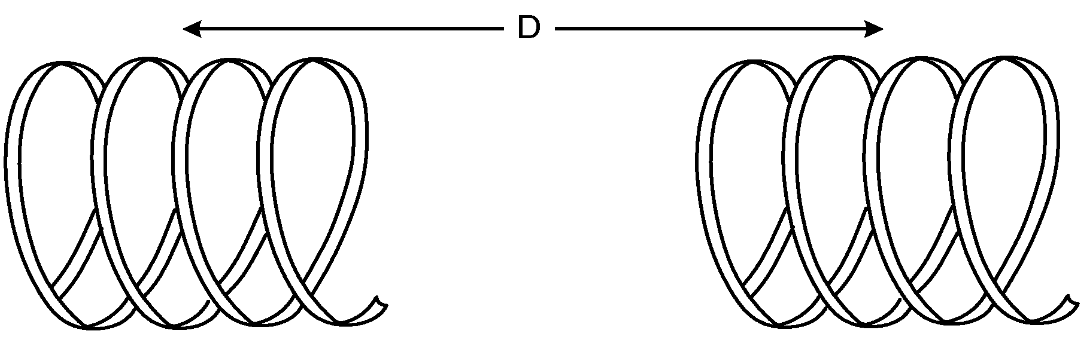


FIG. 4

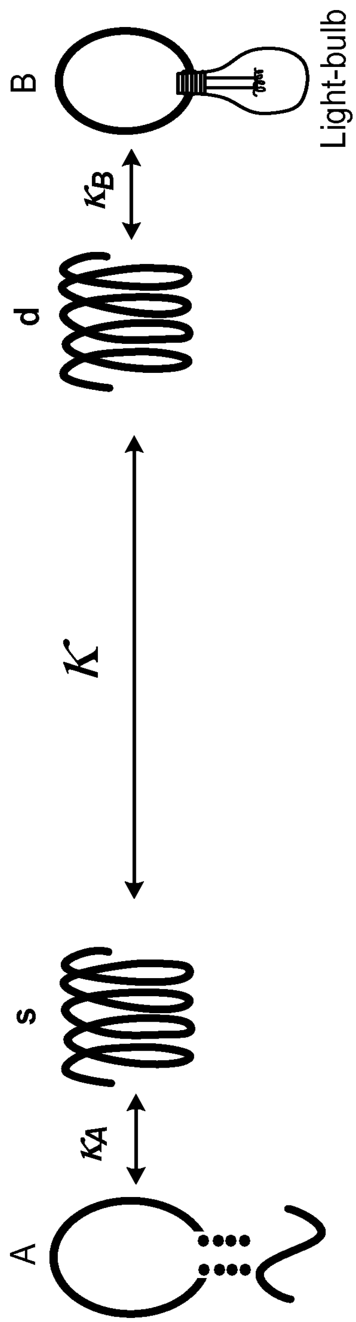
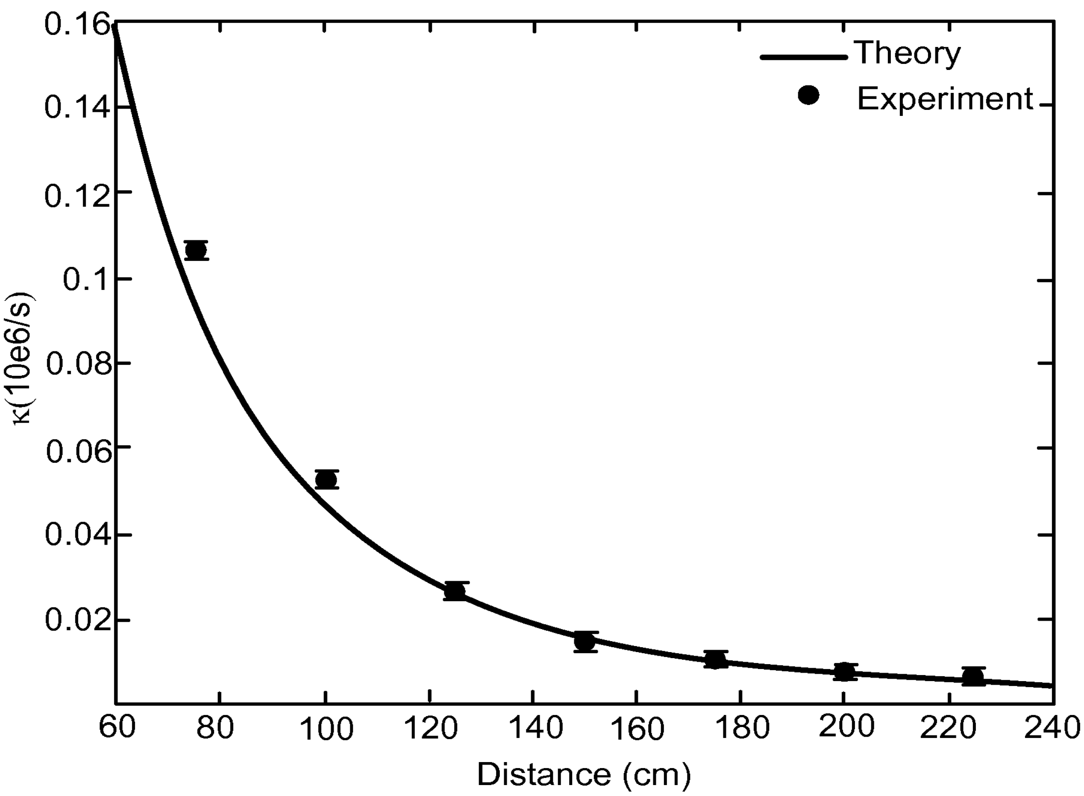
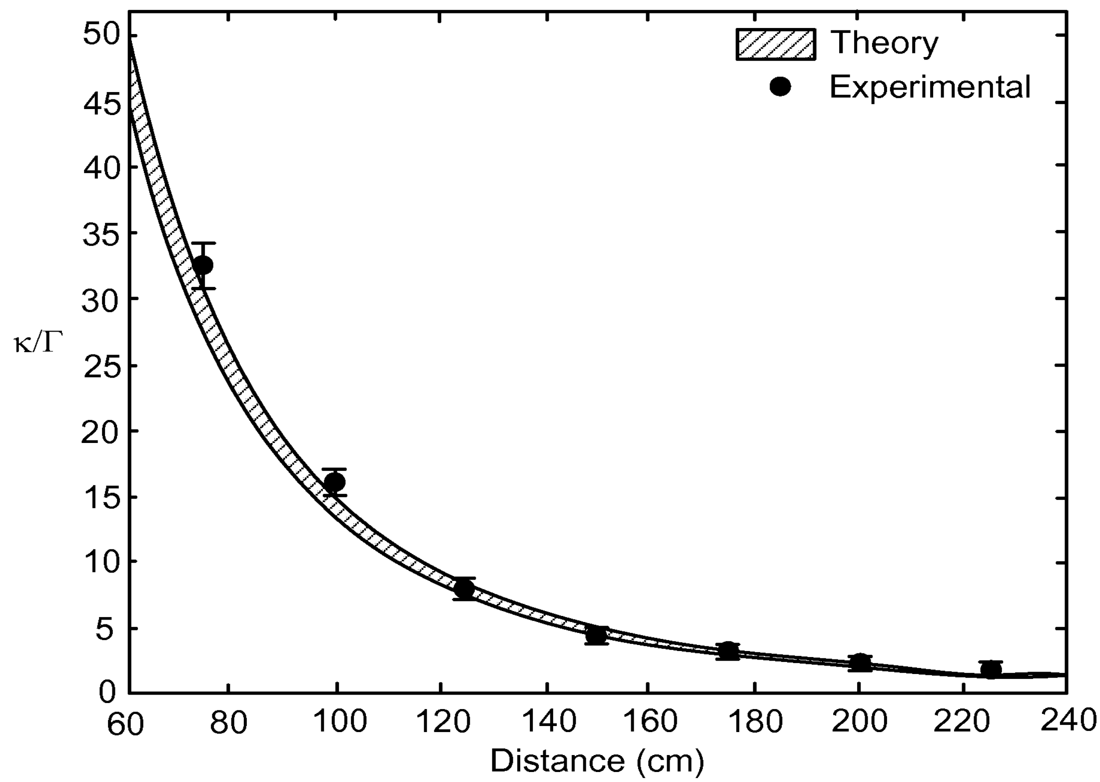


FIG. 5



Comparison of experimental and theoretical values for κ as a function of the separation between the source and device coils.

FIG. 6



Comparison of experimental and theoretical values for the parameter κ/Γ as a function of the separation between the two coils. The theory values are obtained by using the theoretically obtained κ and the experimentally measured Γ. The shaded area represents the spread in the theoretical κ/Γ due to the ~5% uncertainty in Q.

FIG.7

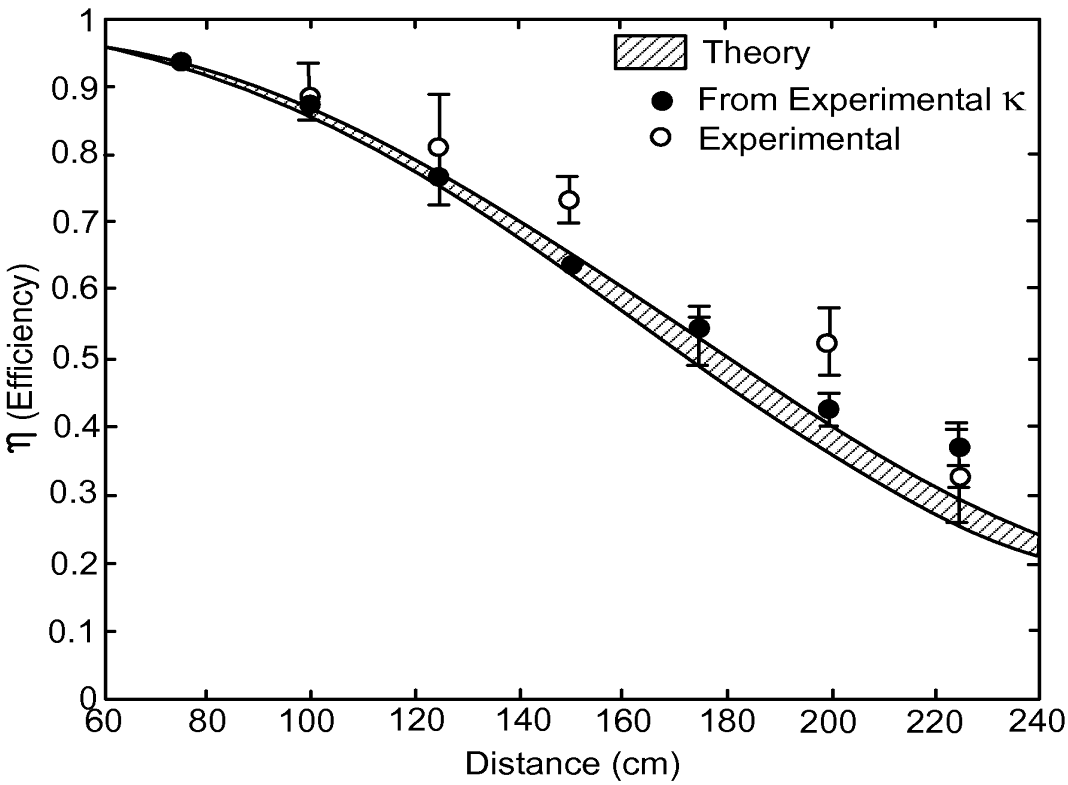

Theory
From Experimental κ
Experimental
η (Efficiency)
Distance (cm)
1
0.9
0.8
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0
60
80
100
120
140
160
180
200
220
240

FIG. 8

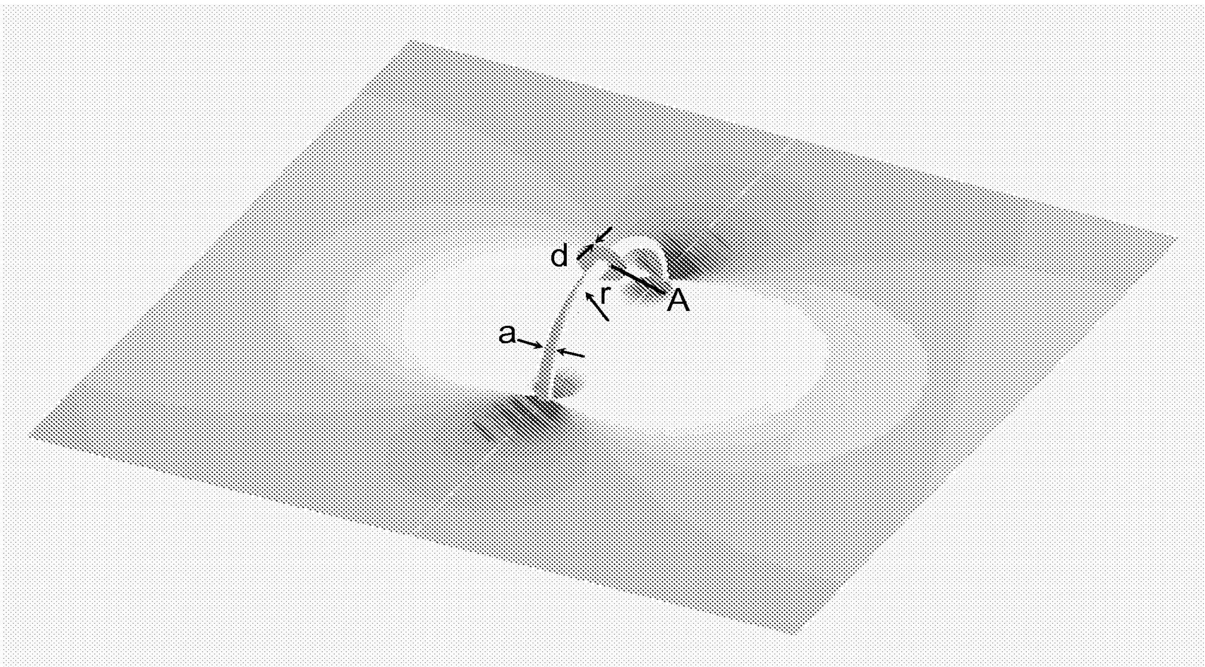


FIG. 9

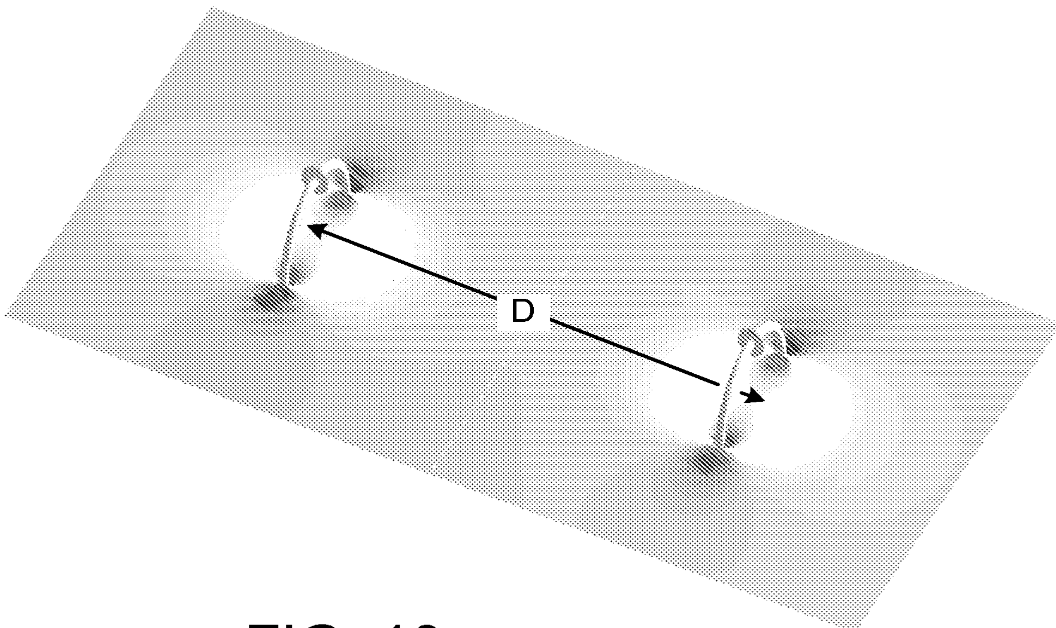


FIG. 10

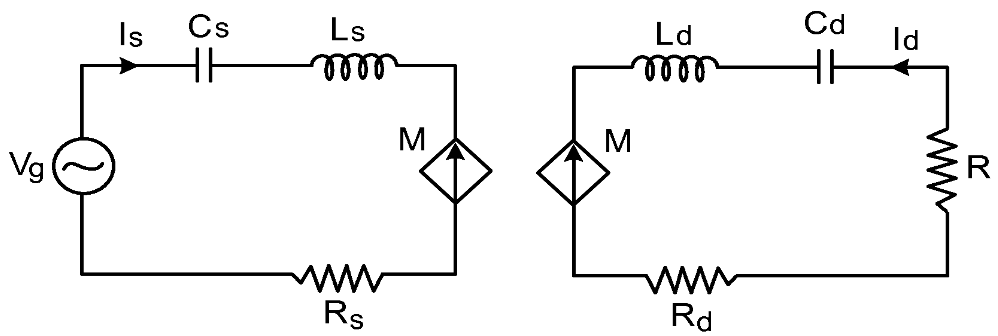


FIG. 11

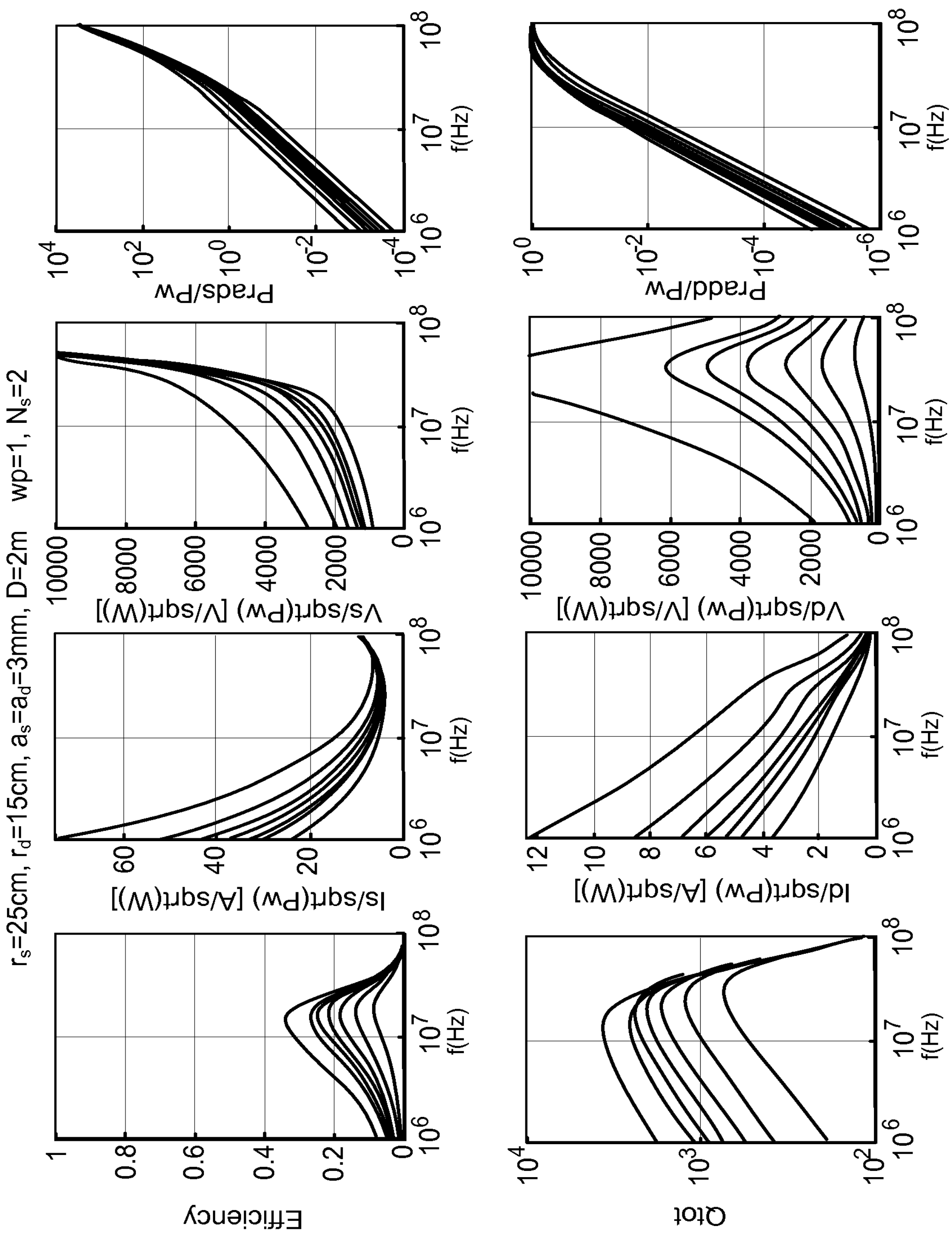

r_s=25cm, r_d=15cm, a_s=a_d=3mm, D=2m wp=1, N_s=2
Efficiency
Is/sqrt(Pw) [A/sqrt(W)]
Vs/sqrt(Pw) [V/sqrt(W)]
Prads/Pw
Qtot
Id/sqrt(Pw) [A/sqrt(W)]
Vd/sqrt(Pw) [V/sqrt(W)]
Pradd/Pw
f(Hz)

FIG. 12

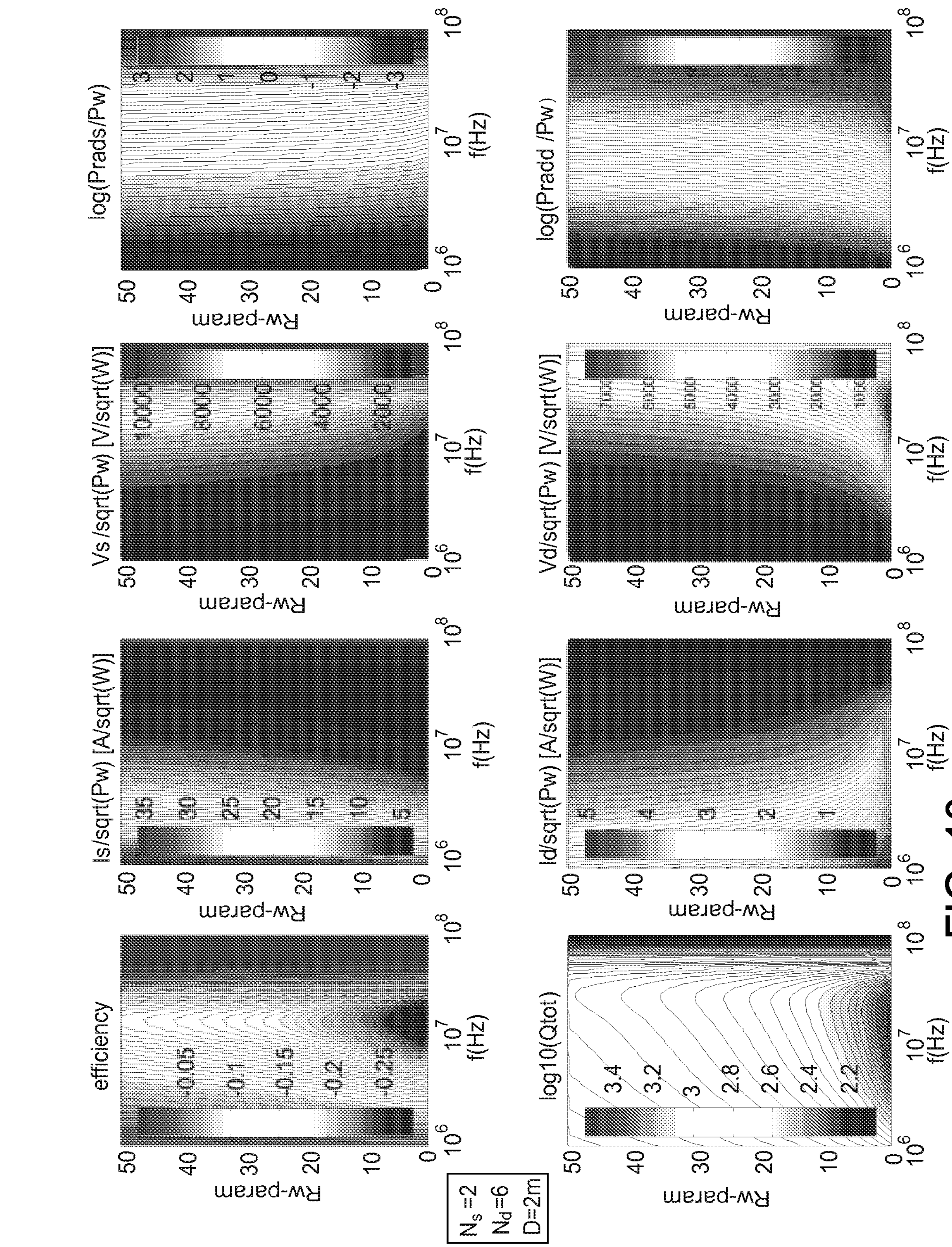

efficiency
Is/sqrt(Pw) [A/sqrt(W)]
Vs /sqrt(Pw) [V/sqrt(W)]
log(Prads/Pw)
log10(Qtot)
Id/sqrt(Pw) [A/sqrt(W)]
Vd/sqrt(Pw) [V/sqrt(W)]
log(Pradd /Pw)
Rw-param
f(Hz)
N_s=2
N_d=6
D=2m

FIG. 13

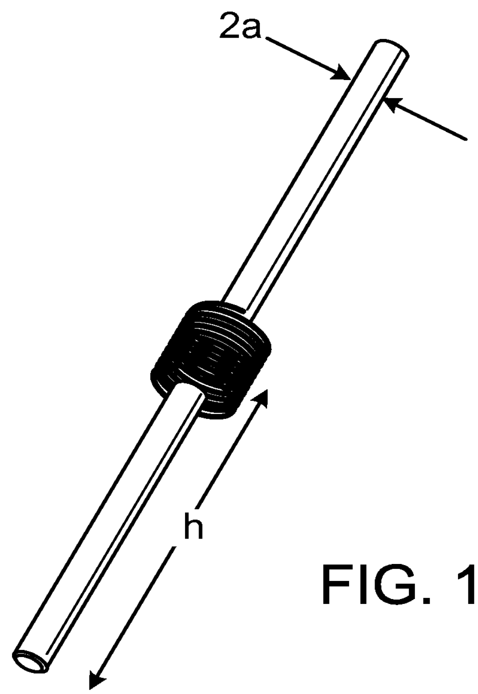
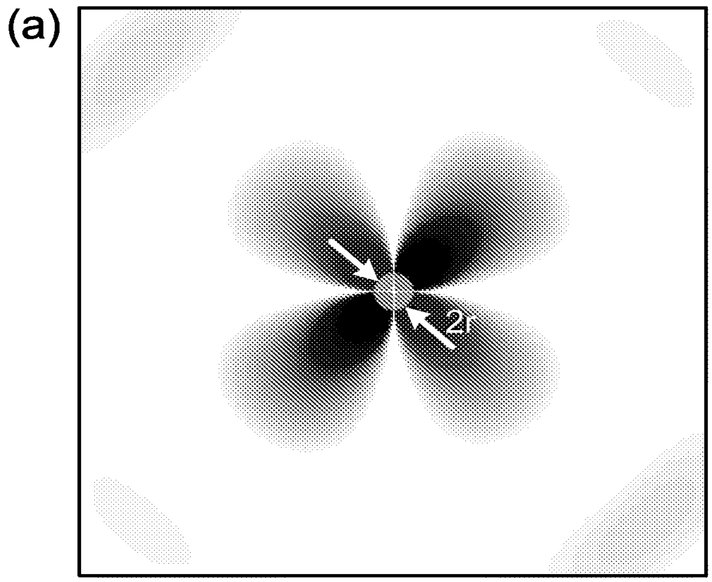


FIG. 14



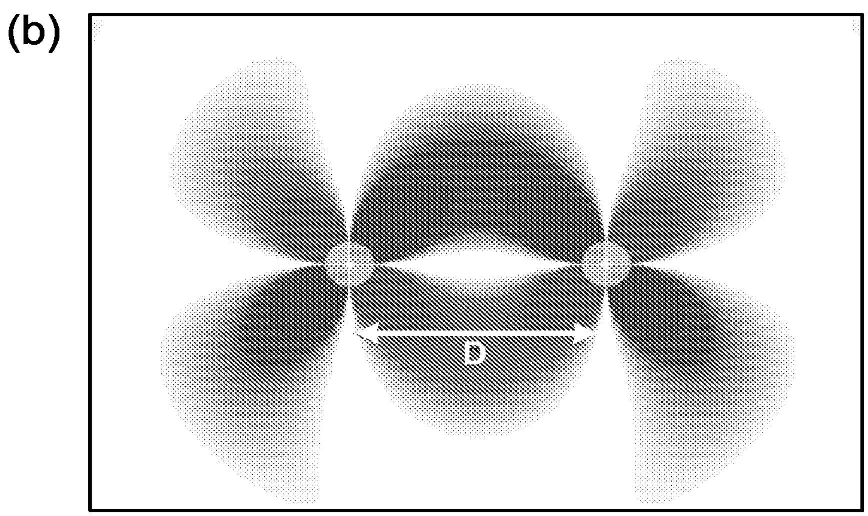


FIG. 15

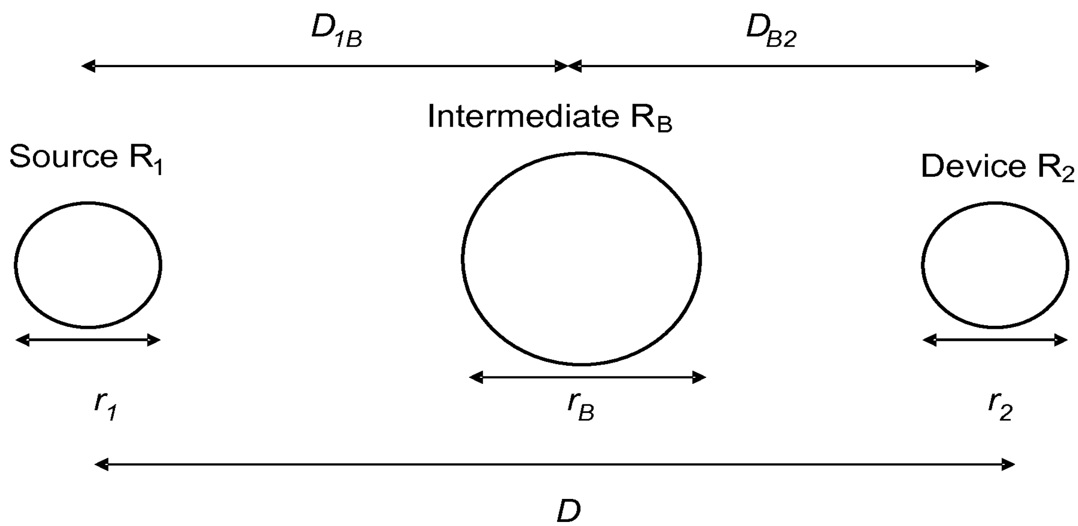


FIG. 16

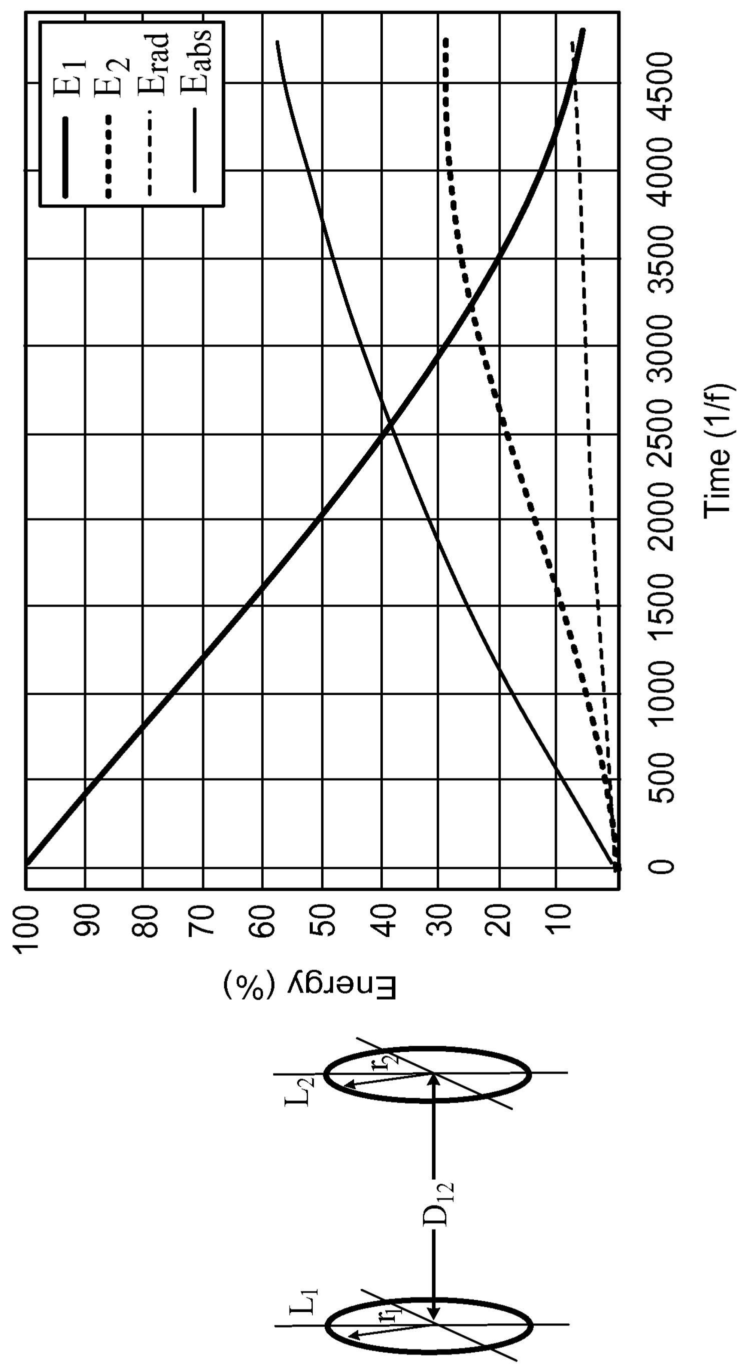

E1
E2
Erad
Eabs
Energy (%)
100
90
80
70
60
50
40
30
20
10
0
500
1000
1500
2000
2500
3000
3500
4000
4500
Time (1/f)
L1
r1
D12
L2
r2

FIG. 17A

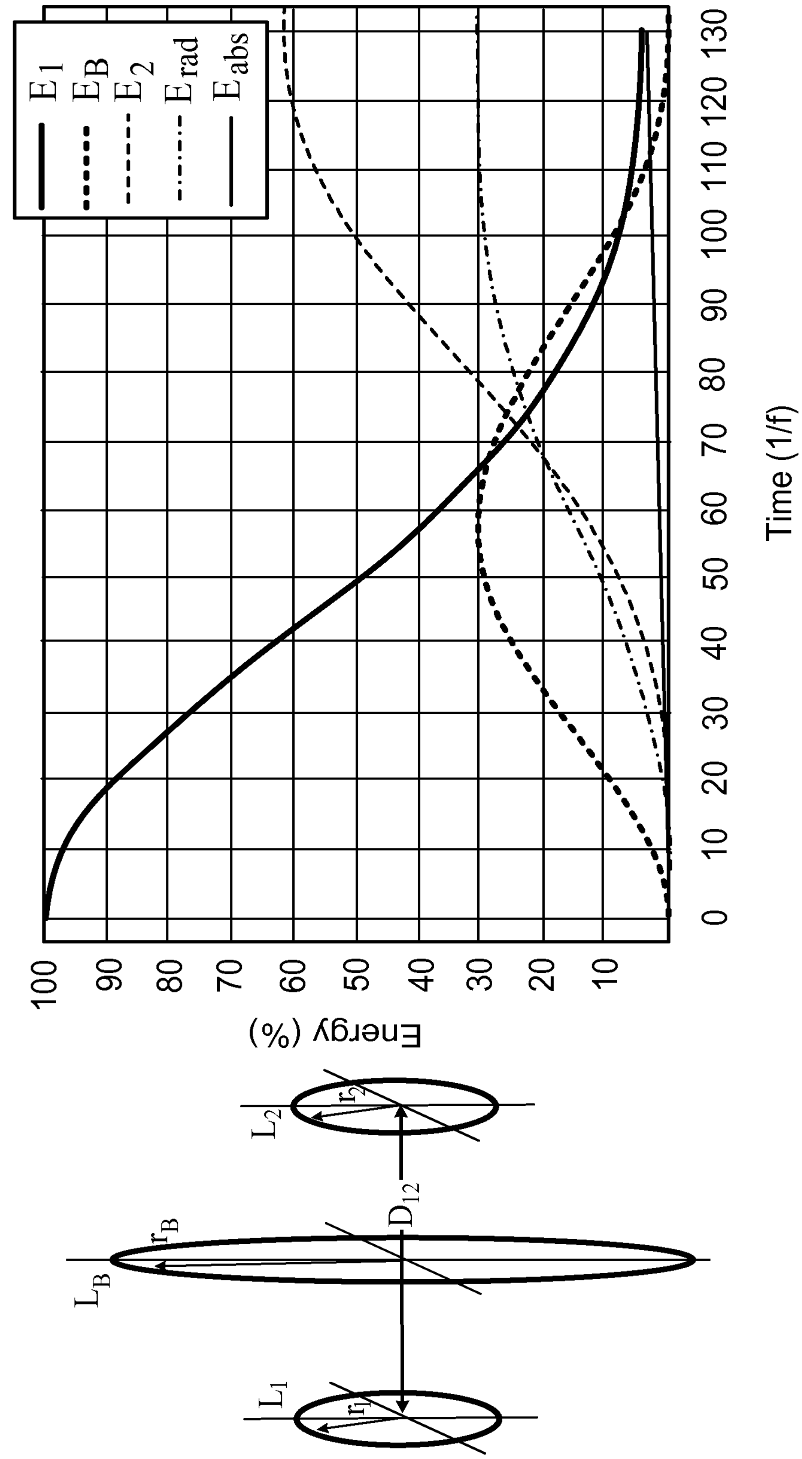

E1
EB
E2
Erad
Eabs
Energy (%)
100
90
80
70
60
50
40
30
20
10
0
10
20
30
40
50
60
70
80
90
100
110
120
130
Time (1/f)
L1
r1
LB
rB
D12
L2
r2

FIG. 17B

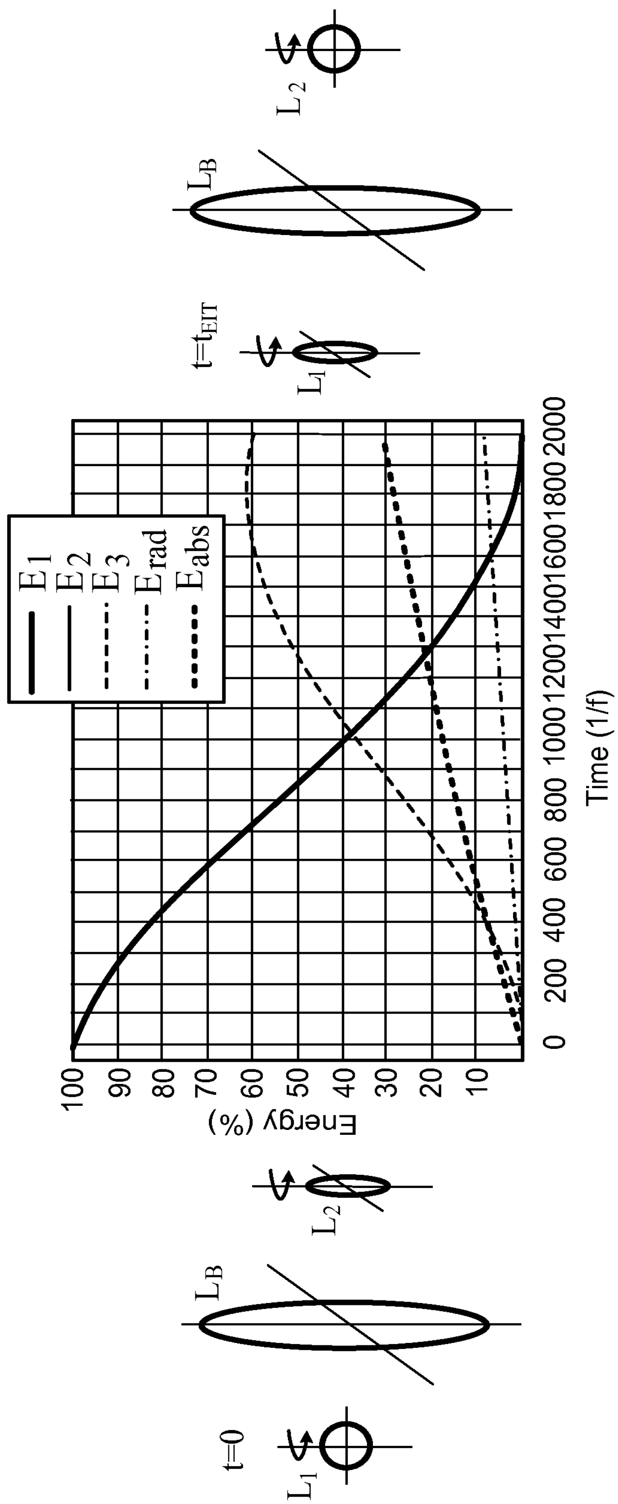

t=0
L_1
L_B
L_2
$t=t_{EIT}$
L_1
L_B
L_2
E_1
E_2
E_3
E_{rad}
E_{abs}
Energy (%)
100 90 80 70 60 50 40 30 20 10
0 200 400 600 800 1000 1200 1400 1600 1800 2000
Time (1/f)

FIG. 17C

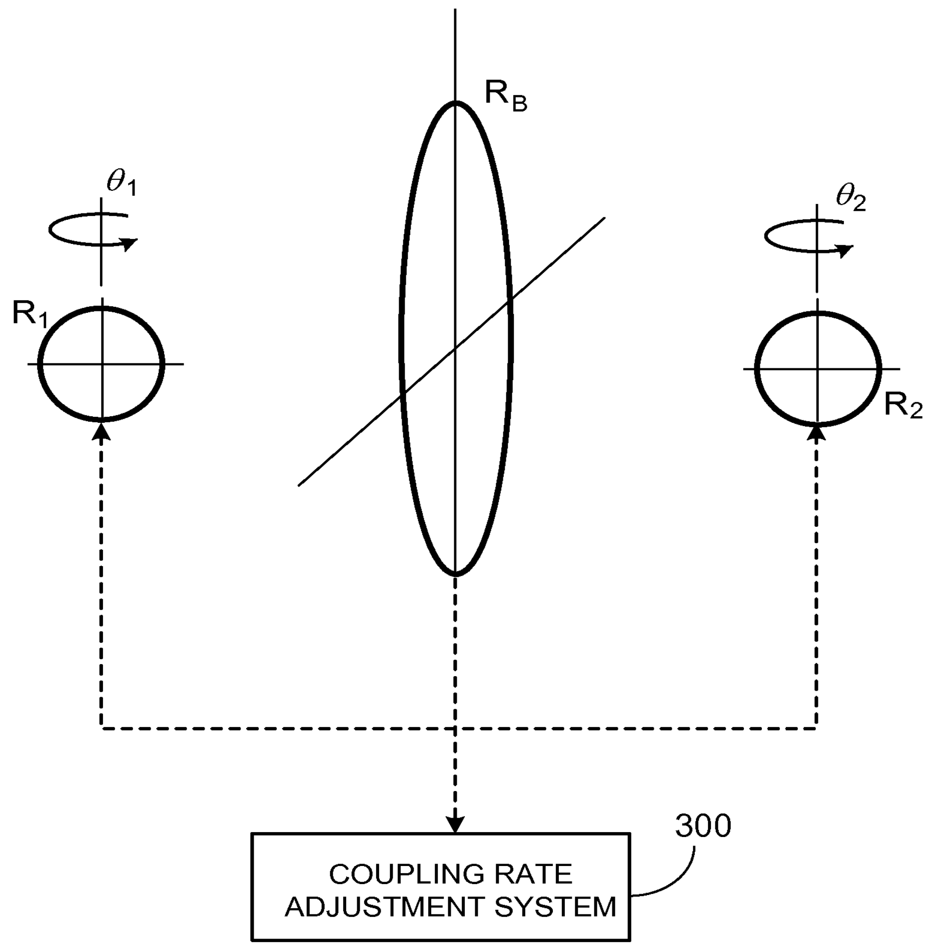


FIG. 18

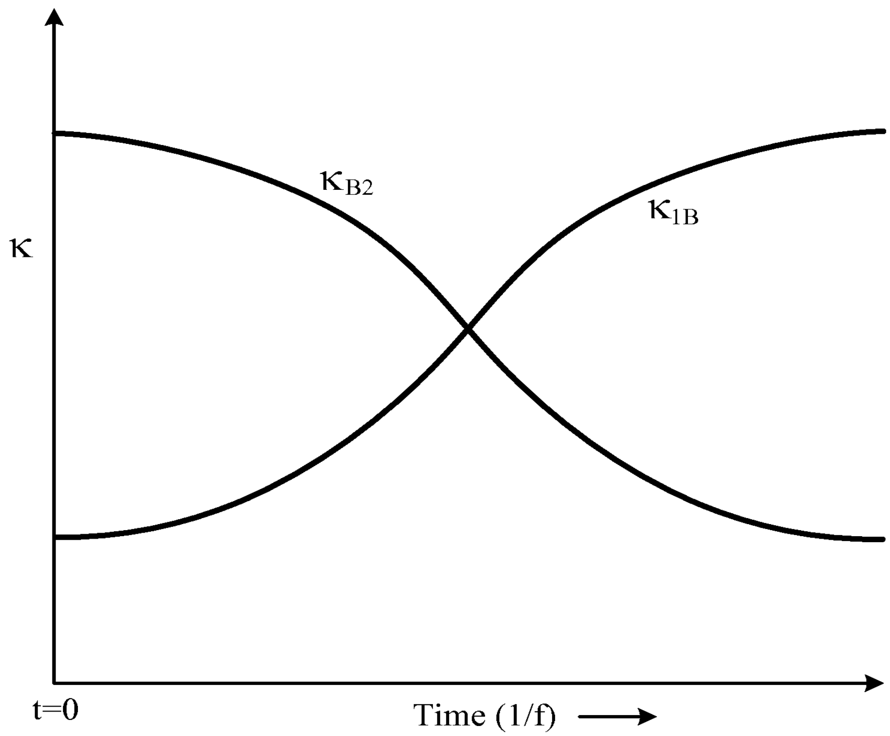


FIG. 19

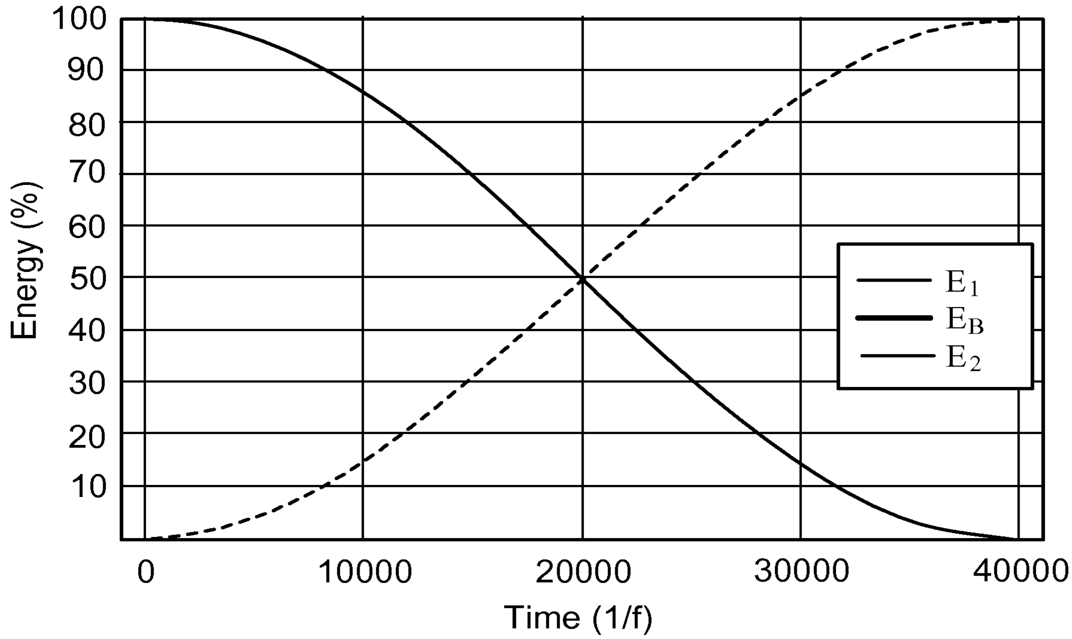

100
90
80
70
60
50
40
30
20
10
Energy (%)
0
10000
20000
30000
40000
Time (1/f)
E_1
E_B
E_2

FIG. 20

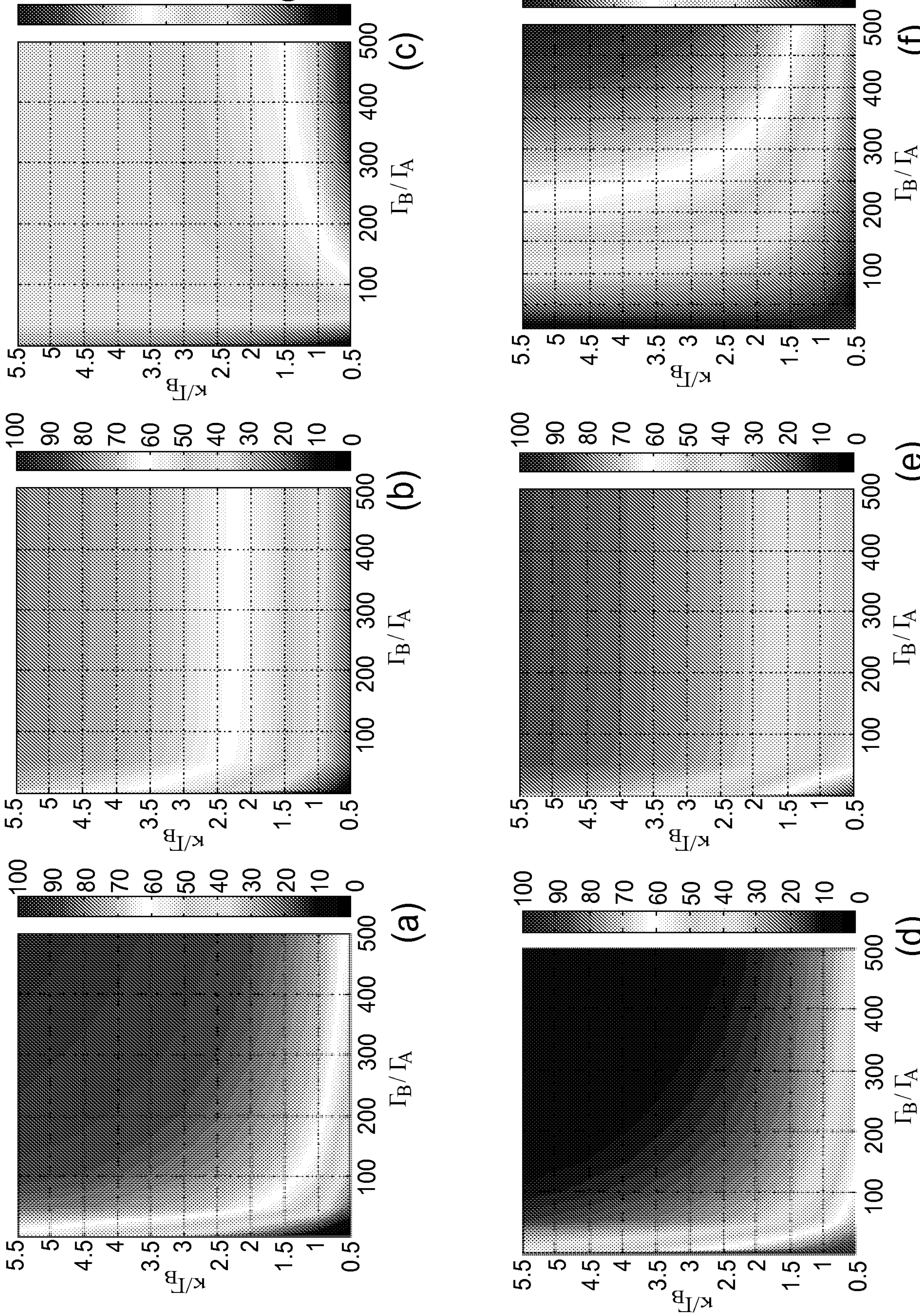


FIG. 21

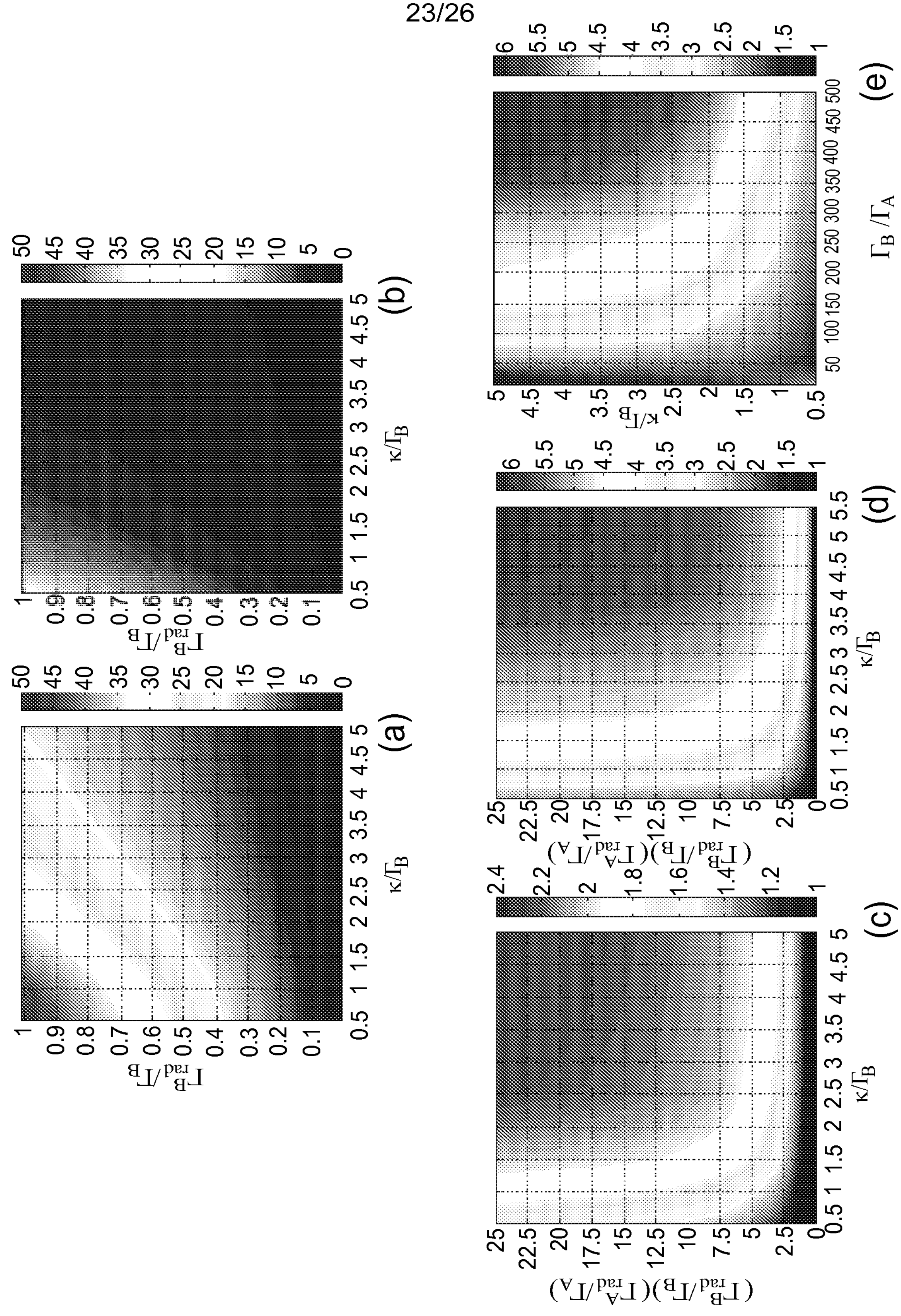

(a)
(b)
(c)
(d)
(e)
κ/Γ_B
Γ_B^{rad}/Γ_B
$(\Gamma_B^{rad}/\Gamma_B)(\Gamma_A^{rad}/\Gamma_A)$
Γ_B/Γ_A

FIG. 22

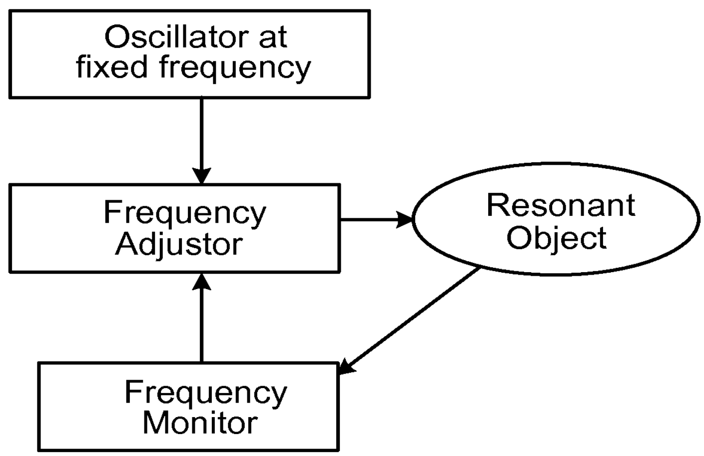


FIG. 23a

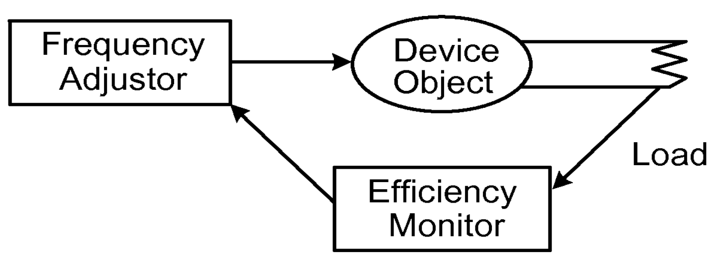
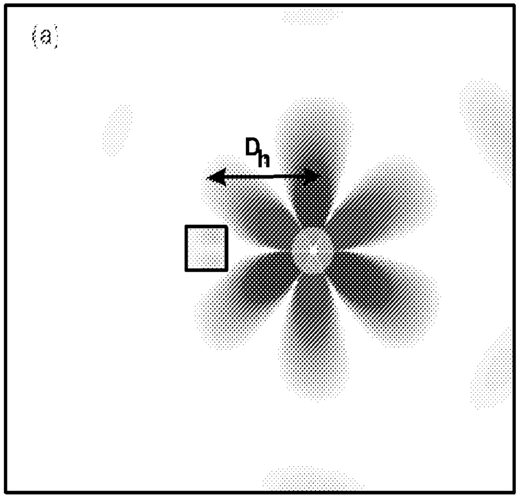
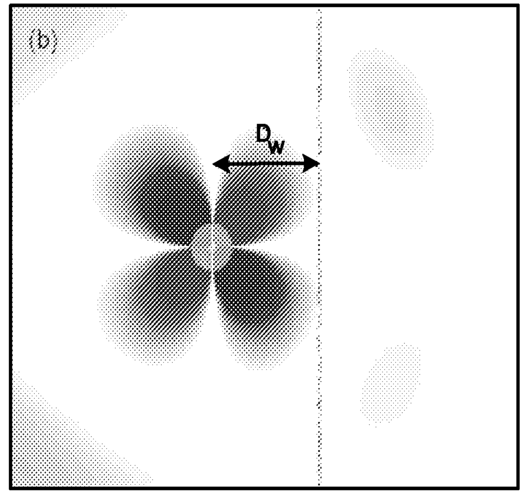


FIG. 23b



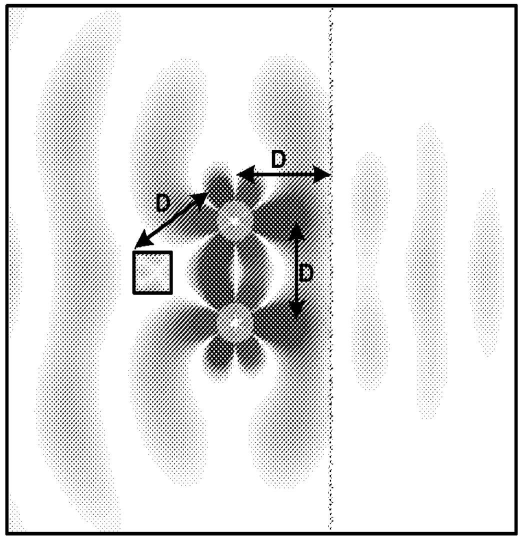
Disk with "human"	D_h/r	$Q^{abs}_{c\text{-}h}$	$Q^{rad}_{c[h]}$	$Q_{c[h]}$
Re{ε}=147.7, m=2 $\lambda/r \approx 20$ $Q^{abs}_c \approx 10096$	3	230	981	183
	5	2917	1984	1057
	7	11573	2230	1578
	10	41496	2201	1732
Re{ε}=65.6, m=3 $\lambda/r \approx 10$ $Q^{abs}_c \approx 10096$	3	1827	6197	1238
	5	**58431**	**11808**	**4978**
	7	249748	9931	4908
	10	867552	9078	4754

FIG. 24a



Disk with "wall"	D_w/r	$Q^{abs}_{c\text{-}w}$	$Q^{rad}_{c[w]}$	$Q_{c[w]}$
Re{ε}=147.7, m=2 $\lambda/r \approx 20$ $Q^{abs}_c \approx 10098$	3	16725	1235	1033
	5	**31659**	**1922**	**1536**
	7	49440	2389	1859
	10	82839	2140	1729
Re{ε}=65.6, m=3 $\lambda/r \approx 10$ $Q^{abs}_c \approx 10097$	3	53154	6228	3592
	5	127402	10988	5053
	7	159192	10168	4910
	10	191506	9510	4775

FIG. 24b



Two disks with "human" and "wall"	D/r	$Q^{abs}_{c\text{-}h}$	$Q^{abs}_{c\text{-}w}$	$Q^{rad}_{c[hw]}$	$Q_{c[hw]} = \omega/2\Gamma_{c[hw]}$	$\omega/2\kappa_{[hw]}$	$\kappa_{[hw]}/\Gamma_{c[hw]}$
Re{ε}=147.7, m=2	3	3300	12774	536	426	48	8.8
$\lambda/r \approx 20$	5	5719	26333	1600	1068	322	3.3
$Q^{abs}_c \approx 10100$	7	13248	50161	3542	2097	973	2.2
	10	18447	68460	3624	2254	1768	1.3
Re{ε}=65.6, m=3	3	2088	36661	6764	1328	141	9.4
$\lambda/r \approx 10$	**5**	**72137**	**90289**	**11945**	**4815**	**2114**	**2.3**
$Q^{abs}_c \approx 10100$	7	237822	129094	12261	5194	8307	0.6

FIG. 24c

INTERNATIONAL SEARCH REPORT

International application No. PCT/US 09/59244

A. CLASSIFICATION OF SUBJECT MATTER

IPC(8) - H01P 7/00 (2009.01)
USPC - 333/219

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
USPC - 333/219

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
USPC - 333/219; 307/126; 455/522 (keyword limited - see terms below)

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)
pubwest (DB=PGPB,USPT,EPAB,JPAB), Google Scholar
Terms - wireless energy, coupling rate, resonate, energy accumulation, near-field, transfer, intermediate

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y	US 2007/0222542 A1 (Joannopoulos et al.) 27 September 2007 (27.09.2007), abstract, para [0013], [0015], [0016], [0026], [0027], [0035]	1-4, 6, 13, 20-24, 28-32, 50-53, 80, 81
Y	US 3,517,350 A (Beaver) 23 June 1970 (23.06.1970), col 2, ln 69-72; col 7, ln 22-39	1-4, 6, 13, 20-24, 28-32, 50-53, 80, 81
X, P	US 2009/0153273 A1 (Chen) 18 June 2009 (18.06.2009), entire document	1-4, 6, 13, 20-24, 28-32, 50-53, 80, 81
X, P	WO 2008/118178 A1 (Karalis et al.) 02 October 2008 (02.10.2008), entire document	1-4, 6, 13, 20-24, 28-32, 50-53, 80, 81
A	US 6,960,968 B2 (Odendaal et al.) 01 November 2005 (01.11.2005), entire document	1-4, 6, 13, 20-24, 28-32, 50-53, 80, 81
A	US 7,069,064 B2 (Gevorgian et al.) 27 June 2006 (27.06.2006), entire document	1-4, 6, 13, 20-24, 28-32, 50-53, 80, 81
A	US 1,119,732 A (Tesla) 01 December 1914 (01.12.1914), entire document	1-4, 6, 13, 20-24, 28-32, 50-53, 80, 81

☐ Further documents are listed in the continuation of Box C. ☐

* Special categories of cited documents:
"A" document defining the general state of the art which is not considered to be of particular relevance
"E" earlier application or patent but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search
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Form PCT/ISA/210 (second sheet) (April 2007)

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US 09/59244

Box No. II Observations where certain claims were found unsearchable (Continuation of item 2 of first sheet)

This international search report has not been established in respect of certain claims under Article 17(2)(a) for the following reasons:

1. ☐ Claims Nos.:
because they relate to subject matter not required to be searched by this Authority, namely:

2. ☐ Claims Nos.:
because they relate to parts of the international application that do not comply with the prescribed requirements to such an extent that no meaningful international search can be carried out, specifically:

3. ☒ Claims Nos.: 5, 7-12, 14-19, 25-27, 33-49 and 54-79
because they are dependent claims and are not drafted in accordance with the second and third sentences of Rule 6.4(a).

Box No. III Observations where unity of invention is lacking (Continuation of item 3 of first sheet)

This International Searching Authority found multiple inventions in this international application, as follows:

1. ☐ As all required additional search fees were timely paid by the applicant, this international search report covers all searchable claims.

2. ☐ As all searchable claims could be searched without effort justifying additional fees, this Authority did not invite payment of additional fees.

3. ☐ As only some of the required additional search fees were timely paid by the applicant, this international search report covers only those claims for which fees were paid, specifically claims Nos.:

4. ☐ No required additional search fees were timely paid by the applicant. Consequently, this international search report is restricted to the invention first mentioned in the claims; it is covered by claims Nos.:

Remark on Protest

☐ The additional search fees were accompanied by the applicant's protest and, where applicable, the payment of a protest fee.

☐ The additional search fees were accompanied by the applicant's protest but the applicable protest fee was not paid within the time limit specified in the invitation.

☐ No protest accompanied the payment of additional search fees.

Form PCT/ISA/210 (continuation of first sheet (2)) (April 2007)

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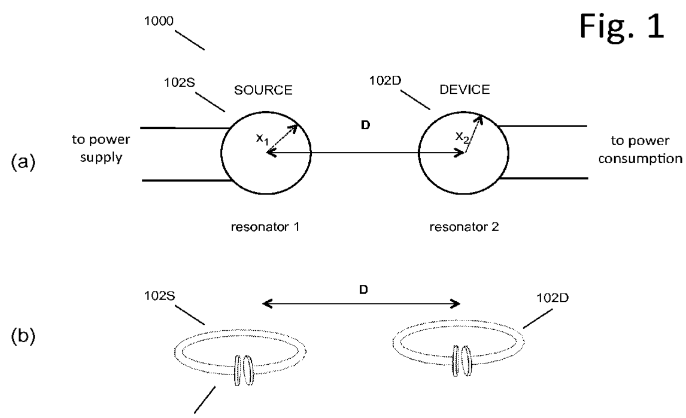
(54) **Title**: WIRELESS ENERGY TRANSFER IN LOSSY ENVIRONMENTS

Fig. 1



(57) **Abstract**: Described herein are improved configurations for a wireless power transfer for electronic devices that include at least one source magnetic resonator including a capacitively- loaded conducting loop coupled to a power source and configured to generate an oscillating magnetic field and at least one device magnetic resonator, distal from said source resonators, comprising a capacitively-loaded conducting loop configured to convert said oscillating magnetic fields into electrical energy, wherein at least one said resonator has a keep-out zone around the resonator that surrounds the resonator with a layer of non-lossy material.

WO 2010/093997 A1

WIRELESS ENERGY TRANSFER IN LOSSY ENVIRONMENTS

CROSS-REFERENCE TO RELATED APPLICATIONS

[0001] This application claims priority to the following U.S. Patent Applications, each of which is hereby incorporated by reference in its entirety; U.S. Patent App. No. 12/567,716 filed September 25, 2009; U.S. App. No. 61/178,508 filed May 15, 2009; U.S. App. No. 61/182,768 filed June 1, 2009; U.S. App. No. 61/156,764 filed March 2, 2009; U.S. App. No. 61/152,390 filed February 13, 2009; U.S. App. No. 61/163,695 filed March 26, 2009; U.S. App. No. 61/172,633 filed April 24, 2009; U.S. App. No. 61/169,240 filed April 14, 2009; U.S. App. No. 61/173,747 filed April 29, 2009; U.S. Patent App. No. 12/639,489 filed December 16, 2009; and U.S. Patent App. No. 12/647,705 filed December 28, 2009.

BACKGROUND

[0002] Field:

[0003] This disclosure relates to wireless energy transfer, also referred to as wireless power transmission.

[0004] Description of the Related Art:

[0005] Energy or power may be transferred wirelessly using a variety of known radiative, or far-field, and non-radiative, or near-field, techniques. For example, radiative wireless information transfer using low-directionality antennas, such as those used in radio and cellular communications systems and home computer networks, may be considered wireless energy transfer. However, this type of radiative transfer is very inefficient because only a tiny portion of the supplied or radiated power, namely, that portion in the direction of, and overlapping with, the receiver is picked up. The vast majority of the power is radiated away in all the other directions and lost in free space. Such inefficient power transfer may be acceptable for data transmission, but is not practical for transferring useful amounts of electrical energy for the purpose of doing work, such as for powering or charging electrical devices. One way to improve the transfer efficiency of some radiative energy transfer schemes is to use directional antennas to confine and preferentially direct the radiated energy towards a receiver. However, these directed radiation schemes may require an uninterruptible line-of-sight and potentially complicated tracking and steering mechanisms in the case of mobile transmitters and/or receivers. In addition, such schemes may pose hazards to objects or people that cross or intersect the beam when modest to high amounts of power are being transmitted. A known non-radiative, or near-field, wireless energy transfer scheme, often referred to as either induction or traditional induction, does not (intentionally) radiate power, but uses an oscillating current passing through a primary coil, to generate an oscillating magnetic near-field that induces currents in a near-by receiving or secondary coil. Traditional induction schemes have demonstrated the transmission of modest to large amounts of power, however only over very short distances, and with very small offset tolerances between the primary power supply unit and the secondary receiver unit. Electric transformers and proximity chargers are examples of devices that utilize this known short range, near-field energy transfer scheme.

[0006] Therefore a need exists for a wireless power transfer scheme that is capable of transferring useful amounts of electrical power over mid-range distances or alignment offsets. Such a wireless power transfer scheme should enable useful energy transfer over greater distances and alignment offsets than those realized with traditional induction schemes, but without the limitations and risks inherent in radiative transmission schemes.

SUMMARY

[0007] There is disclosed herein a non-radiative or near-field wireless energy transfer scheme that is capable of transmitting useful amounts of power over mid-range distances and alignment offsets. This inventive technique uses coupled electromagnetic resonators with long-lived oscillatory resonant modes to transfer power from a power supply to a power drain. The technique is general and may be applied to a wide range of resonators, even where the specific examples disclosed herein relate to electromagnetic resonators. If the resonators are designed such that the energy stored by the electric field is primarily confined within the structure and that the energy stored by the magnetic field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant magnetic near-field. These types of resonators may be referred to as magnetic resonators. If the resonators are designed such that the energy stored by the magnetic field is primarily confined within the structure and that the energy stored by the electric field is primarily in the region surrounding the resonator. Then, the

energy exchange is mediated primarily by the resonant electric near-field. These types of resonators may be referred to as electric resonators. Either type of resonator may also be referred to as an electromagnetic resonator. Both types of resonators are disclosed herein.

[0008] The omni-directional but stationary (non-lossy) nature of the near-fields of the resonators we disclose enables efficient wireless energy transfer over mid-range distances, over a wide range of directions and resonator orientations, suitable for charging, powering, or simultaneously powering and charging a variety of electronic devices. As a result, a system may have a wide variety of possible applications where a first resonator, connected to a power source, is in one location, and a second resonator, potentially connected to electrical/electronic devices, batteries, powering or charging circuits, and the like, is at a second location, and where the distance from the first resonator to the second resonator is on the order of centimeters to meters. For example, a first resonator connected to the wired electricity grid could be placed on the ceiling of a room, while other resonators connected to devices, such as robots, vehicles, computers, communication devices, medical devices, and the like, move about within the room, and where these devices are constantly or intermittently receiving power wirelessly from the source resonator. From this one example, one can imagine many applications where the systems and methods disclosed herein could provide wireless power across mid-range distances, including consumer electronics, industrial applications, infrastructure power and lighting, transportation vehicles, electronic games, military applications, and the like.

[0009] Energy exchange between two electromagnetic resonators can be optimized when the resonators are tuned to substantially the same frequency and when the losses in the system are minimal. Wireless energy transfer systems may be designed so that the "coupling-time" between resonators is much shorter than the resonators' "loss-times". Therefore, the systems and methods described herein may utilize high quality factor (high-Q) resonators with low intrinsic-loss rates. In addition, the systems and methods described herein may use sub-wavelength resonators with near-fields that extend significantly longer than the characteristic sizes of the resonators, so that the near-fields of the resonators that exchange energy overlap at mid-range distances. This is a regime of operation that has not been practiced before and that differs significantly from traditional induction designs.

[0010] It is important to appreciate the difference between the high-Q magnetic resonator scheme disclosed here and the known close-range or proximity inductive schemes, namely, that those known schemes do not conventionally utilize high-Q resonators. Using coupled-mode theory (CMT), (see, for example, *Waves and Fields in Optoelectronics*, H.A. Haus, Prentice Hall, 1984), one may show that a high-Q resonator-coupling mechanism can enable orders of magnitude more efficient power delivery between resonators spaced by mid-range distances than is enabled by traditional inductive schemes. Coupled high-Q resonators have demonstrated efficient energy transfer over mid-range distances and improved efficiencies and offset tolerances in short range energy transfer applications.

[0011] The systems and methods described herein may provide for near-field wireless energy transfer via strongly coupled high-Q resonators, a technique with the potential to transfer power levels from picowatts to kilowatts, safely, and over distances much larger than have been achieved using traditional induction techniques. Efficient energy transfer may be realized for a variety of general systems of strongly coupled resonators, such as systems of strongly coupled acoustic resonators, nuclear resonators, mechanical resonators, and the like, as originally described by researchers at M.I.T. in their publications, "Efficient wireless non-radiative mid-range energy transfer", *Annals of Physics*, vol. 323, Issue 1, p. 34 (2008) and "Wireless Power Transfer via Strongly Coupled Magnetic Resonances", *Science*, vol. 317, no. 5834, p. 83, (2007). Disclosed herein are electromagnetic resonators and systems of coupled electromagnetic resonators, also referred to more specifically as coupled magnetic resonators and coupled electric resonators, with operating frequencies below 10 GHz.

[0012] This disclosure describes wireless energy transfer technologies, also referred to as wireless power transmission technologies. Throughout this disclosure, we may use the terms wireless energy transfer, wireless power transfer, wireless power transmission, and the like, interchangeably. We may refer to supplying energy or power from a source, an AC or DC source, a battery, a source resonator, a power supply, a generator, a solar panel, and thermal collector, and the like, to a device, a remote device, to multiple remote devices, to a device resonator or resonators, and the like. We may describe intermediate resonators that extend the range of the wireless energy transfer system by allowing energy to hop, transfer through, be temporarily stored, be partially dissipated, or for the transfer to be mediated in any way, from a source resonator to any combination of other device and intermediate resonators, so that energy transfer networks, or strings, or extended paths may be realized. Device resonators may receive energy from a source resonator, convert a portion of that energy to electric power for powering or charging a device, and simultaneously pass a portion of the received energy onto other device or mobile device resonators. Energy may be transferred from a source resonator to multiple device resonators, significantly extending the distance over which energy may be wirelessly transferred. The wireless power transmission systems may be implemented using a variety of system architectures and resonator designs. The systems may include a single source or multiple sources transmitting power to a single device or multiple devices. The resonators may be designed to be source or device resonators, or they may be designed to be repeaters. In some cases, a resonator may be a device and source resonator simultaneously, or it may be switched from operating as a source to operating as a device or a repeater. One skilled in the art will understand that a variety of system architectures may be supported by the wide range of resonator designs and functionalities described in this application.

[0013] In the wireless energy transfer systems we describe, remote devices may be powered directly, using the wirelessly supplied power or energy, or the devices may be coupled to an energy storage unit such as a battery, a super-capacitor, an ultra-capacitor, or the like (or other kind of power drain), where the energy storage unit may be charged or re-charged wirelessly, and/or where the wireless power transfer mechanism is simply supplementary to the main power source of the device. The devices may be powered by hybrid battery/energy storage devices such as batteries with integrated storage capacitors and the like. Furthermore, novel battery and energy storage devices may be designed to take advantage of the operational improvements enabled by wireless power transmission systems.

[0014] Other power management scenarios include using wirelessly supplied power to recharge batteries or charge energy storage units while the devices they power are turned off, in an idle state, in a sleep mode, and the like. Batteries or energy storage units may be charged or recharged at high (fast) or low (slow) rates. Batteries or energy storage units may be trickle charged or float charged. Multiple devices may be charged or powered simultaneously in parallel or power delivery to multiple devices may be serialized such that one or more devices receive power for a period of time after which other power delivery is switched to other devices. Multiple devices may share power from one or more sources with one or more other devices either simultaneously, or in a time multiplexed manner, or in a frequency multiplexed manner, or in a spatially multiplexed manner, or in an orientation multiplexed manner, or in any combination of time and frequency and spatial and orientation multiplexing. Multiple devices may share power with each other, with at least one device being reconfigured continuously, intermittently, periodically, occasionally, or temporarily, to operate as wireless power sources. It would be understood by one of ordinary skill in the art that there are a variety of ways to power and/or charge devices, and the variety of ways could be applied to the technologies and applications described herein.

[0015] Wireless energy transfer has a variety of possible applications including for example, placing a source (e.g. one connected to the wired electricity grid) on the ceiling, under the floor, or in the walls of a room, while devices such as robots, vehicles, computers, PDAs or similar are placed or move freely within the room. Other applications may include powering or recharging electric-engine vehicles, such as buses and/or hybrid cars and medical devices, such as wearable or implantable devices. Additional example applications include the ability to power or recharge autonomous electronics (e.g. laptops, cell-phones, portable music players, house-hold robots, GPS navigation systems, displays, etc), sensors, industrial and manufacturing equipment, medical devices and monitors, home appliances and tools (e.g. lights, fans, drills, saws, heaters, displays, televisions, counter-top appliances, etc.), military devices, heated or illuminated clothing, communications and navigation equipment, including equipment built into vehicles, clothing and protective-wear such as helmets, body armor and vests, and the like, and the ability to transmit power to physically isolated devices such as to implanted medical devices, to hidden, buried, implanted or embedded sensors or tags, to and/or from roof-top solar panels to indoor distribution panels, and the like.

[0016] In one aspect, disclosed herein is a system including a source resonator having a Q-factor Q_1 and a characteristic size x_1, coupled to a power generator with direct electrical connections; and a second resonator having a Q-factor Q_2 and a characteristic size x_2, coupled to a load with direct electrical connections, and located a distance D from the source resonator, wherein the source resonator and the second resonator are coupled to exchange energy wirelessly among the source resonator and the second resonator in order to transmit power from the power generator to the load, and wherein $\sqrt{Q_1 Q_2}$ is greater than 100.

[0017] Q_1 may be greater than 100 and Q_2 may be less than 100. Q_1 may be greater than 100 and Q_2 may be greater than 100. A useful energy exchange may be maintained over an operating distance from 0 to D, where D is larger than the smaller of x_1 and x_2. At least one of the source resonator and the second resonator may be a coil of at least one turn of a conducting material connected to a first network of capacitors. The first network of capacitors may include at least one tunable capacitor. The direct electrical connections of at least one of the source resonator to the ground terminal of the power generator and the second resonator to the ground terminal of the load may be made at a point on an axis of electrical symmetry of the first network of capacitors. The first network of capacitors may include at least one tunable butterfly-type capacitor, wherein the direct electrical connection to the ground terminal is made on a center terminal of the at least one tunable butterfly-type capacitor. The direct electrical connection of at least one of the source resonator to the power generator and the second resonator to the load may be made via a second network of capacitors, wherein the first network of capacitors and the second network of capacitors form an impedance matching network. The impedance matching network may be designed to match the coil to a characteristic impedance of the power generator or the load at a driving frequency of the power generator.

[0018] At least one of the first network of capacitors and the second network of capacitors may include at least one tunable capacitor. The first network of capacitors and the second network of capacitors may be adjustable to change an impedance of the impedance matching network at a driving frequency of the power generator. The first network of capacitors and the second network of capacitors may be adjustable to match the coil to the characteristic impedance of the power generator or the load at a driving frequency of the power generator. At least one of the first network of capacitors and the second network of capacitors may include at least one fixed capacitor that reduces a voltage across the at least one tunable capacitor. The direct electrical connections of at least one of the source resonator to the power generator and the second resonator to the load may be configured to substantially preserve a resonant mode. At least one of the source resonator and the second resonator may be a tunable resonator. The source resonator may be physically separated from the power generator and the second resonator may be physically separated from the load. The second resonator may be coupled to a power conversion circuit to deliver DC power to the load. The second resonator may be coupled to a power conversion circuit to deliver AC power to the load. The second resonator may be coupled to a power conversion circuit to deliver both AC and DC power to the load. The second resonator may be coupled to a power conversion circuit to deliver power to a plurality of loads.

[0019] In another aspect, a system disclosed herein includes a source resonator having a Q-factor Q_1 and a characteristic size x_1, and a second resonator having a Q-factor Q_2

and a characteristic size x_2, and located a distance D from the source resonator; wherein the source resonator and the second resonator are coupled to exchange energy wirelessly among the source resonator and the second resonator; and wherein $\sqrt{Q_1 Q_2}$ is greater than 100, and wherein at least one of the resonators is enclosed in a low loss tangent material.

[0020] In another aspect, a system disclosed herein includes a source resonator having a Q-factor Q_1 and a characteristic size x_1, and a second resonator having a Q-factor Q_2 and a characteristic size x_2, and located a distance D from the source resonator; wherein the source resonator and the second resonator are coupled to exchange energy wirelessly among the source resonator and the second resonator, and wherein $\sqrt{Q_1 Q_2}$ is greater than 100; and wherein at least one of the resonators includes a coil of a plurality of turns of a conducting material connected to a network of capacitors, wherein the plurality of turns are in a common plane, and wherein a characteristic thickness of the at least one of the resonators is much less than a characteristic size of the at least one of the resonators.

[0021] Throughout this disclosure we may refer to the certain circuit components such as capacitors, inductors, resistors, diodes, switches and the like as circuit components or elements. We may also refer to series and parallel combinations of these components as elements, networks, topologies, circuits, and the like. We may describe combinations of capacitors, diodes, varactors, transistors, and/or switches as adjustable impedance networks, tuning networks, matching networks, adjusting elements, and the like. We may also refer to "self-resonant" objects that have both capacitance, and inductance distributed (or partially distributed, as opposed to solely lumped) throughout the entire object. It would be understood by one of ordinary skill in the art that adjusting and controlling variable components within a circuit or network may adjust the performance of that circuit or network and that those adjustments may be described generally as tuning, adjusting, matching, correcting, and the like. Other methods to tune or adjust the operating point of the wireless power transfer system may be used alone, or in addition to adjusting tunable components such as inductors and capacitors, or banks of inductors and capacitors.

[0022] Unless otherwise defined, all technical and scientific terms used herein have the same meaning as commonly understood by one of ordinary skill in the art to which this disclosure belongs. In case of conflict with publications, patent applications, patents, and other references mentioned or incorporated herein by reference, the present specification, including definitions, will control.

[0023] Any of the features described above may be used, alone or in combination, without departing from the scope of this disclosure. Other features, objects, and advantages of the systems and methods disclosed herein will be apparent from the following detailed description and figures.

BRIEF DESCRIPTION OF FIGURES

[0024] Fig. 1 (a) and (b) depict exemplary wireless power systems containing a source resonator 1 and device resonator 2 separated by a distance D.

[0025] Fig. 2 shows an exemplary resonator labeled according to the labeling convention described in this disclosure. Note that there are no extraneous objects or additional resonators shown in the vicinity of resonator 1.

[0026] Fig. 3 shows an exemplary resonator in the presence of a "loading" object, labeled according to the labeling convention described in this disclosure.

[0027] Fig. 4 shows an exemplary resonator in the presence of a "perturbing" object, labeled according to the labeling convention described in this disclosure.

[0028] Fig. 5 shows a plot of efficiency, η, vs. strong coupling factor, $U = \kappa / \sqrt{\Gamma_s \Gamma_d} = k\sqrt{Q_s Q_d}$.

[0029] Fig. 6 (a) shows a circuit diagram of one example of a resonator (b) shows a diagram of one example of a capacitively-loaded inductor loop magnetic resonator, (c) shows a drawing of a self-resonant coil with distributed capacitance and inductance, (d) shows a simplified drawing of the electric and magnetic field lines associated with an exemplary magnetic resonator of the current disclosure, and (e) shows a diagram of one example of an electric resonator.

[0030] Fig. 7 shows a plot of the "quality factor", Q (solid line), as a function of frequency, of an exemplary resonator that may be used for wireless power transmission at MHz frequencies. The absorptive Q (dashed line) increases with frequency, while the radiative Q (dotted line) decreases with frequency, thus leading the overall Q to peak at a particular frequency.

[0031] Fig. 8 shows a drawing of a resonator structure with its characteristic size, thickness and width indicated.

[0032] Fig. 9 (a) and (b) show drawings of exemplary inductive loop elements.

[0033] Fig. 10 (a) and (b) show two examples of trace structures formed on printed circuit boards and used to realize the inductive element in magnetic resonator structures.

[0034] Fig. 11 (a) shows a perspective view diagram of a planar magnetic resonator, (b) shows a perspective view diagram of a two planar magnetic resonator with various geometries, and c) shows is a perspective view diagram of a two planar magnetic resonators separated by a distance D.

[0035] Fig. 12 is a perspective view of an example of a planar magnetic resonator.

[0036] Fig. 13 is a perspective view of a planar magnetic resonator arrangement with a circular resonator coil.

[0037] Fig. 14 is a perspective view of an active area of a planar magnetic resonator.

[0038] Fig. 15 is a perspective view of an application of the wireless power transfer system with a source at the center of a table powering several devices placed around the source.

[0039] Fig. 16(a) shows a 3D finite element model of a copper and magnetic material structure driven by a square loop of current around the choke point at its center. In this example, a structure may be composed of two boxes made of a conducting material such as copper, covered by a layer of magnetic material, and connected by a block of magnetic material. The inside of the two conducting boxes in this example would be shielded from AC electromagnetic fields generated outside the boxes and may house lossy objects that might lower the Q of the resonator or sensitive components that might be adversely affected by the AC electromagnetic fields. Also shown are the calculated magnetic field streamlines generated by this structure, indicating that the magnetic field lines tend to follow the lower reluctance path in the magnetic material. Fig. 16(b) shows interaction, as indicated by the calculated magnetic field streamlines, between two identical structures as shown in (a). Because of symmetry, and to reduce computational complexity, only one half of the system is modeled (but the computation assumes the symmetrical arrangement of the other half).

[0040] Fig. 17 shows an equivalent circuit representation of a magnetic resonator including a conducting wire wrapped N times around a structure, possibly containing magnetically permeable material. The inductance is realized using conducting loops wrapped around a structure comprising a magnetic material and the resistors represent loss mechanisms in the system (R_{wire} for resistive losses in the loop, R_{μ} denoting the equivalent series resistance of the structure surrounded by the loop). Losses may be minimized to realize high-Q resonators.

[0041] Fig. 18 shows a Finite Element Method (FEM) simulation of two high conductivity surfaces above and below a disk composed of lossy dielectric material, in an external magnetic field of frequency 6.78 MHz. Note that the magnetic field was uniform before the disk and conducting materials were introduced to the simulated environment. This simulation is performed in cylindrical coordinates. The image is azimuthally symmetric around the r= 0 axis. The lossy dielectric disk has $\epsilon_r = 1$ and σ= 10 S/m.

[0042] Fig. 19 shows a drawing of a magnetic resonator with a lossy object in its vicinity completely covered by a high-conductivity surface.

[0043] Fig. 20 shows a drawing of a magnetic resonator with a lossy object in its vicinity partially covered by a high-conductivity surface.

[0044] Fig. 21 shows a drawing of a magnetic resonator with a lossy object in its vicinity placed on top of a high-conductivity surface.

[0045] Fig. 22 shows a diagram of a completely wireless projector.

[0046] Fig. 23 shows the magnitude of the electric and magnetic fields along a line that contains the diameter of the circular loop inductor and along the axis of the loop inductor.

[0047] Fig. 24 shows a drawing of a magnetic resonator and its enclosure along with a necessary but lossy object placed either (a) in the corner of the enclosure, as far away from the resonator structure as possible or (b) in the center of the surface enclosed by the inductive element in the magnetic resonator.

[0048] Fig. 25 shows a drawing of a magnetic resonator with a high-conductivity surface above it and a lossy object, which may be brought into the vicinity of the resonator, but above the high-conductivity sheet.

[0049] Fig. 26(a) shows an axially symmetric FEM simulation of a thin conducting (copper) cylinder or disk (20 cm in diameter, 2 cm in height) exposed to an initially uniform, externally applied magnetic field (gray flux lines) along the z-axis. The axis of symmetry is at r=0. The magnetic streamlines shown originate at $z = -\infty$, where they are spaced from r=3 cm to r=10 cm in intervals of 1 cm. The axes scales are in meters. Fig. 26 (b) shows the same structure and externally applied field as in (a), except that the conducting cylinder has been modified to include a 0.25 mm layer of magnetic material (not visible) with $\mu_r'=40$, on its outside surface. Note that the magnetic streamlines are deflected away from the cylinder significantly less than in (a).

[0050] Fig. 27 shows an axi-symmetric view of a variation based on the system shown in Fig. 26. Only one surface of the lossy material is covered by a layered structure of copper and magnetic materials. The inductor loop is placed on the side of the copper and magnetic material structure opposite to the lossy material as shown.

[0051] Fig. 28 (a) depicts a general topology of a matching circuit including an indirect coupling to a high-Q inductive element.

[0052] Fig. 28 (b) shows a block diagram of a magnetic resonator that includes a conductor loop inductor and a tunable impedance network. Physical electrical connections to this resonator may be made to the terminal connections.

[0053] Fig. 28 (c) depicts a general topology of a matching circuit directly coupled to a high-Q inductive element.

[0054] Fig. 28 (d) depicts a general topology of a symmetric matching circuit directly coupled to a high-Q inductive element and driven anti-symmetrically (balanced drive).

[0055] Fig. 28 (e) depicts a general topology of a matching circuit directly coupled to a high-Q inductive element and connected to ground at a point of symmetry of the main resonator (unbalanced drive).

[0056] Figs. 29(a) and 29(b) depict two topologies of matching circuits transformer-coupled (i.e. indirectly or inductively) to a high-Q inductive element. The highlighted portion of the Smith chart in (c) depicts the complex impedances (arising from L and R of the inductive element) that may be matched to an arbitrary real impedance Z_0 by the topology of Fig. 31(b) in the case $\omega L_2=1/\omega C_2$.

[0057] Figs. 30(a),(b),(c),(d),(e),(f) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in series with Z_0. The topologies shown in Figs. 30(a),(b),(c) are driven with a common-mode signal at the input terminals, while the topologies shown in Figs 30(d),(e),(f) are symmetric and receive a balanced drive. The highlighted portion of the Smith chart in 30(g) depicts the complex impedances that may be matched by these topologies. Figs. 30(h),(i),(j),(k),(l),(m) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in series with Z_0.

[0058] Figs. 31(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in series with Z_0. They are connected to ground at the center point of a capacitor and receive an unbalanced drive. The highlighted portion of the Smith chart in Fig. 31(d) depicts the complex impedances that may be matched by these topologies. Figs. 31(e),(f),(g) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in series with Z_0.

[0059] Figs. 32(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in series with Z_0. They are connected to ground by tapping at the center point of the inductor loop and receive an unbalanced drive. The highlighted portion of the Smith chart in (d) depicts the complex impedances that may be matched by these topologies, (e),(f),(g) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in series with Z_0.

[0060] Figs. 33(a),(b),(c),(d),(e),(f) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in parallel with Z_o. The topologies shown in Figs. 33(a),(b),(c) are driven with a common-mode signal at the input terminals, while the topologies shown in Figs 33(d),(e),(f) are symmetric and receive a balanced drive. The highlighted portion of the Smith chart in Fig. 33(g) depicts the complex impedances that may be matched by these topologies. Figs. 33(h),(i),(j),(k),(l),(m) depict six topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in parallel with Z_o.

[0061] Figs. 34(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in parallel with Z_0. They are connected to ground at the center point of a capacitor and receive an unbalanced drive. The highlighted portion of the Smith chart in (d) depicts the complex impedances that may be matched by these topologies. Figs. 34(e),(f),(g) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including inductors in parallel with Z_0.

[0062] Figs. 35(a),(b),(c) depict three topologies of matching circuits directly coupled to a high-Q inductive element and including capacitors in parallel with Z_0. They are connected to ground by tapping at the center point of the inductor loop and receive an unbalanced drive. The highlighted portion of the Smith chart in Figs. 35(d),(e), and (f) depict the complex impedances that may be matched by these topologies.

[0063] Figs. 36(a),(b),(c),(d) depict four topologies of networks of fixed and variable capacitors designed to produce an overall variable capacitance with finer tuning resolution and some with reduced voltage on the variable capacitor.

[0064] Figs. 37(a) and 37(b) depict two topologies of networks of fixed capacitors and a variable inductor designed to produce an overall variable capacitance.

[0065] Fig. 38 depicts a high level block diagram of a wireless power transmission system.

[0066] Fig. 39 depicts a block diagram of an exemplary wirelessly powered device.

[0067] Fig. 40 depicts a block diagram of the source of an exemplary wireless power transfer system.

[0068] Fig. 41 shows an equivalent circuit diagram of a magnetic resonator. The slash through the capacitor symbol indicates that the represented capacitor may be fixed or variable. The port parameter measurement circuitry may be configured to measure certain electrical signals and may measure the magnitude and phase of signals.

[0069] Fig. 42 shows a circuit diagram of a magnetic resonator where the tunable impedance network is realized with voltage controlled capacitors. Such an implementation may be adjusted, tuned or controlled by electrical circuits including programmable or controllable voltage sources and/or computer processors. The voltage controlled capacitors may be adjusted in response to data measured by the port parameter measurement circuitry and processed by measurement analysis and control algorithms and hardware. The voltage controlled capacitors may be a switched bank of capacitors.

[0070] Fig. 43 shows an end-to-end wireless power transmission system. In this example, both the source and the device contain port measurement circuitry and a processor. The box labeled “coupler/switch” indicates that the port measurement circuitry may be connected to the resonator by a directional coupler or a switch, enabling the measurement, adjustment and control of the source and device resonators to take place in conjunction with, or separate from, the power transfer functionality.

[0071] Fig. 44 shows an end-to-end wireless power transmission system. In this example, only the source contains port measurement circuitry and a processor. In this case, the device resonator operating characteristics may be fixed or may be adjusted by analog control circuitry and without the need for control signals generated by a processor.

[0072] Fig. 45 shows an end-to-end wireless power transmission system. In this example, both the source and the device contain port measurement circuitry but only the source contains a processor. Data from the device is transmitted through a wireless communication channel, which could be implemented either with a separate antenna, or through some modulation of the source drive signal.

[0073] Fig. 46 shows an end-to-end wireless power transmission system. In this example, only the source contains port measurement circuitry and a processor. Data from the device is transmitted through a wireless communication channel, which could be implemented either with a separate antenna, or through some modulation of the source drive signal.

[0074] Fig. 47 shows coupled magnetic resonators whose frequency and impedance may be automatically adjusted using algorithms implemented using a processor or a computer.

[0075] Fig. 48 shows a varactor array.

[0076] Fig. 49 shows a device (laptop computer) being wirelessly powered or charged by a source, where both the source and device resonator are physically separated from, but electrically connected to, the source and device.

[0077] Fig. 50 (a) is an illustration of a wirelessly powered or charged laptop application where the device resonator is inside the laptop case and is not visible.

[0078] Fig. 50 (b) is an illustration of a wirelessly powered or charged laptop application where the resonator is underneath the laptop base and is electrically connected to the laptop power input by an electrical cable.

[0079] Fig. 50 (c) is an illustration of a wirelessly powered or charged laptop application where the resonator is attached to the laptop base.

[0080] Fig. 50 (d) is an illustration of a wirelessly powered or charged laptop application where the resonator is attached to the laptop display.

[0081] Fig. 51 is a diagram of rooftop PV panels with wireless power transfer.

[0082] Fig. 52 a) is a diagram showing routing of individual traces in four layers of a layered PCB b) is a perspective three dimensional diagram showing routing of individual traces and via connections.

[0083] Fig. 53 a) is a diagram showing routing of individual traces in four layers of a layered PCB with one of the individual traces highlighted to show its path through the layer, b) is a perspective three dimensional diagram showing routing of conductor traces and via connection with one of the conductor traces highlighted to show its path through the layers for the stranded trace.

[0084] Fig. 54 is a diagram showing examples of alternative routing of individual traces.

[0085] Fig. 55 is a diagram showing routing of individual traces in one layer of a PCB.

[0086] Fig. 56 is a diagram showing routing direction between conducting layers of a PCB.

[0087] Fig. 57 is a diagram showing sharing of via space of two stranded traces routed next to each other.

[0088] Fig. 58(a)-(d) are diagrams of cross sections of stranded traces with various feature sizes and aspect ratios.

[0089] Fig. 59(a) is a plot of wireless power transfer efficiency between a fixed size device resonator and different sized source resonators as a function of separation distance and (b) is a diagram of the resonator configuration used for generating the plot.

[0090] Fig. 60(a) is a plot of wireless power transfer efficiency between a fixed size device resonator and different sized source resonators as a function of lateral offset and (b) is a diagram of the resonator configuration used for generating the plot.

[0091] Fig. 61 is a diagram of a conductor arrangement of an exemplary system embodiment.

[0092] Fig. 62 is a diagram of another conductor arrangement of an exemplary system embodiment.

[0093] Fig. 63 is a diagram of an exemplary system embodiment of a source comprising an array of equally sized resonators.

[0094] Fig. 64 is a diagram of an exemplary system embodiment of a source comprising an array of multi-sized resonators.

[0095] Fig. 65 is a diagram of an exemplary embodiment of an adjustable size source comprising planar resonator structures.

[0096] Fig. 66(a)-(d) are diagrams showing usage scenarios for an adjustable source size.

[0097] Fig. 67(a-b) is a diagram showing resonators with different keep out zones.

[0098] Fig. 68 is a diagram showing a resonator with a symmetric keep out zone.

[0099] Fig. 69 is a diagram showing a resonator with an asymmetric keep out zone.

[00100] Fig. 70 is a diagram showing an application of wireless power transfer.

[00101] Fig. 71(a-b) is a diagram arrays of resonators used to reduce lateral and angular alignment dependence between the source and device.

[00102] Fig. 72 is a plot showing the effect of resonator orientation on efficiency due to resonator displacement.

[00103] Fig. 73(a-b) is a diagram showing lateral and angular misalignments between resonators.

DETAILED DESCRIPTION

[00104] As described above, this disclosure relates to coupled electromagnetic resonators with long-lived oscillatory resonant modes that may wirelessly transfer power from a power supply to a power drain. However, the technique is not restricted to electromagnetic resonators, but is general and may be applied to a wide variety of resonators and resonant objects. Therefore, we first describe the general technique, and then disclose electromagnetic examples for wireless energy transfer.

[00105] Resonators

[00106] A resonator may be defined as a system that can store energy in at least two different forms, and where the stored energy is oscillating between the two forms. The resonance has a specific oscillation mode with a resonant (modal) frequency, f, and a resonant (modal) field. The angular resonant frequency, ω, may be defined as $\omega=2\pi f$, the resonant wavelength, λ, may be defined as $\lambda = c/f$, where c is the speed of light, and the resonant period, T, may be defined as $T=1/f=2\pi/\omega$. In the absence of loss mechanisms, coupling mechanisms or external energy supplying or draining mechanisms, the total resonator stored energy, W, would stay fixed and the two forms of energy would oscillate, wherein one would be maximum when the other is minimum and vice versa.

[00107] In the absence of extraneous materials or objects, the energy in the resonator 102 shown in Fig. 1 may decay or be lost by intrinsic losses. The resonator fields then obey the following linear equation:

$$\frac{da(t)}{dt} = -i\left(\omega - i\Gamma\right)a(t),$$

where the variable $a(t)$ is the resonant field amplitude, defined so that the energy contained within the resonator is given by $|a(t)|^2$. Γ is the intrinsic energy decay or loss rate (e.g. due to absorption and radiation losses).

[00108] The Quality Factor, or Q-factor, or Q, of the resonator, which characterizes the energy decay, is inversely proportional to these energy losses. It may be defined as $Q=\omega*W/P$, where P is the time-averaged power lost at steady state. That is, a resonator 102 with a high-Q has relatively low intrinsic losses and can store energy for a relatively long time. Since the resonator loses energy at its intrinsic decay rate, 2Γ, its Q, also referred to as its intrinsic Q, is given by $Q=\omega/2\Gamma$. The quality factor also represents the number of oscillation periods, T, it takes for the energy in the resonator to decay by a factor of e.

[00109] As described above, we define the quality factor or Q of the resonator as that due only to intrinsic loss mechanisms. A subscript index such as Q_1, indicates the resonator (resonator 1 in this case) to which the Q refers. Fig. 2 shows an electromagnetic resonator 102 labeled according to this convention. Note that in this figure, there are no extraneous objects or additional resonators in the vicinity of resonator 1.

[00110] Extraneous objects and/or additional resonators in the vicinity of a first resonator may perturb or load the first resonator, thereby perturbing or loading the Q of the first resonator, depending on a variety of factors such as the distance between the resonator and object or other resonator, the material composition of the object or other resonator, the structure of the first resonator, the power in the first resonator, and the like. Unintended external energy losses or coupling mechanisms to extraneous materials and objects in the vicinity of the resonators may be referred to as "perturbing" the Q of a resonator, and may be indicated by a subscript within rounded parentheses, (). Intended external energy losses, associated with energy transfer via coupling to other resonators and to generators and loads in the wireless energy transfer system may be referred to as "loading" the Q of the resonator, and may be indicated by a subscript within square brackets, [].

[00111] The Q of a resonator 102 connected or coupled to a power generator, g, or load 302 , l, may be called the "loaded quality factor" or the "loaded Q" and may be denoted by $Q_{[g]}$ or $Q_{[l]}$, as illustrated in Fig. 3. In general, there may be more than one generator or load 302 connected to a resonator 102. However, we do not list those generators or loads separately but rather use "*g*" and "*l*" to refer to the equivalent circuit loading imposed by the combinations of generators and loads. In general descriptions, we may use the subscript "*l*" to refer to either generators or loads connected to the resonators.

[00112] In some of the discussion herein, we define the "loading quality factor" or the "loading *Q*" due to a power generator or load connected to the resonator, as $\delta Q_{[l]}$, where, $1/\delta Q_{[l]} \equiv 1/Q_{[l]} - 1/Q$. Note that the larger the loading *Q*, $\delta Q_{[l]}$, of a generator or load, the less the loaded *Q*, $Q_{[l]}$, deviates from the unloaded *Q* of the resonator.

[00113] The *Q* of a resonator in the presence of an extraneous object 402, *p,* that is not intended to be part of the energy transfer system may be called the "perturbed quality factor" or the "perturbed *Q*" and may be denoted by $Q_{(p)}$, as illustrated in Fig. 4. In general, there may be many extraneous objects, denoted as *p1*, *p2*, etc., or a set of extraneous objects *{p}*, that perturb the *Q* of the resonator 102. In this case, the perturbed *Q* may be denoted $Q_{(p1+p2+...)}$ or $Q_{(\{p\})}$. For example, $Q_{1(brick+wood)}$ may denote the perturbed quality factor of a first resonator in a system for wireless power exchange in the presence of a brick and a piece of wood, and $Q_{2(\{office\})}$ may denote the perturbed quality factor of a second resonator in a system for wireless power exchange in an office environment.

[00114] In some of the discussion herein, we define the "perturbing quality factor" or the "perturbing *Q*" due to an extraneous object, *p*, as $\delta Q_{(p)}$, where $1/\delta Q_{(p)} \equiv 1/Q_{(p)} - 1/Q$. As stated before, the perturbing quality factor may be due to multiple extraneous objects, *p1*, *p2*, etc. or a set of extraneous objects, *{p}*. The larger the perturbing *Q*, $\delta Q_{(p)}$, of an object, the less the perturbed *Q*, $Q_{(p)}$, deviates from the unperturbed *Q* of the resonator.

[00115] In some of the discussion herein, we also define $\Theta_{(p)} \equiv Q_{(p)} / Q$ and call it the "quality factor insensitivity" or the "Q-insensitivity" of the resonator in the presence of an extraneous object. A subscript index, such as $\Theta_{1(p)}$, indicates the resonator to which the perturbed and unperturbed quality factors are referring, namely, $\Theta_{1(p)} \equiv Q_{1(p)} / Q_1$.

[00116] Note that the quality factor, *Q,* may also be characterized as "unperturbed", when necessary to distinguish it from the perturbed quality factor, $Q_{(p)}$, and "unloaded", when necessary to distinguish it from the loaded quality factor, $Q_{[l]}$. Similarly, the perturbed quality factor, $Q_{(p)}$, may also be characterized as "unloaded", when necessary to distinguish them from the loaded perturbed quality factor, $Q_{(p)[l]}$.

[00117] Coupled Resonators

[00118] Resonators having substantially the same resonant frequency, coupled through any portion of their near-fields may interact and exchange energy. There are a variety of physical pictures and models that may be employed to understand, design, optimize and characterize this energy exchange. One way to describe and model the energy exchange between two coupled resonators is using coupled mode theory (CMT).

[00119] In coupled mode theory, the resonator fields obey the following set of linear equations:

$$\frac{da_m(t)}{dt} = -i(\omega_m - i\Gamma_m)a_m(t) + i\sum_{n\neq m}\kappa_{mn}a_n(t)$$

where the indices denote different resonators and and κ_{mn} are the coupling coefficients between the resonators. For a reciprocal system, the coupling coefficients may obey the relation $\kappa_{mn} = \kappa_{nm}$. Note that, for the purposes of the present specification, far-field radiation interference effects will be ignored and thus the coupling coefficients will be considered real. Furthermore, since in all subsequent calculations of system performance in this specification the coupling coefficients appear only with their square, κ_{mn}^2, we use κ_{mn} to denote the absolute value of the real coupling coefficients.

[00120] Note that the coupling coefficient, κ_{mn}, from the CMT described above is related to the so-called coupling factor, k_{mn}, between resonators m and n by $k_{mn} = 2\kappa_{mn}/\sqrt{\omega_m\omega_n}$. We define a "strong-coupling factor", U_{mn}, as the ratio of the coupling and loss rates between resonators m and n, by $U_{mn} = \kappa_{mn}/\sqrt{\Gamma_m\Gamma_n} = k_{mn}\sqrt{Q_mQ_n}$.

[00121] The quality factor of a resonator *m,* in the presence of a similar frequency resonator *n* or additional resonators, may be loaded by that resonator *n* or additional resonators, in a fashion similar to the resonator being loaded by a connected power generating or consuming device. The fact that resonator *m* may be loaded by resonator *n* and vice versa is simply a different way to see that the resonators are coupled.

[00122] The loaded *Q*'s of the resonators in these cases may be denoted as $Q_{m[n]}$ and $Q_{n[m]}$. For multiple resonators or loading supplies or devices, the total loading of a resonator may be determined by modeling each load as a resistive loss, and adding the multiple loads in the appropriate parallel and/or series combination to determine the equivalent load of the ensemble.

[00123] In some of the discussion herein, we define the "loading quality factor" or the "loading Q_m" of resonator *m* due to resonator *n* as $\delta Q_{m[n]}$, where $1/\delta Q_{m[n]} \equiv 1/Q_{m[n]} - 1/Q_m$. Note that resonator *n* is also loaded by resonator *m* and its "loading Q_n" is given by $1/\delta Q_{n[m]} \equiv 1/Q_{n[m]} - 1/Q_n$.

[00124] When one or more of the resonators are connected to power generators or loads, the set of linear equations is modified to:

$$\frac{da_m(t)}{dt} = -i(\omega_m - i\Gamma_m)a_m(t) + i\sum_{n\neq m}\kappa_{mn}a_n(t) - \kappa_m a_m(t) + \sqrt{2\kappa_m}s_{+m}(t)$$
$$s_{-m}(t) = \sqrt{2\kappa_m}a_m(t) - s_{+m}(t),$$

where $s_{+m}(t)$ and $s_{-m}(t)$ are respectively the amplitudes of the fields coming from a generator into the resonator *m* and going out of the resonator *m* either back towards the generator or into a load, defined so that the power they carry is given by $|s_{+m}(t)|^2$ and $|s_{-m}(t)|^2$. The loading coefficients κ_m relate to the rate at which energy is exchanged between the resonator *m* and the generator or load connected to it.

[00125] Note that the loading coefficient, κ_m, from the CMT described above is related to the loading quality factor, $\delta Q_{m[l]}$, defined earlier, by $\delta Q_{m[l]} = \omega_m/2\kappa_m$.

[00126] We define a "strong-loading factor", $U_{m[l]}$, as the ratio of the loading and loss rates of resonator *m*, $U_{m[l]} = \kappa_m/\Gamma_m = Q_m/\delta Q_{m[l]}$.

[00127] Fig. 1(a) shows an example of two coupled resonators 1000, a first resonator 102S, configured as a source resonator and a second resonator 102D, configured as a device resonator. Energy may be transferred over a distance *D* between the resonators. The source resonator 102S may be driven by a power supply or generator (not shown). Work may be extracted from the device resonator 102D by a power consuming drain or load (e.g. a load resistor, not shown). Let us use the subscripts "s" for the source, "d" for the device, "g" for the generator, and "l" for the load, and, since in this example there are only two resonators and $\kappa_{sd}=\kappa_{ds}$, let us drop the indices on κ_{sd}, k_{sd}, and U_{sd}, and denote them as κ, k, and U, respectively.

[00128] The power generator may be constantly driving the source resonator at a constant driving frequency, f, corresponding to an angular driving frequency, ω, where $\omega=2\pi f$.

[00129] In this case, the efficiency, $\eta=|s_{-d}|^2/|s_{+s}|^2$, of the power transmission from the generator to the load (via the source and device resonators) is maximized under the following conditions: The source resonant frequency, the device resonant frequency and the generator driving frequency have to be matched, namely

$$\omega_s=\omega_d=\omega.$$

Furthermore, the loading Q of the source resonator due to the generator, $\delta Q_{s[g]}$, has to be matched (equal) to the loaded Q of the source resonator due to the device resonator and the load, $Q_{s[dl]}$, and inversely the loading Q of the device resonator due to the load, $\delta Q_{d[l]}$, has to be matched (equal) to the loaded Q of the device resonator due to the source resonator and the generator, $Q_{d[sg]}$, namely

$$\delta Q_{s[g]}=Q_{s[dl]} \text{ and } \delta Q_{d[l]}=Q_{d[sg]}.$$

These equations determine the optimal loading rates of the source resonator by the generator and of the device resonator by the load as

$$U_{d[l]}=\kappa_d/\Gamma_d=Q_d/\delta Q_{d[l]}=\sqrt{1+U^2}=\sqrt{1+\left(\kappa/\sqrt{\Gamma_s\Gamma_d}\right)^2}=Q_s/\delta Q_{s[g]}=\kappa_s/\Gamma_s=U_{s[g]}.$$

Note that the above frequency matching and Q matching conditions are together known as "impedance matching" in electrical engineering.

[00130] Under the above conditions, the maximized efficiency is a monotonically increasing function of only the strong-coupling factor, $U=\kappa/\sqrt{\Gamma_s\Gamma_d}=k\sqrt{Q_sQ_d}$, between the source and device resonators and is given by, $\eta=U^2/\left(1+\sqrt{1+U^2}\right)^2$, as shown in Fig. 5. Note that the coupling efficiency, η, is greater than 1% when U is greater than 0.2, is greater than 10% when U is greater than 0.7, is greater than 17% when U is greater than 1, is greater than 52% when U is greater than 3, is greater than 80% when U is greater than 9, is greater than 90% when U is greater than 19, and is greater than 95% when U is greater than 45. In some applications, the regime of operation where $U>1$ may be referred to as the "strong-coupling" regime.

[00131] Since a large $U=\kappa/\sqrt{\Gamma_s\Gamma_d}=\left(2\kappa/\sqrt{\omega_s\omega_d}\right)\sqrt{Q_sQ_d}$ is desired in certain circumstances, resonators may be used that are high-Q. The Q of each resonator may be high. The geometric mean of the resonator Q's, $\sqrt{Q_sQ_d}$ may also or instead be high.

[00132] The coupling factor, k, is a number between $0\le k\le 1$, and it may be independent (or nearly independent) of the resonant frequencies of the source and device resonators, rather it may determined mostly by their relative geometry and the physical decay-law of the field mediating their coupling. In contrast, the coupling coefficient, $\kappa=k\sqrt{\omega_s\omega_d}/2$, may be a strong function of the resonant frequencies. The resonant frequencies of the resonators may be chosen preferably to achieve a high Q rather than to achieve a low Γ, as these two goals may be achievable at two separate resonant frequency regimes.

[00133] A high-Q resonator may be defined as one with Q>100. Two coupled resonators may be referred to as a system of high-Q resonators when each resonator has a Q greater than 100, Q_s>100 and Q_d>100. In other implementationss, two coupled resonators may be referred to as a system of high-Q resonators when the geometric mean of the resonator *Q's* is greater than 100, $\sqrt{Q_sQ_d}>100$.

[00134] The resonators may be named or numbered. They may be referred to as source resonators, device resonators, first resonators, second resonators, repeater resonators, and the like. It is to be understood that while two resonators are shown in Fig. 1, and in many of the examples below, other implementations may include three (3) or more resonators. For example, a single source resonator 102S may transfer energy to multiple device resonators 102D or multiple devices. Energy may be transferred from a first device to a second, and then from the second device to the third, and so forth. Multiple sources may transfer energy to a single device or to multiple devices connected to a single device resonator or to multiple devices connected to multiple device resonators. Resonators 102 may serve alternately or simultaneously as sources, devices, or they may be used to relay power from a source in one location to a device in another location. Intermediate electromagnetic resonators 102 may be used to extend the distance range of wireless energy transfer systems. Multiple resonators 102 may be daisy chained together, exchanging energy over extended distances and with a wide range of sources and devices. High power levels may be split between multiple sources 102S, transferred to multiple devices and recombined at a distant location.

[00135] The analysis of a single source and a single device resonator may be extended to multiple source resonators and/or multiple device resonators and/or multiple intermediate resonators. In such an analysis, the conclusion may be that large strong-coupling factors, U_{mn}, between at least some or all of the multiple resonators is preferred for a high system efficiency in the wireless energy transfer. Again, implementations may use source, device and intermediate resonators that have a high *Q*. The *Q* of each resonator may be high. The geometric mean $\sqrt{Q_m Q_n}$ of the *Q*'s for pairs of resonators *m* and *n*, for which a large U_{mn} is desired, may also or instead be high.

[00136] Note that since the strong-coupling factor of two resonators may be determined by the relative magnitudes of the loss mechanisms of each resonator and the coupling mechanism between the two resonators, the strength of any or all of these mechanisms may be perturbed in the presence of extraneous objects in the vicinity of the resonators as described above.

[00137] Continuing the conventions for labeling from the previous sections, we describe *k* as the coupling factor in the absence of extraneous objects or materials. We denote the coupling factor in the presence of an extraneous object, *p*, as $k_{(p)}$, and call it the "perturbed coupling factor" or the "perturbed *k*". Note that the coupling factor, *k*, may also be characterized as "unperturbed", when necessary to distinguish from the perturbed coupling factor $k_{(p)}$.

[00138] We define $\delta k_{(p)} \equiv k_{(p)} - k$ and we call it the "perturbation on the coupling factor" or the "perturbation on *k*" due to an extraneous object, *p*.

[00139] We also define $\beta_{(p)} \equiv k_{(p)}/k$ and we call it the "coupling factor insensitivity" or the "k-insensitivity". Lower indices, such as $\beta_{12(p)}$, indicate the resonators to which the perturbed and unperturbed coupling factor is referred to, namely $\beta_{12(p)} \equiv k_{12(p)}/k_{12}$.

[00140] Similarly, we describe *U* as the strong-coupling factor in the absence of extraneous objects. We denote the strong-coupling factor in the presence of an extraneous object, *p*, as $U_{(p)}$, $U_{(p)} = k_{(p)}\sqrt{Q_{1(p)}Q_{2(p)}}$, and call it the "perturbed strong-coupling factor" or the "perturbed *U*". Note that the strong-coupling factor *U* may also be characterized as "unperturbed", when necessary to distinguish from the perturbed strong-coupling factor $U_{(p)}$.

Note that the strong-coupling factor U may also be characterized as "unperturbed", when necessary to distinguish from the perturbed strong-coupling factor $U_{(p)}$.

[00141] We define $\delta U_{(p)} \equiv U_{(p)} - U$ and call it the "perturbation on the strong-coupling factor" or the "perturbation on U" due to an extraneous object, p.

[00142] We also define $\Xi_{(p)} \equiv U_{(p)}/U$ and call it the "strong-coupling factor insensitivity" or the "U-insensitivity". Lower indices, such as $\Xi_{12(p)}$, indicate the resonators to which the perturbed and unperturbed coupling factor refers, namely $\Xi_{12(p)} \equiv U_{12(p)}/U_{12}$.

[00143] The efficiency of the energy exchange in a perturbed system may be given by the same formula giving the efficiency of the unperturbed system, where all parameters such as strong-coupling factors, coupling factors, and quality factors are replaced by their perturbed equivalents. For example, in a system of wireless energy transfer including one source and one device resonator, the optimal efficiency may calculated as $\eta_{(p)} = \left[U_{(p)} \Big/ \left(1 + \sqrt{1 + U_{(p)}^{\ 2}} \right) \right]^2$. Therefore, in a system of wireless energy exchange which is perturbed by extraneous objects, large perturbed strong-coupling factors, $U_{mn(p)}$, between at least some or all of the multiple resonators may be desired for a high system efficiency in the wireless energy transfer. Source, device and/or intermediate resonators may have a high $Q_{(p)}$.

[00144] Some extraneous perturbations may sometimes be detrimental for the perturbed strong-coupling factors (via large perturbations on the coupling factors or the quality factors). Therefore, techniques may be used to reduce the effect of extraneous perturbations on the system and preserve large strong-coupling factor insensitivites.

[00145] **Efficiency of Energy Exchange**

[00146] The so-called "useful" energy in a useful energy exchange is the energy or power that must be delivered to a device (or devices) in order to power or charge the device. The transfer efficiency that corresponds to a useful energy exchange may be system or application dependent. For example, high power vehicle charging applications that transfer kilowatts of power may need to be at least 80% efficient in order to supply useful amounts of power resulting in a useful energy exchange sufficient to recharge a vehicle battery, without significantly heating up various components of the transfer system. In some consumer electronics applications, a useful energy exchange may include any energy transfer efficiencies greater than 10%, or any other amount acceptable to keep rechargeable batteries "topped off" and running for long periods of time. For some wireless sensor applications, transfer efficiencies that are much less than 1% may be adequate for powering multiple low power sensors from a single source located a significant distance from the sensors. For still other applications, where wired power transfer is either impossible or impractical, a wide range of transfer efficiencies may be acceptable for a useful energy exchange and may be said to supply useful power to devices in those applications. In general, an operating distance is any distance over which a useful energy exchange is or can be maintained according to the principles disclosed herein.

[00147] A useful energy exchange for a wireless energy transfer in a powering or recharging application may be efficient, highly efficient, or efficient enough, as long as the wasted energy levels, heat dissipation, and associated field strengths are within tolerable limits. The tolerable limits may depend on the application, the environment and the system location. Wireless energy transfer for powering or recharging applications may be efficient, highly efficient, or efficient enough, as long as the desired system performance may be attained for the reasonable cost restrictions, weight restrictions, size restrictions, and the like. Efficient energy transfer may be determined relative to that which could be achieved using traditional inductive techniques that are not high-Q systems. Then, the energy transfer may be defined as being efficient, highly efficient, or efficient enough, if more energy is delivered than could be delivered by similarly sized coil structures in traditional inductive schemes over similar distances or alignment offsets.

[00148] Note that, even though certain frequency and Q matching conditions may optimize the system efficiency of energy transfer, these conditions may not need to be exactly met in order to have efficient enough energy transfer for a useful energy exchange. Efficient energy exchange may be realized so long as the relative offset of the resonant frequencies $(|\omega_m - \omega_n|/\sqrt{\omega_m \omega_n})$ is less than approximately the maximum among $1/Q_{m(p)}$, $1/Q_{n(p)}$ and $k_{mn(p)}$. The Q matching condition may be less critical than the frequency matching condition for efficient energy exchange. The degree by which the strong-loading factors, $U_{m[l]}$, of the resonators due to generators and/or loads may be away from their optimal values and still have efficient enough energy exchange depends on the particular system, whether all or some of the generators and/or loads are Q-mismatched and so on.

[00149] Therefore, the resonant frequencies of the resonators may not be exactly matched, but may be matched within the above tolerances. The strong-loading factors of at least some of the resonators due to generators and/or loads may not be exactly matched to their optimal value. The voltage levels, current levels, impedance values, material parameters, and the like may not be at the exact values described in the disclosure but will be within some acceptable tolerance of those values. The system optimization may include cost, size, weight, complexity, and the like, considerations, in addition to efficiency, Q, frequency, strong coupling factor, and the like, considerations. Some system performance parameters, specifications, and designs may be far from optimal in order to optimize other system performance parameters, specifications and designs.

[00150] In some applications, at least some of the system parameters may be varying in time, for example because components, such as sources or devices, may be mobile or aging or because the loads may be variable or because the perturbations or the environmental conditions are changing etc. In these cases, in order to achieve acceptable matching conditions, at least some of the system parameters may need to be dynamically adjustable or tunable. All the system parameters may be dynamically adjustable or tunable to achieve approximately the optimal operating conditions. However, based on the discussion above, efficient enough energy exchange may be realized even if some system parameters are not variable. In some examples, at least some of the devices may not be dynamically adjusted. In some examples, at least some of the sources may not be dynamically adjusted. In some examples, at least some of the intermediate resonators may not be dynamically adjusted. In some examples, none of the system parameters may be dynamically adjusted.

[00151] **Electromagnetic Resonators**

[00152] The resonators used to exchange energy may be electromagnetic resonators. In such resonators, the intrinsic energy decay rates, Γ_m, are given by the absorption (or resistive) losses and the radiation losses of the resonator.

[00153] The resonator may be constructed such that the energy stored by the electric field is primarily confined within the structure and that the energy stored by the magnetic field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant magnetic near-field. These types of resonators may be referred to as magnetic resonators.

[00154] The resonator may be constructed such that the energy stored by the magnetic field is primarily confined within the structure and that the energy stored by the electric field is primarily in the region surrounding the resonator. Then, the energy exchange is mediated primarily by the resonant electric near-field. These types of resonators may be referred to as electric resonators.

[00155] Note that the total electric and magnetic energies stored by the resonator have to be equal, but their localizations may be quite different. In some cases, the ratio of the average electric field energy to the average magnetic field energy specified at a distance from a resonator may be used to characterize or describe the resonator.

[00156] Electromagnetic resonators may include an inductive element, a distributed inductance, or a combination of inductances with inductance, *L*, and a capacitive element, a distributed capacitance, or a combination of capacitances, with capacitance, *C*. A minimal circuit model of an electromagnetic resonator 102 is shown in Fig. 6a. The resonator may include an inductive element 108 and a capacitive element 104. Provided with initial energy, such as electric field energy stored in the capacitor 104, the system will oscillate as the capacitor discharges transferring energy into magnetic field energy stored in the inductor 108 which in turn transfers energy back into electric field energy stored in the capacitor 104.

[00157] The resonators 102 shown in Figs. 6(b)(c)(d) may be referred to as magnetic resonators. Magnetic resonators may be preferred for wireless energy transfer applications in populated environments because most everyday materials including animals, plants, and humans are non-magnetic (i.e., $\mu_r \approx 1$), so their interaction with magnetic fields is minimal and due primarily to eddy currents induced by the time-variation of the magnetic fields, which is a second-order effect. This characteristic is important both for safety reasons and because it reduces the potential for interactions with extraneous environmental objects and materials that could alter system performance.

[00158] Fig. 6d shows a simplified drawing of some of the electric and magnetic field lines associated with an exemplary magnetic resonator 102B. The magnetic resonator 102B may include a loop of conductor acting as an inductive element 108 and a capacitive element 104 at the ends of the conductor loop. Note that this drawing depicts most of the energy in the region surrounding the resonator being stored in the magnetic field, and most of the energy in the resonator (between the capacitor plates) stored in the electric field. Some electric field, owing to fringing fields, free charges, and the time varying magnetic field, may be stored in the region around the resonator, but the magnetic resonator may be designed to confine the electric fields to be close to or within the resonator itself, as much as possible.

[00159] The inductor 108 and capacitor 104 of an electromagnetic resonator 102 may be bulk circuit elements, or the inductance and capacitance may be distributed and may result from the way the conductors are formed, shaped, or positioned, in the structure. For example, the inductor 108 may be realized by shaping a conductor to enclose a surface area, as shown in Figs. 6(b)(c)(d). This type of resonator 102 may be referred to as a capacitively-loaded loop inductor. Note that we may use the terms “loop” or “coil” to indicate generally a conducting structure (wire, tube, strip, etc.), enclosing a surface of any shape and dimension, with any number of turns. In Fig. 6b, the enclosed surface area is circular, but the surface may be any of a wide variety of other shapes and sizes and may be designed to achieve certain system performance specifications. As an example to indicate how inductance scales with physical dimensions, the inductance for a length of circular conductor arranged to form a circular single-turn loop is approximately,

$$L = \mu_0 x(\ln\frac{8x}{a} - 2),$$

where μ_0 is the magnetic permeability of free space, x, is the radius of the enclosed circular surface area and, a, is the radius of the conductor used to form the inductor loop. A more precise value of the inductance of the loop may be calculated analytically or numerically.

[00160] The inductance for other cross-section conductors, arranged to form other enclosed surface shapes, areas, sizes, and the like, and of any number of wire turns, may be calculated analytically, numerically or it may be determined by measurement. The inductance may be realized using inductor elements, distributed inductance, networks, arrays, series and parallel combinations of inductors and inductances, and the like. The inductance may be fixed or variable and may be used to vary impedance matching as well as resonant frequency operating conditions.

[00161] There are a variety of ways to realize the capacitance required to achieve the desired resonant frequency for a resonator structure. Capacitor plates 110 may be formed and utilized as shown in Fig. 6b, or the capacitance may be distributed and be realized between adjacent windings of a multi-loop conductor 114, as shown in Fig. 6c. The capacitance may be