IN THE UNITED STATES DISTRICT COURT
FOR THE DISTRICT OF DELAWARE

FRAUNHOFER-GESELLSCHAFT ZUR FÖRDERUNG DER ANGEWANDTEN FORSCHUNG E.V.,	)	Case No. _________________
Plaintiff,	)	**DEMAND FOR JURY TRIAL**
vs.	)	
SIRIUS XM RADIO INC.,	)	
Defendant.	)	

COMPLAINT

Plaintiff Fraunhofer-Gesellschaft Zur Förderung der angewandten Forschung e.V. ("Fraunhofer") hereby pleads the following claims for relief against Defendant Sirius XM Radio Inc. ("Sirius XM") and alleges as follows:

NATURE OF THE ACTION

1. Fraunhofer is one of the largest and most successful applied research organizations in Europe and the world. Founded in 1946, Fraunhofer now includes over 60 institutes and research units dedicated to developing real-world innovations in the fields of health, communications, security, transportation, and energy, both for private industry contracts and publicly funded projects. One of Fraunhofer's most famous inventions is the MP3, a compressed audio recording format that has become the standard for digital audio around the world. Fraunhofer has been named to the Thomson Reuters Top 100 Global Innovators list three times in the five years since the list's inception.

2. In the late 1990s, Fraunhofer's institutes began developing technology for a Digital Audio Radio Service ("DARS"), now more commonly known as satellite radio. Satellite radio has transformed the way that consumers experience radio, both in their homes and while traveling. Before satellite radio, radio listeners were restricted by the

geographical limitations of traditional AM and FM radio. Most traditional radio signals extend only 30 or 40 miles from their source, and radio listeners were forced to change stations as they traveled in and out of range.

3. Satellite radio eliminates these geographic restrictions. With satellite radio, instead of broadcasting directly from a ground station, a radio broadcaster transmits its signal more than 22,000 miles away to a satellite, which then transmits the signal back to radio receivers installed in cars and homes nationwide.

4. Satellite radio operates by transmitting a signal from a satellite orbiting the planet to terrestrial radio receivers or ground repeaters. However, because the satellite and the ground receivers and repeaters are moving at different relative speeds, the signal can become disrupted in a process known as "channel fading." To reduce the effect of channel fading—and thereby ensure the clarity of satellite broadcasts—Fraunhofer developed certain multicarrier modulation ("MCM") technologies for use in satellite radio broadcasting. MCM is a method of transmitting data by splitting it into several components and sending each of the components over separate carrier signals. Fraunhofer's novel inventions contributed to the development of a satellite radio system that permit listeners to access hundreds of music, news, and entertainment broadcasts anywhere in the United States, with absolute clarity.

5. Fraunhofer has sought to protect its hard work and ingenuity by seeking and obtaining extensive intellectual property protection for its innovations. As of the end of 2015, Fraunhofer held a portfolio of more than 6,500 patent families. This portfolio includes more than 1,500 issued United States patents on Fraunhofer's inventions, including more than 50 directly related to Fraunhofer's work on satellite communications, such as

patented inventions relating to signal transmission, data coding, and receiver processing, among others.

6. Sirius XM is the only satellite radio provider in the United States. It broadcasts hundreds of radio channels from studios in New York City, Washington, D.C., Nashville, Memphis, and Los Angeles to more than 30.6 million subscribers across the country. In addition to the subscription service, Sirius XM also sells and supplies satellite radio receiver units directly to consumers and to major auto manufacturers for pre-installation into new vehicles.

7. Sirius XM is capitalizing on Fraunhofer's innovation and success by making, selling, and using satellite radios and subscription satellite radio services that infringe Fraunhofer's patents. Sirius XM is utilizing Fraunhofer's patented inventions without license or authority from Fraunhofer.

8. Fraunhofer has brought this action to remedy Sirius XM's infringement.

PARTIES

9. Plaintiff Fraunhofer is a non-profit corporation organized and existing under the laws of the Federal Republic of Germany, with its principal place of business at Postfach 20 07 33, 80007 Munich, Germany.

10. Defendant Sirius XM is a corporation organized and existing under the laws of Delaware with its principal place of business at 1221 Avenue of the Americas, New York, New York 10020. Upon information and belief, Sirius operates through an agent, The Corporation Trust Company, with its principal place of business at 1209 Orange Street, Wilmington, Delaware 19801.

JURISDICTION AND VENUE

11. This is an action for patent infringement arising under the patent laws of the United States of America, 35 U.S.C. § 1, *et seq.*, including 35 U.S.C. § 271. This Court has subject matter jurisdiction over the claims of patent infringement alleged in this Complaint under 28 U.S.C. §§ 1331 and 1338(a).

12. This Court has personal jurisdiction over Sirius XM because it is organized and existing under the laws of Delaware. Furthermore, Fraunhofer is informed and believes, and on that basis alleges, that Sirius has committed and continues to commit acts of infringement in Delaware in violation of 35 U.S.C. § 271 by, among other things, using, selling, and offering to sell infringing products and services in this judicial district.

13. Venue in the District of Delaware is proper pursuant to 28 U.S.C. §§ 1391(b) and (c) because, among other reasons, Sirius XM resides and is subject to personal jurisdiction in this judicial district. Furthermore, a substantial part of the events or omissions giving rise to claims alleged herein occurred in this judicial district. Sirius XM has committed acts of infringement in this judicial district by, among other things, marketing, selling, and offering for sale infringing products and services in this judicial district.

FACTUAL BACKGROUND

14. Fraunhofer's investment in research and development has resulted in more than 1,500 patents recognized by the United States Patent and Trademark Office. This case involves four of those patents.

15. Fraunhofer is the owner of the entire right, title, and interest in and to U.S. Patent No. 6,314,289 ("the '289 Patent"), entitled "Apparatus and Method for Transmitting Information and Apparatus and Method for Receiving Information," which was duly issued

on November 6, 2001. Fraunhofer holds the exclusive rights to bring suit with respect to any past, present, and future infringement of the ’289 Patent. A copy of the ’289 Patent is attached as Exhibit A hereto.

16. Fraunhofer is the owner of the entire right, title, and interest in and to U.S. Patent No. 6,931,084 (“the ’1084 Patent”), entitled “Differential Coding and Carrier Recovery for Multicarrier Systems,” which was duly issued on August 16, 2005. Fraunhofer holds the exclusive rights to bring suit with respect to any past, present, and future infringement of the ’1084 Patent. A copy of the ’1084 Patent is attached as Exhibit B hereto.

17. Fraunhofer is the owner of the entire right, title, and interest in and to U.S. Patent No. 6,993,084 (“the ’3084 Patent”), entitled “Coarse Frequency Synchronization in Multicarrier Systems,” which was duly issued on January 31, 2006. Fraunhofer holds the exclusive rights to bring suit with respect to any past, present, and future infringement of the ’3084 Patent. A copy of the ’3084 Patent is attached as Exhibit C hereto.

18. Fraunhofer is the owner of the entire right, title and interest in and to U.S. Patent No. 7,061,997 (“the ’997 Patent”), entitled “Method and Apparatus for Fine Frequency Synchronization in Multi-Carrier Demodulation Systems,” which was duly issued on June 13, 2006. Fraunhofer holds the exclusive rights to bring suit with respect to any past, present, and future infringement of the ’997 Patent. A copy of the ’997 Patent is attached as Exhibit D hereto.

19. The ’289, ’1084, ’3084, and ’997 Patents are directed to apparatuses and methods used to receive and decode encoded satellite signals, identify any “channel fading” effects, such as offsets in the phase shift, frequency, or amplitude of the signal waves, and

correct for those offsets. This correction process occurs using a channel decoder, which is a component of a satellite radio receiver.

20. Fraunhofer developed the patented MCM technologies under a "Frame Agreement" it entered with WorldSpace on July 3, 1996.

21. In 1998, Fraunhofer granted an exclusive right to WorldSpace International Network Inc. ("WorldSpace") to license all Fraunhofer intellectual property rights for MCM technologies suitable for use with digital satellite broadcasts, including patents to be sought relating to MCM. Fraunhofer subsequently applied for and obtained the '289, '1084, '3084, and '997 Patents, all of which relate to MCM technologies and are covered by this exclusive license to WorldSpace (the "MCM License").

22. WorldSpace in turn granted a sub-license under the MCM License to American Mobile Radio Corporation, which was later renamed as XM Satellite Radio, Inc. ("XM Satellite"). XM Satellite sought to use the sublicensed technology in connection with the development of XM Satellite's Digital Audio Radio Services System (the "XM DARS System").

23. In July 1999, XM Satellite entered into a Technical Consulting Contract with Fraunhofer to facilitate its development efforts (the "XM Radio Contract"). Under the terms of that contract, Fraunhofer was to engineer the XM DARS System for XM Satellite, and develop and test the equipment needed to generate, receive, and decode satellite and terrestrial signals. Although the parties acknowledged that certain "background intellectual property" owned by Fraunhofer would be used to complete the development of the DARS System, the contract specifically excluded the MCM technologies that Fraunhofer had exclusively licensed to WorldSpace, for which XM Satellite had already obtained a separate

sublicense from WorldSpace. A copy of the XM Radio Contract is attached as Exhibit E hereto.

24. To fulfill its obligations under the XM Radio Contract, Fraunhofer worked in close cooperation with STMicroelectronics (an agent of XM Satellite) to develop the channel decoder IC, a decoding device that performed the phase offset correction. The channel decoder IC was the initial basis for all XM receiver developments, and is a critical component of the XM DARS System. Since 2001, STMicroelectronics chips bearing the designations STA400 (Channel Decoder), STA850 (Channel and Service & Source Decoder), STA875 (Channel and Service & Source Decoder with EEPROM), and STA895 (system on a chip ("SOC") containing a Channel Decoder) have been used in XM Satellite's satellite radio receivers.

25. Fraunhofer specifically developed the channel decoder to implement the MCM methods and apparatuses patented by Fraunhofer and licensed to WorldSpace. Thus, Fraunhofer built the infringing aspects of the XM DARS System at the request of XM Satellite using the technologies covered by the '289, '1084, '3084, and '997 Patents.

26. In 2008, XM Satellite merged with a company named Sirius Satellite Radio, which had developed its own satellite radio system, to form Sirius XM.

27. On October 17, 2008, WorldSpace filed a voluntary petition for relief under chapter 11 of the Bankruptcy Code in the United States Bankruptcy Court for the District of Delaware. On June 1, 2010, the Bankruptcy Court approved an agreement between Fraunhofer, WorldSpace, and a potential buyer of WorldSpace's assets that unambiguously rejected the MCM License.[1] As a result, WorldSpace lost all rights relating to Fraunhofer's

[1] The agreement stated: "In the event that Fraunhofer and Yazmi [the buyer] have not entered into an agreement on the going forward business arrangements as provided for in

MCM technologies. By the same token, all WorldSpace sub-licensees, including Sirius XM, also lost any rights to the MCM technologies that had been conferred or sublicensed based on WorldSpace's License.

28. On June 12, 2012, WorldSpace's Chapter 11 case was converted to Chapter 7. Pursuant to 11 U.S.C. § 365(d), the MCM License was confirmed rejected as of August 12, 2012, when the bankruptcy trustee failed to assume either the MCM License or the sub-license to Sirius XM within sixty days after the conversion. Accordingly, Sirius XM's license terminated, at the latest, by August 12, 2012.

29. In operating the XM DARS System as well as making, using, and selling satellite radios as part of that system, Sirius XM continues to use the technologies covered by the '289, '1084, '3084, and '997 Patents for which Sirius XM no longer holds a license.

30. In October 2015, Fraunhofer informed Sirius XM that Sirius XM was infringing on Fraunhofer's valid patents and attempted to ascertain the basis upon which Sirius XM was continuing to use the patented MCM technologies. Sirius XM advised Fraunhofer that it believed that it had authority to continue using the patented MCM technologies. However, as shown above, any license or authority that Sirius XM may have had to the patented MCM technologies was terminated in connection with the bankruptcy proceedings involving WorldSpace.

31. Sirius XM offers the XM DARS System on a subscription basis to its more than 30.6 million customers. Fraunhofer is informed and believes, and on that basis alleges, that certain Sirius XM radio receivers, which are required to use the XM DARS System, contain a channel decoder IC. Thus, the XM DARS System uses Fraunhofer's MCM

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the agreement, the license agreement will be deemed rejected." Fraunhofer and Yazmi never entered into an agreement relating to the MCM License.

technology, including the ’289, ’1084, ’3084, and ’997 Patents, to compensate for channel fading effects, without license or authority from Fraunhofer.

32. Sirius XM also develops and supplies the equipment, including satellite radios, needed to use the XM DARS system. Sirius XM sells these satellite radios directly to consumers and businesses. Sirius XM also supplies satellite radios to the largest automakers in the world, including but not limited to Toyota, General Motors, Volkswagen, Nissan, and Hyundai. Sirius XM’s website identifies the full list of automakers that make Sirius XM satellite radios available for installation. *See* http://www.siriusxm.com/vehicleavailability. In 2015, Sirius XM radios were available in more than 75% of all new cars sold, and approximately one third of cars currently on the road in America are equipped with a factory-installed satellite radio. Thus, Sirius XM has caused its infringing satellite radios to be widely deployed without license or authority from Fraunhofer. These satellite radios include but are not limited to radios bearing the model designations XEZ1V1, SXPL1V1, SSV7V1, SST8V1, and GDI-SXBR1 and other radios having the same or similar functionality with respect to the ’289, ’1084, ’3084, and ’997 Patents (the “Sirius XM Satellite Radios”). The ability to access the XM DARS System on the Sirius XM Satellite Radios is an essential feature and function of these devices.

33. Fraunhofer is informed and believes that Sirius XM was well aware of Fraunhofer and had knowledge of Fraunhofer’s ’289, ’1084, ’3084, and ’997 Patents before the filing of this action.

34. For example, Sirius XM obtained knowledge of Fraunhofer’s patents through the contractual agreements between and among Fraunhofer, WorldSpace, and XM Satellite related to the licensing of Fraunhofer’s MCM technologies.

35. In addition, Sirius XM obtained knowledge of Fraunhofer's patents and its infringement through communications from Fraunhofer in 2015 informing Sirius XM of its infringement.

36. Sirius XM also gained knowledge of Fraunhofer, Fraunhofer's patents, and its infringement of the '289, '1084, '3084, and '997 Patents as a result of this Complaint.

FIRST CLAIM FOR RELIEF
Infringement of U.S. Patent No. 6,314,289

37. Fraunhofer realleges and incorporates by reference Paragraphs 1 through 36 of this Complaint, as if fully set forth herein.

38. Fraunhofer is the owner of the entire right, title, and interest in and to the '289 Patent.

39. Sirius XM has infringed and is currently infringing one or more claims of the '289 Patent, in violation of 35 U.S.C. § 271.

40. Sirius XM has infringed and continues to infringe the '289 Patent in violation of 35 U.S.C. § 271(a), literally and/or under the doctrine of equivalents by, among other things, making, using, offering for sale, selling, and/or importing within this judicial district and elsewhere in the United States, without license or authority, the XM DARS System and the Sirius XM Satellite Radios and related products and/or processes falling within the scope of one or more claims of the '289 Patent.

41. Sirius XM has infringed and continues to infringe the '289 Patent in violation of 35 U.S.C. § 271(b) by actively inducing infringement of the '289 Patent, literally and/or under the doctrine of equivalents, with knowledge of the '289 Patent and knowledge that it was inducing the infringement of the '289 Patent by, among other things, actively and knowingly aiding and abetting, assisting and encouraging others, including without limitation Toyota, General Motors, Volkswagen, Nissan, and Hyundai, and other

automakers, and other customers and end users of Sirius XM products and services, to directly infringe the ’289 Patent with respect to the making, using, offering for sale, selling, and/or importing within this judicial district and elsewhere in the United States, without license or authority, the XM DARS System and the Sirius XM Satellite Radios and related products and/or processes falling within the scope of one or more claims of the ’289 Patent. Moreover, Sirius XM had knowledge of its infringement of the Fraunhofer patents, as set forth in Paragraphs 32 through 36.

42. Sirius XM has infringed and continues to infringe the ’289 Patent in violation of 35 U.S.C. § 271(c) by contributing to infringement of the ’289 Patent, literally and/or under the doctrine of equivalents, by, among other things, selling, offering for sale, and/or importing within this judicial district and elsewhere in the United States, without license or authority, the XM DARS System and the Sirius XM Satellite Radios and related products and/or processes falling within the scope of one or more claims of the ’289 Patent, with knowledge of the ’289 Patent knowing that such products and/or components are especially made or especially adapted for use in the infringement of the ’289 Patent, and not staple articles or commodities of commerce suitable for substantial noninfringing use. Moreover, Sirius XM had knowledge of its infringement of the Fraunhofer patents, as set forth in Paragraphs 32 through 36.

43. Fraunhofer is informed and believes, and on that basis alleges, that Sirius XM’s infringement of the ’289 Patent has been and continues to be willful and deliberate, entitling Fraunhofer to increased damages under 35 U.S.C. § 284. Fraunhofer is informed and believes, and on that basis alleges, that Sirius XM had knowledge of the ’289 Patent and its infringement, and nonetheless engaged in objectively reckless conduct by continuing to

engage in infringing activity in the face of an objectively high risk that Sirius XM was infringing Fraunhofer's valid '289 Patent.

44. Sirius XM's acts of infringement have caused damage to Fraunhofer in an amount to be proven at trial. As a consequence of Sirius XM's infringement, Fraunhofer is entitled to recover damages adequate to compensate it for the infringement complained of herein, but in no event less than a reasonable royalty.

45. Fraunhofer has suffered irreparable injury as a direct and proximate result of Sirius XM's infringement for which there is no adequate remedy at law. Unless Sirius XM is enjoined, Fraunhofer will continue to suffer such irreparable injury as a direct and proximate result of Sirius XM's conduct.

SECOND CLAIM FOR RELIEF
Infringement of U.S. Patent No. 6,931,084

46. Fraunhofer realleges and incorporates by reference Paragraphs 1 through 45 of this Complaint, as if fully set forth herein.

47. Fraunhofer is the owner of the entire right, title, and interest in and to the '1084 Patent.

48. Sirius XM has infringed and is currently infringing one or more claims of the '1084 Patent, in violation of 35 U.S.C. § 271.

49. Sirius XM has infringed and continues to infringe the '1084 Patent in violation of 35 U.S.C. § 271(a), literally and/or under the doctrine of equivalents by, among other things, making, using, offering for sale, selling, and/or importing within this judicial district and elsewhere in the United States, without license or authority, the XM DARS System and the Sirius XM Satellite Radios and related products and/or processes falling within the scope of one or more claims of the '1084 Patent.

50. Sirius XM has infringed and continues to infringe the ’1084 Patent in violation of 35 U.S.C. § 271(b) by actively inducing infringement of the ’1084 Patent, literally and/or under the doctrine of equivalents, with knowledge of the ’1084 Patent and knowledge that it was inducing the infringement of the ’1084 Patent by, among other things, actively and knowingly aiding and abetting, assisting and encouraging others, including without limitation Toyota, General Motors, Volkswagen, Nissan, and Hyundai, and other automakers, and other customers and end users of Sirius XM products and services, to directly infringe the ’1084 Patent with respect to the making, using, offering for sale, selling, and/or importing within this judicial district and elsewhere in the United States, without license or authority, the XM DARS System and the Sirius XM Satellite Radios and related products and/or processes falling within the scope of one or more claims of the ’1084 Patent. Moreover, Sirius XM had knowledge of its infringement of the Fraunhofer patents, as set forth in Paragraphs 32 through 36.

51. Sirius XM has infringed and continues to infringe the ’1084 Patent in violation of 35 U.S.C. § 271(c) by contributing to infringement of the ’1084 Patent, literally and/or under the doctrine of equivalents, by, among other things, selling, offering for sale, and/or importing within this judicial district and elsewhere in the United States, without license or authority, the XM DARS System and the Sirius XM Satellite Radios and related products and/or processes falling within the scope of one or more claims of the ’1084 Patent, with knowledge of the ’1084 Patent knowing that such products and/or components are especially made or especially adapted for use in the infringement of the ’1084 Patent, and not staple articles or commodities of commerce suitable for substantial noninfringing use. Moreover, Sirius XM had knowledge of its infringement of the Fraunhofer patents, as set forth in Paragraphs 32 through 36.

52. Fraunhofer is informed and believes, and on that basis alleges, that Sirius XM's infringement of the '1084 Patent has been and continues to be willful and deliberate, entitling Fraunhofer to increased damages under 35 U.S.C. § 284. Fraunhofer is informed and believes, and on that basis alleges, that Sirius XM, had knowledge of the '1084 Patent and its infringement, and nonetheless engaged in objectively reckless conduct by continuing to engage in infringing activity in the face of an objectively high risk that Sirius XM was infringing Fraunhofer's valid '1084 Patent.

53. Sirius XM's acts of infringement have caused damage to Fraunhofer in an amount to be proven at trial. As a consequence of Sirius XM's infringement, Fraunhofer is entitled to recover damages adequate to compensate it for the infringement complained of herein, but in no event less than a reasonable royalty.

54. Fraunhofer has suffered irreparable injury as a direct and proximate result of Sirius XM's infringement for which there is no adequate remedy at law. Unless Sirius XM is enjoined, Fraunhofer will continue to suffer such irreparable injury as a direct and proximate result of Sirius XM's conduct.

THIRD CLAIM FOR RELIEF
Infringement of U.S. Patent No. 6,993,084

55. Fraunhofer realleges and incorporates by reference Paragraphs 1 through 54 of this Complaint, as if fully set forth herein.

56. Fraunhofer is the owner of the entire right, title, and interest in and to the '3084 Patent.

57. Sirius XM has infringed and is currently infringing one or more claims of the '3084 Patent, in violation of 35 U.S.C. § 271.

58. Sirius XM has infringed and continues to infringe the '3084 Patent in violation of 35 U.S.C. § 271(a), literally and/or under the doctrine of equivalents by, among

other things, making, using, offering for sale, selling, and/or importing within this judicial district and elsewhere in the United States, without license or authority, the XM DARS System and the Sirius XM Satellite Radios and related products and/or processes falling within the scope of one or more claims of the ’3084 Patent.

59. Sirius XM has infringed and continues to infringe the ’3084 Patent in violation of 35 U.S.C. § 271(b) by actively inducing infringement of the ’3084 Patent, literally and/or under the doctrine of equivalents, with knowledge of the ’3084 Patent and knowledge that it was inducing the infringement of the ’3084 Patent by, among other things, actively and knowingly aiding and abetting, assisting and encouraging others, including without limitation Toyota, General Motors, Volkswagen, Nissan, and Hyundai, and other automakers, and other customers and end users of Sirius XM products and services, to directly infringe the ’3084 Patent with respect to the making, using, offering for sale, selling, and/or importing within this judicial district and elsewhere in the United States, without license or authority, the XM DARS System and the Sirius XM Satellite Radios and related products and/or processes falling within the scope of one or more claims of the ’3084 Patent. Moreover, Sirius XM had knowledge of its infringement of the Fraunhofer patents, as set forth in Paragraphs 32 through 36.

60. Sirius XM has infringed and continues to infringe the ’3084 Patent in violation of 35 U.S.C. § 271(c) by contributing to infringement of the ’3084 Patent, literally and/or under the doctrine of equivalents, by, among other things, selling, offering for sale, and/or importing within this judicial district and elsewhere in the United States, without license or authority, the XM DARS System and the Sirius XM Satellite Radios and related products and/or processes falling within the scope of one or more claims of the ’3084 Patent, with knowledge of the ’3084 Patent knowing that such products and/or components are

especially made or especially adapted for use in the infringement of the ’3084 Patent, and not staple articles or commodities of commerce suitable for substantial noninfringing use. Moreover, Sirius XM had knowledge of its infringement of the Fraunhofer patents, as set forth in Paragraphs 32 through 36.

61. Fraunhofer is informed and believes, and on that basis alleges, that Sirius XM’s infringement of the ’3084 Patent has been and continues to be willful and deliberate, entitling Fraunhofer to increased damages under 35 U.S.C. § 284. Fraunhofer is informed and believes, and on that basis alleges, that Sirius XM, had knowledge of the ’3084 Patent and its infringement, and nonetheless engaged in objectively reckless conduct by continuing to engage in infringing activity in the face of an objectively high risk that Sirius XM was infringing Fraunhofer’s valid ’3084 Patent.

62. Sirius XM’s acts of infringement have caused damage to Fraunhofer in an amount to be proven at trial. As a consequence of Sirius XM’s infringement, Fraunhofer is entitled to recover damages adequate to compensate it for the infringement complained of herein, but in no event less than a reasonable royalty.

63. Fraunhofer has suffered irreparable injury as a direct and proximate result of Sirius XM’s infringement for which there is no adequate remedy at law. Unless Sirius XM is enjoined, Fraunhofer will continue to suffer such irreparable injury as a direct and proximate result of Sirius XM’s conduct.

FOURTH CLAIM FOR RELIEF
Infringement of U.S. Patent No. 7,061,997

64. Fraunhofer realleges and incorporates by reference Paragraphs 1 through 63 of this Complaint, as if fully set forth herein.

65. Fraunhofer is the owner of the entire right, title, and interest in and to the ’997 Patent.

66. Sirius XM has infringed and is currently infringing one or more claims of the ’997 Patent, in violation of 35 U.S.C. § 271.

67. Sirius XM has infringed and continues to infringe the ’997 Patent in violation of 35 U.S.C. § 271(a), literally and/or under the doctrine of equivalents by, among other things, making, using, offering for sale, selling, and/or importing within this judicial district and elsewhere in the United States, without license or authority, the XM DARS System and the Sirius XM Satellite Radios and related products and/or processes falling within the scope of one or more claims of the ’997 Patent.

68. Sirius XM has infringed and continues to infringe the ’997 Patent in violation of 35 U.S.C. § 271(b) by actively inducing infringement of the ’997 Patent, literally and/or under the doctrine of equivalents, with knowledge of the ’997 Patent and knowledge that it was inducing the infringement of the ’997 Patent by, among other things, actively and knowingly aiding and abetting, assisting and encouraging others, including without limitation Toyota, General Motors, Volkswagen, Nissan, and Hyundai, and other automakers, and other customers and end users of Sirius XM products and services, to directly infringe the ’997 Patent with respect to the making, using, offering for sale, selling, and/or importing within this judicial district and elsewhere in the United States, without license or authority, the XM DARS System and the Sirius XM Satellite Radios and related products and/or processes falling within the scope of one or more claims of the ’997 Patent. Moreover, Sirius XM had knowledge of its infringement of the Fraunhofer patents, as set forth in Paragraphs 32 through 36.

69. Sirius XM has infringed and continues to infringe the ’997 Patent in violation of 35 U.S.C. § 271(c) by contributing to infringement of the ’997 Patent, literally and/or under the doctrine of equivalents, by, among other things, selling, offering for sale, and/or

importing within this judicial district and elsewhere in the United States, without license or authority, the XM DARS System and the Sirius XM Satellite Radios and related products and/or processes falling within the scope of one or more claims of the ’997 Patent, with knowledge of the ’997 Patent knowing that such products and/or components are especially made or especially adapted for use in the infringement of the ’997 Patent, and not staple articles or commodities of commerce suitable for substantial noninfringing use. Moreover, Sirius XM had knowledge of its infringement of the Fraunhofer patents, as set forth in Paragraphs 32 through 36.

70. Fraunhofer is informed and believes, and on that basis alleges, that Sirius XM’s infringement of the ’997 Patent has been and continues to be willful and deliberate, entitling Fraunhofer to increased damages under 35 U.S.C. § 284. Fraunhofer is informed and believes, and on that basis alleges, that Sirius XM, had knowledge of the ’997 Patent and its infringement, and nonetheless engaged in objectively reckless conduct by continuing to engage in infringing activity in the face of an objectively high risk that Sirius XM was infringing Fraunhofer’s valid ’997 Patent.

71. Sirius XM’s acts of infringement have caused damage to Fraunhofer in an amount to be proven at trial. As a consequence of Sirius XM’s infringement, Fraunhofer is entitled to recover damages adequate to compensate it for the infringement complained of herein, but in no event less than a reasonable royalty.

72. Fraunhofer has suffered irreparable injury as a direct and proximate result of Sirius XM’s infringement for which there is no adequate remedy at law. Unless Sirius XM is enjoined, Fraunhofer will continue to suffer such irreparable injury as a direct and proximate result of Sirius XM’s conduct.

DEMAND FOR JURY TRIAL

Pursuant to Federal Rule of Civil Procedure 38(b), Plaintiff Fraunhofer hereby demands a trial by jury on all issues so triable of right by a jury raised in this Complaint.

PRAYER FOR RELIEF

WHEREFORE, Fraunhofer respectfully requests that the Court enter judgment as follows:

1. That Sirius XM has directly infringed the ’289, ’1084, ’3084, and ’997 Patents;

2. That Sirius XM has induced the infringement of the ’289, ’1084, ’3084, and ’997 Patents;

3. That Sirius XM has contributorily infringed the ’289, ’1084, ’3084, and ’997 Patents;

4. That Sirius XM and any of their affiliates, subsidiaries, officers, directors, employees, agents, representatives, licensees, successors, assigns, and all those acting for any of them and/or on any of their behalf, or acting in concert with any of them directly or indirectly, be enjoined from infringing, inducing others to infringe or contributing to the infringement of the ’289, ’1084, ’3084, and ’997 Patents;

5. That Sirius XM be ordered to pay compensatory damages to Fraunhofer, together with pre-judgment interest and post-judgment interest as allowed by law;

6. That Sirius XM be ordered to provide an accounting;

7. That Sirius XM be ordered to pay supplemental damages to Fraunhofer, including without limitation interest;

8. That the infringement by Sirius XM be adjudged willful and that the damages be increased under 35 U.S.C. § 284 to three times the amount found or measured;

9. That the Court enter judgment against Sirius XM, and in favor of Fraunhofer, in all respects;

10. That the Court determine this is an exceptional case under 35 U.S.C. § 285 and an award of attorneys' fees and costs to Fraunhofer is warranted in this action; and

11. For any such other and further relief as this Court may deem just and proper.

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Dated: February 22, 2017

Exhibit A

US006314289B1

(12) **United States Patent**
Eberlein et al.

(10) **Patent No.: US 6,314,289 B1**
(45) **Date of Patent: Nov. 6, 2001**

(54) **APPARATUS AND METHOD FOR TRANSMITTING INFORMATION AND APPARATUS AND METHOD FOR RECEIVING INFORMATION**

(75) Inventors: **Ernst Eberlein**, Grossenseebach; **Marco Breiling**, Erlangen; **Jan Stoessel**, Nürnberg; **Heinz Gerhäuser**, Waischenfeld, all of (DE)

(73) Assignee: **Fraunhofer-Gesellschaft zur Förderung der angewandten Forschung e.V.**, Munich (DE)

(*) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 0 days.

(21) Appl. No.: **09/202,729**

(22) PCT Filed: **Dec. 3, 1998**

(86) PCT No.: **PCT/EP98/07850**

§ 371 Date: **May 17, 1999**

§ 102(e) Date: **May 17, 1999**

(87) PCT Pub. No.: **WO00/36783**

PCT Pub. Date: **Jun. 22, 2000**

(51) **Int. Cl.**^{7} **H04Q 7/20**
(52) **U.S. Cl.** **455/427**; 455/3.02; 455/10; 714/746; 714/758; 375/225
(58) **Field of Search** 455/427, 3.02, 455/10, 12.1, 137; 714/701, 746, 764, 751, 752, 758, 767, 763, 786; 375/290, 264, 225, 250

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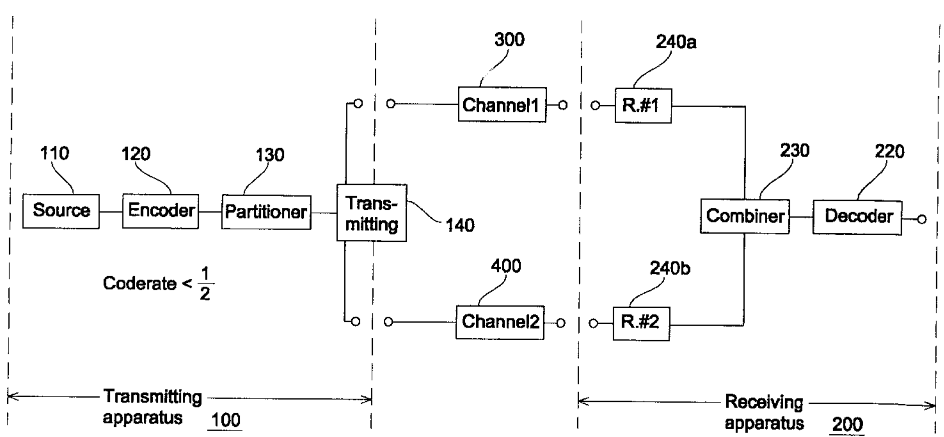
Assistant Examiner—Congvan Tran
(74) *Attorney, Agent, or Firm*—Dougherty & Clements LLP

(57) **ABSTRACT**

An apparatus for transmitting information comprises a bitstream source for providing a bitstream representing the information, a redundancy adding encoder for generating an encoded bitstream, which is arranged to output, for a first number of input bits, a second number of output bits, the second number of output bits having at least twice as many output bits as the first number of input bits, wherein the second number of output bits includes two portions of output bits, each portion of output bits individually allowing the retrieval of information represented by the first number of input bits, and the first portion of output bits being coded based on the bitstream in a different way with respect to the second portion of output bits. The apparatus further comprises a partitioner for partitioning the second number of output bits into the two portions of output bits and a transmitter for transmitting the output bits of the first portion via a first channel and the output bits of the second portion via a second channel, the second channel being spatially different from the first channel. An inventive receiving apparatus combines the signals received via the first and second channels and uses both channel signals for channel decoding by removing redundancy. Thus, the transmitting receiving system is suitable for providing time and/or space diversity and, in the optimal case, provides a C/N value which is greater than 4.3 dB with respect to a two-channel system comprising a duplicator in the transmitter and a channel-controlled switch in the receiver.

35 Claims, 4 Drawing Sheets



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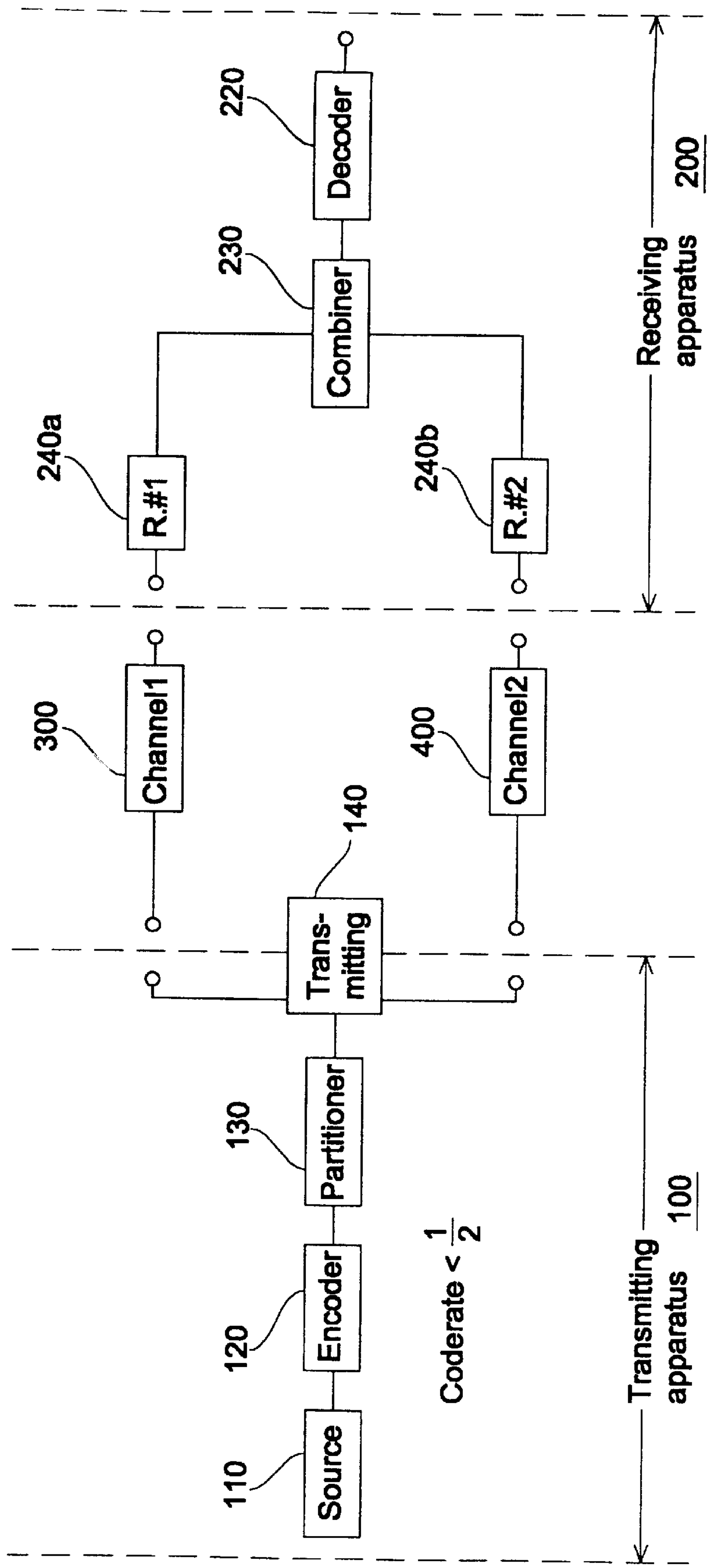

Source
110
Encoder
120
Partitioner
130
Trans-
mitting
140
Coderate < 1/2
Transmitting
apparatus
100
Channel1
300
Channel2
400
R.#1
240a
R.#2
240b
Combiner
230
Decoder
220
Receiving
apparatus
200

Fig. 1

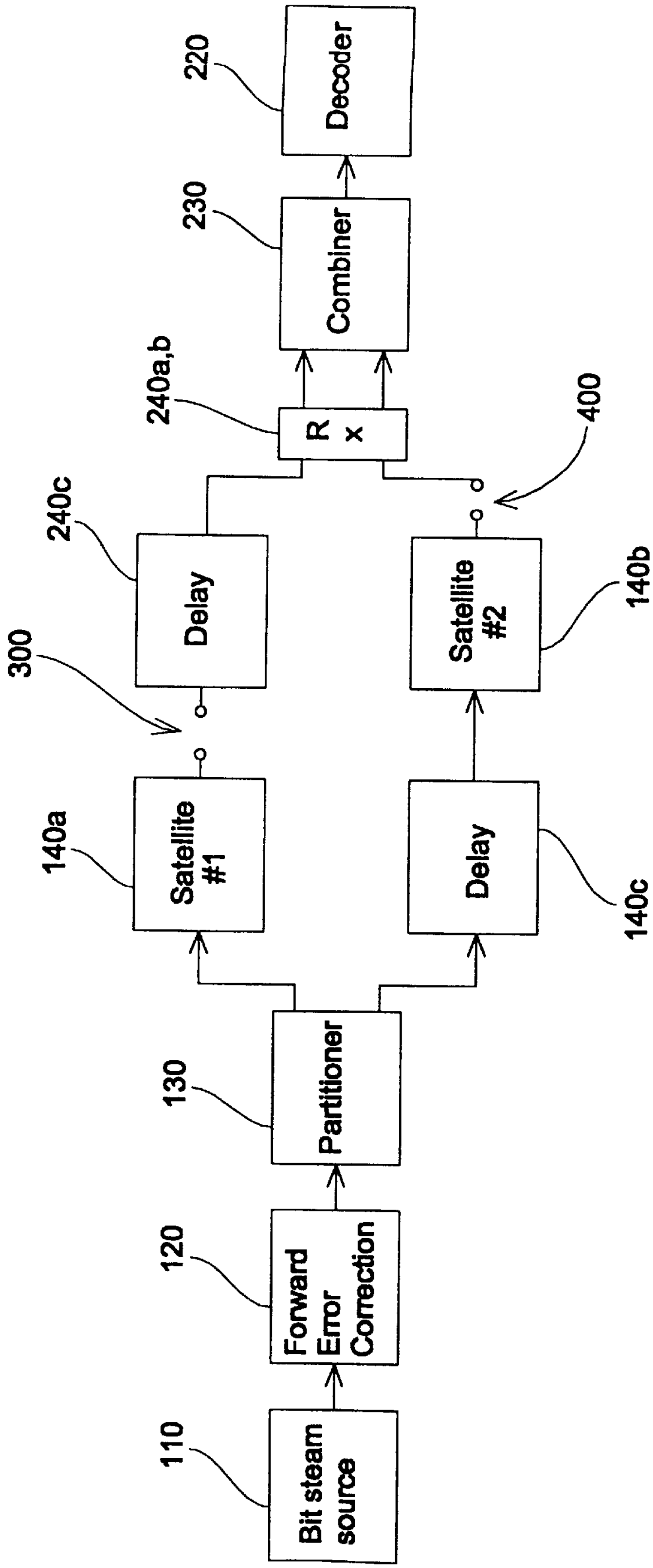
110
Bit steam source
120
Forward Error Correction
130
Partitioner
140a
Satellite #1
140c
Delay
300
240c
Delay
140b
Satellite #2
400
240a,b
R x
230
Combiner
220
Decoder

Fig. 2

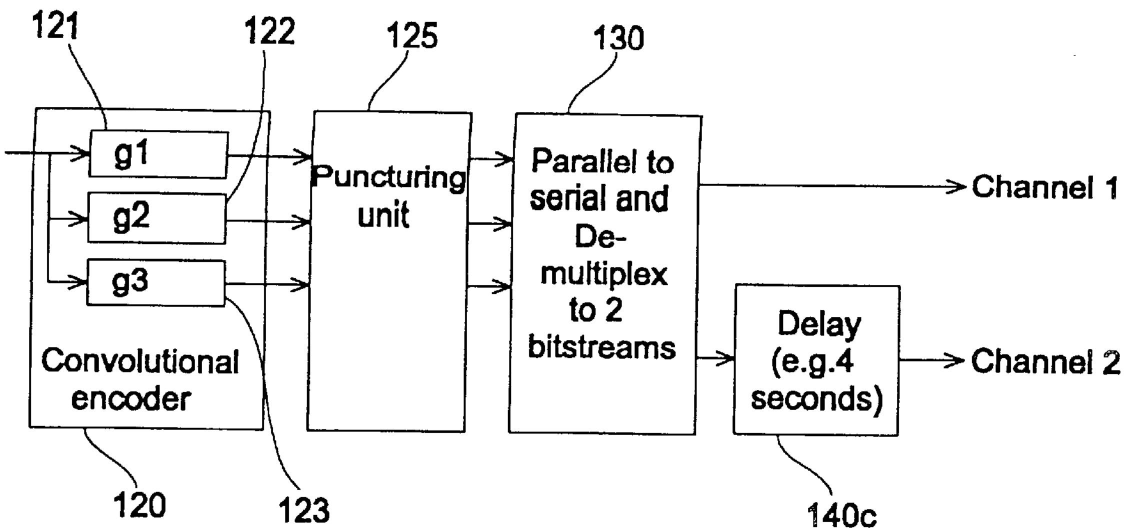
121
122
125
130
g1
g2
g3
Puncturing unit
Parallel to serial and De-multiplex to 2 bitstreams
Channel 1
Delay (e.g.4 seconds)
Channel 2
Convolutional encoder
120
123
140c

Fig. 3

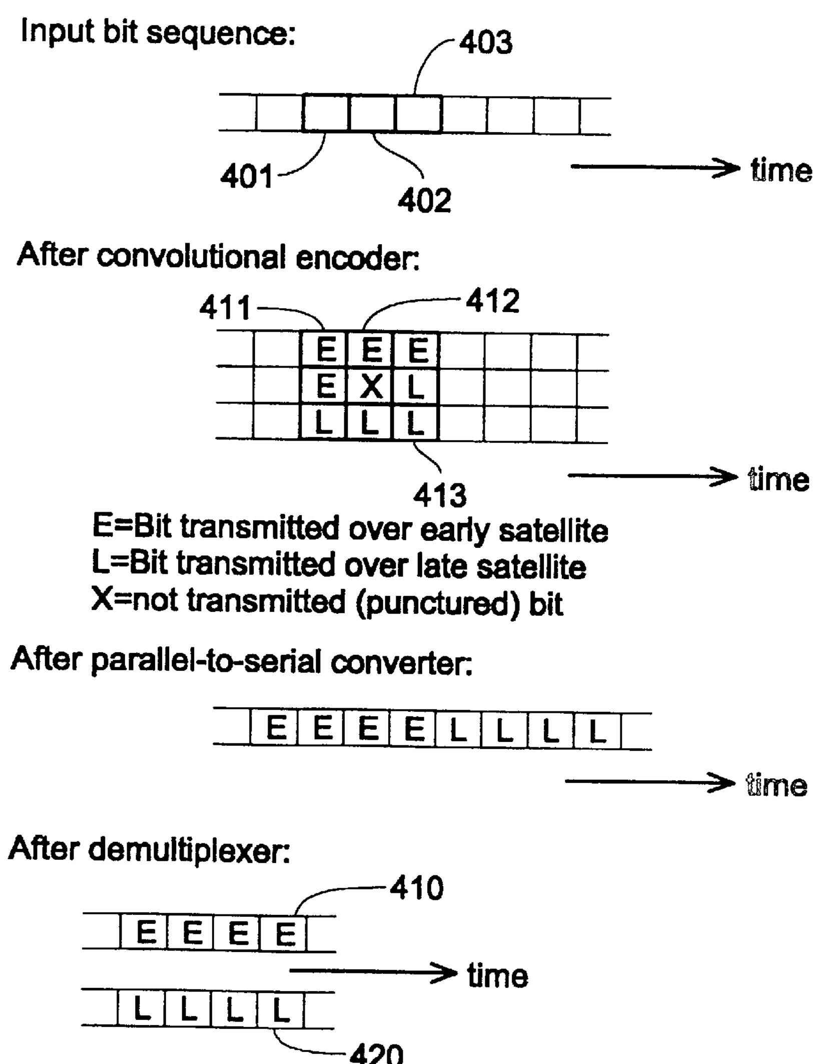
Input bit sequence:
403
401
402
time
After convolutional encoder:
411
412
E E E
E X L
L L L
413
time
E=Bit transmitted over early satellite
L=Bit transmitted over late satellite
X=not transmitted (punctured) bit
After parallel-to-serial converter:
E E E E L L L L
time
After demultiplexer:
410
E E E E
time
L L L L
420

Fig. 4

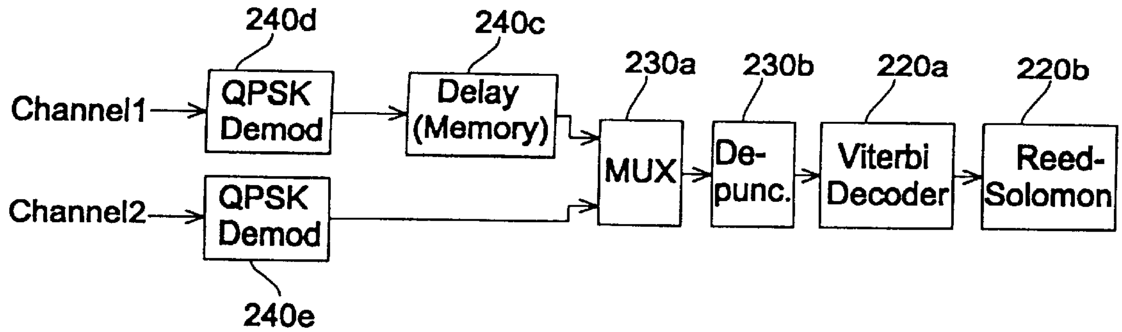

Fig. 5

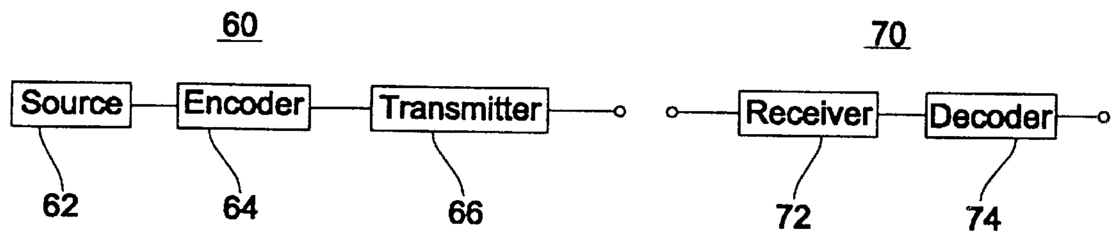

Fig. 6

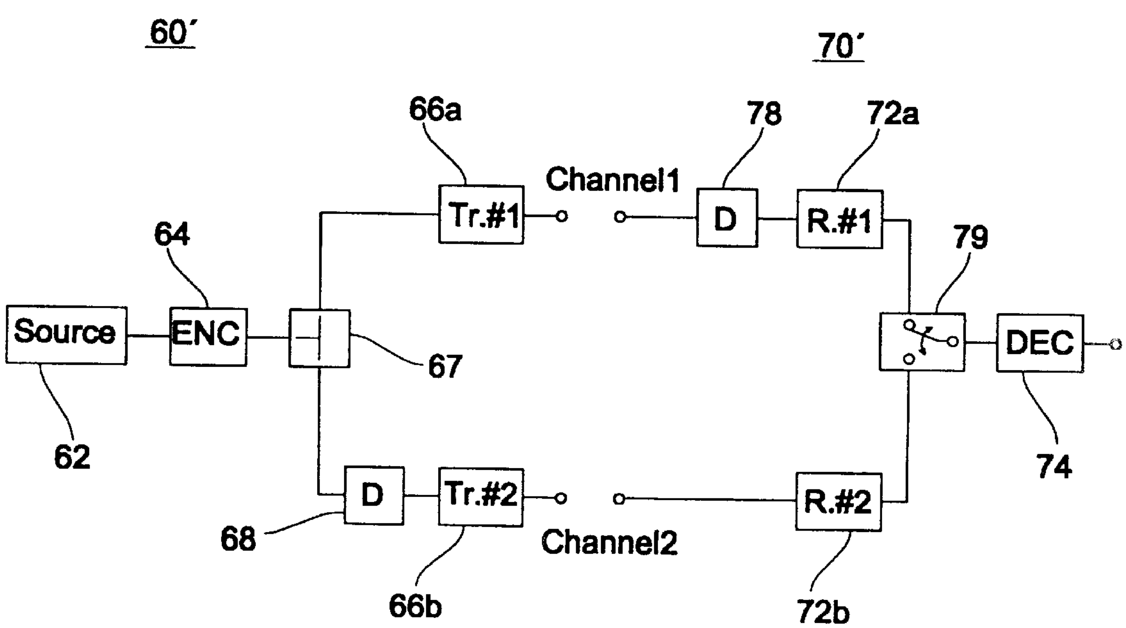

Fig. 7

APPARATUS AND METHOD FOR TRANSMITTING INFORMATION AND APPARATUS AND METHOD FOR RECEIVING INFORMATION

FIELD OF THE INVENTION

The present invention relates to concepts for digital broadcasting and, in particular, concepts for digital broadcasting suited for fading channels for wireless communication.

BACKGROUND OF THE INVENTION

Satellite-based broadcasting systems provide an adequate communication link only in rural areas, in which only a small number of e.g. bridges exist. Additionally, rural areas usually do not have skyscrapers. Skyscrapers as well as bridges or, generally, densly built-up areas are obstacles to satellite-based communication systems, since carrier frequencies used for such communication links involve that a channel between a sender, e.g., a satellite, and a receiver, i. e. a mobile or stationary receiver, is characterised by the line of visual contact (line of sight) between the sender and the receiver. If a skyscraper comes into the line of visual contact, i.e., the transmission channel between the satellite and the receiver, which may be positioned in a car, the received signal power will decrease substantially.

Generally, it can be stated that in wireless systems (radio systems), changes in the physical environment cause the channel to fade. These changes include both relative movement between transmitter and receiver and moving scatters/reflectors in the surrounding space. In theoretical studies of wireless systems, the real channels are usually modelled so that they result in trackable analysis. The two major classes of fading characteristics are known as Rayleigh and Rician. A Rayleigh-fading environment assumes no line of sight and no fixed reflectors/scatters. The expected value of the fading is zero. If there is a line of sight, this can be modelled by Rician-fading, which has the same characteristics as the Rayleigh-fading, except for a non-zero expected radio.

Modern digital broadcasting systems know several means for reducing the impact of a channel fading. These concepts comprise channel coding on the one hand and several kinds of diversity on the other hand. The European standard for digital audio broadcasting (DAB), set out in Radio Broadcasting Systems; Digital Audio Broadcasting (DAB) To Mobile, Portable and Fixed Receivers, ETS 300 401, ETS I—European Telecommunications Standards Institute, Valbonne, France, February 1995, uses differential quadrature phase-shift keying (DQPSK) as modulation technique. The channel encoding process is based on punctured convolutional coding, which allows both equal and unequal error protection. As a mother code, a convolutional code having a code rate of 1/4, a constraint length 7, and octal polynominals is used. The puncturing procedure allows the effective code rate to vary between 8/9 and 1/4. Channel coding by means of punctured convolutional codes is described in "Punctured Convolutional Codes of Rate (n−1)/n and Simplified Maximum Likelihood Decoding", J. Bibb Cain et al., IEEE Transactions on Information Theory, Vol. IT-25, No. 1, January 1979.

Punctured convolutional codes can be used in connection with many modulation techniques, such as OFDM, BPSK, QAM, etc.

Different channel encoding techniques are outlined in "Channel Coding with Multilevel/Phase Signals", Gottfried Ungerboeck, IEEE Transactions on Information Theory, Vol. IT 28, No. 1, pages 55 to 66, January 1982.

Bitstreams encoded by means of a convolutional encoder can be decoded by a decoder, in which the well-known Viterbi algorithm is implemented. This algorithm is capable of using the channel state information (see P. Hoeher "TCM on Frequency-Selective Length-Mobile Fading Channels", Proc. Tirrenia International Workshop Digital Communication, Tirrenia, Italy, September 1991). The Viterbi algorithm can be modified to provide reliability estimates together with the decoded sequence. This enables soft decoding. By applying a soft-output Viterbi algorithm, an improvement of about 2 dB is obtained in comparison to systems that implement "hard" decision.

DESCRIPTION OF PRIOR ART

With reference to FIG. **6**, a simplified overview of a transmitter receiver system described in the European DAB Standard is illustrated. The transmitter receiver system generally comprises a transmitter section **60** and a receiver section **70**. The transmitter section **60**, in the simplest case, comprises a bitstream source **62**, a channel encoder **64** and a transmitter **66**. The receiver section **70**, in the simplest case, comprises a receiver **72** and a channel decoder **74**.

FIG. **7** illustrates a transmitting receiving setup providing for time diversity as well as space diversity. The transmitter section **60'** comprises the bitstream source **62** and the encoder **64** that have already been described with respect to FIG. **6**. In addition, the receiver section **60'** comprises a first transmitter **66*a*** and a second transmitter **66*b***. Both transmitters **66*a*** and **66*b*** are fed by the same signal output by the encoder **64** that is duplicated by a duplicator **67**.

To obtain time diversity, a delay element **68** is coupled between the duplicator **67** and the second transmitter **66*b***.

In the case of satellite communication, the transmitters **66*a*** and **66*b*** are realised by two satellites that reside on different orbital positions spaced apart from each other.

The first channel is defined by the line of sight between the first transmitter and the receiver, for example, a car, whereas the second channel is defined by the line of sight between the second transmitter **66*b*** and the car that comprises the receiving section **70'**. In the scenario, in which the car travels on a street to the right and to the left of which are high buildings, the possibility is increased that the car will receive the transmitted signal from at least one satellite.

When the case is considered, in which the car is driving through a tunnel or under a bridge, the lines of sight to both transmitters **66*a*** and **66*b*** are interrupted. The time diversity method implemented by this system shown in FIG. **7**, however, ensures that the receiver will not be affected by the interrupted channel, since the transmission signal is delayed by the delay stage **68**. Optimally, no transmission interruption will result, when the delay time is equal to or greater than the travelling time of the car through the tunnel or under the bridge. Thus, the receiving section will, once again, receive the transmission signal sent by the transmitter **66*a***, when it was under the bridge, via a channel **2**. Naturally, the receiving section **70'** comprises another delay stage **78**. As it is shown in FIG. **7**, the delay stage **78** of the receiving section has to be in the channel that has not been delayed in the transmitter section. Thus, the signals at the output of the receivers **72*a*** and **72*b*** are identical, when the delay values of the delay stages **78** and **68** are equal.

A decision stage **79**, which is symbolised as a switch in FIG. **7**, determines which channel provides the signal with the better signal to noise ratio. When it is determined that channel **1** provides the stronger signal, the decision stage **79** is operative to conduct the signal received by the receiver

72*a* into the channel decoder **74** When it is determined in block **79** that the signal transmitted over the other channel (channel **2**) is the stronger one, the decision stage **79** is operative to conduct the signal received by the receiver **72***b* to the channel decoder **74**.

To summarise, the system illustrated in FIG. **7** comprises the following essential features:

- the signal output by the encoder **64** is duplicated by the duplicator **67**;
- exactly the same signals, whether delayed or not, are transmitted via both channels;
- the signals transmitted over both channels are derived from the bitstream output by the bitstream source **62** in exactly the same way by means of the encoding process carried out in the redundancy adding encoder **64** (repetition code);
- the decision stage **79** compares the signal to noise ratio of both channels and selects the channel in which the signal having the better signal to noise ratio is transmitted;
- the signal transmitted via the other channel is discarded; and
- the channel decoder **74** only uses one channel, i.e., the channel determined by the decision stage **79**, for channel decoding.

Besides the technique of channel encoding using a redundancy adding encoder like a convolutional encoder, different types of diversity, e.g., time diversity and space diversity, can be implemented to ease the impact of fading channels.

The bitstream source **62** can be implemented as an audio encoder as defined by ISO-MPEG. It provides a bitstream comprising useful information, i.e., encoded spectral values of a block of audio samples, and side information. To enhance the robustness of the communication link, a forward error correction encoding is performed by the convolutional encoder **64**. In general, the convolutional encoding procedure generates redundancy in the transmitted datastream in order to provide ruggedness against transmission distortion.

Usually, convolutional encoders consist of a specific number of shift registers and a number of XOR gates. The convolutional encoder described in the ETS Standard is a convolutional encoder having a code rate of 1/4. This means that the convolutional encoder produces four output bits for one input bit. As it is well known in the art, each output bit is derived from the current input bit and a specific combination of a certain number of preceding input bits stored in the shift registers. The specific combination of the current input bit and certain preceding input bits for each encoder output bit is defined by the so-called generator polynominals. The octal forms of the generator polynominals defined in the ETS 300 401 are **133**, **171**, **145** and **133**.

The encoded bitstream can be punctured for raising the code rate from 1/4 to another code rate, e.g., 8/9. "Puncturing" means that certain bits in the convolutional encoder output bits are discarded and not forwarded to the transmitter **66**. Thus, puncturing operates to again reduce redundancy in an encoded bitstream, which has been added by the convolutional encoder.

The transmitter **66** may comprise usual transmitter elements, such as a QPSK modulator, an IFFT block (IFFT= Inverse Fast Fourier Transform) for performing orthogonal frequency division multiplexing, a guard interval inserter, a synchronisation sequence inserter and modulation means for modulating the signal onto a high frequency carrier.

Analogously, the receiver **72** comprises an HF front end, an analog/digital converter, and a QPSK demodulator. The signal output by the receiver is input in the decoder **74**. The decoder **74** is operative to decode the encoded bitstream output by the receiver **72**. In modern communication systems, the decoder **74** implements the above-outlined soft-input Viterbi algorithm. As it has already been outlined, the Viterbi decoder performs a maximum likelihood decoding using the channel state information, which is also called "metric". Different algorithms are known for Rician and Rayleigh channels.

Especially in satellite-based communication systems, design engineers are confronted with strong demands for reducing transmitter power. Reduced transmitter power directly translates into system costs. Generally, the costs for designing and transporting the satellite(s) into its (their) orbital position(s) are directly proportional to the power supply needed on board of the satellite. Higher transmitter power on board of the satellite also means higher energy producing capabilities of the satellite. Thus, it can be stated that, under costs aspects, reducing transmitter power is essential.

Therefore, the system described in FIG. **7** is disadvantageous in that, in the receiver, only one channel is used for retrieving information, whereas the other channel is discarded. In extreme situations, in which one channel has faded totally, no transmitter power from one transmitter, i.e., one satellite, will reach the receiver. Normally, however, the channels will not fade totally. Instead, both channels will fade more or less. Thus, the decision stage **79** has to select one out of two useful signals. When the case is considered that both signals output by the receivers **72***a* and **72***b* have identical signal to noise ratios, only one signal is selected, whereby the transmitter power from the satellite transmitting via the other channel is wasted totally.

SUMMARY OF THE INVENTION

It is the object of the present invention to provide an apparatus and method of transmitting information and an apparatus and method of receiving information, which result in better receiver output signal quality and/or reduced transmitter power demands.

In accordance with a first aspect of the present invention, this object is attained by an apparatus for transmitting information, comprising a bitstream source for providing a bitstream representing the information; a redundancy adding encoder for generating an encoded bitstream based on the bitstream provided by the bitstream source wherein the encoder is arranged to output, for a first number of input bits, a second number of output bits, the second number of output bits having at least twice as many output bits as the first number of input bits, and wherein the second number of output bits includes two portions of output bits, each portion of output bits individually allowing the retrieval of information represented by the first number of input bits, and the first portion of output bits being coded based on the bitstream in a different way with respect to the second portion of output bits; a partitioner for partitioning the second number of output bits into the two portions of output bits; and a transmitter for transmitting the output bits of the first portion via a first channel and the output bits of the second portion via a second channel, the second channel being spatially different from the first channel.

In accordance with a second aspect of the present invention, this object is attained by an apparatus for receiving information, the information being represented by an encoded bitstream, the encoded bitstream being encoded such that its redundancy is at least doubled with respect to a bitstream from which the encoded bitstream is derived, and that, for a first number of bits of the bitstream, the encoded bitstream comprises a second number of bits, the second number of bits having at least twice as many bits as the first number, and wherein the second number of bits includes two portions of bits, each portion of bits individually allowing the retrieval of information represented by the first number of bits, and the first portion of the bits being encoded in a different way with respect to the second portion of bits, the apparatus comprising a receiver for receiving the first portion of bits via a first channel and the second portion of bits via a second channel, the first and the second channels being spatially different from each other; a combiner for combining the first and the second portions; and a decoder for decoding the coded bitstream by removing redundancy from the coded bitstream, the decoder using the first and second portions of bits combined by the combiner.

In accordance with a third aspect of the present invention, this object is attained by a method of transmitting information, comprising the following steps: providing a bitstream representing the information; generating a redundancy added encoded bitstream based on the bitstream provided in the step of providing, wherein for a first number of input bits, a second number of output bits is generated, the second number of output bits having at least twice as many output bits as the first number of input bits, and wherein the second number of output bits includes two portions of output bits, each portion of output bits individually allowing the retrieval of information represented by the first number of input bits, and the first portion of output bits being coded based on the bitstream in a different way with respect to the second portion of output bits; partitioning the second number of output bits into the two portions of output bits; and transmitting the output bits of the first portion via a first channel and the output bits of the second portion via a second channel, the second channel being spatially different from the first channel.

In accordance with a fourth aspect of the present invention, this object is attained by a method of receiving information, the information being represented by an encoded bitstream, the encoded bitstream being encoded such that its redundancy is at least doubled with respect to a bitstream from which the encoded bitstream is derived, and that, for a first number of bits of the bitstream, the encoded bitstream comprises a second number of bits, the second number of bits having at least twice as many bits as the first number, and wherein the second number of bits includes two portions of bits, each portion of bits individually allowing the retrieval of information represented by the first number of bits, and the first portion of the bits being encoded in a different way with respect to the second portion of bits, the method comprising the following steps: receiving the first portion of bits via a first channel and the second portion of bits via a second channel, the first and the second channels being spatially different from each other; combining the first and the second portions; and decoding the coded bitstream by removing redundancy from the coded bitstream, wherein the first and second portions of bits combined in the step of combining are used in the step of decoding.

The present invention is based on the finding that, although there are two physically different channels both channels are considered as one single channel from the viewpoint of the channel decoder located in the receiving section. This means that the channel decoder in the receiving section does not know that the signals it decodes stem from two physically, i. e. spatially, different channels. However, the inventive system, in fact, provides two different physical channels to allow for time and/or space diversity.

The space diversity can be obtained by two terrestrial transmitters, by two satellite transmitters or by one satellite transmitter and one terrestrial transmitter.

In accordance with the present invention, an apparatus for transmitting information comprises a bitstream source for providing a bitstream representing the information. A redundancy adding encoder for generating an encoded bitstream based on the bitstream provided by the bitstream source is arranged to output, for a first number of input bits, a second number of output bits, the second number of output bits having at least twice as many output bits as the first number of input bits, and wherein the second number of output bits includes two portions of output bits, each portion of output bits individually allowing the retrieval of information represented by the first number of input bits, and the first portion of output bits being coded based on the bitstream in a different way with respect to the second portion of output bits. A means for partitioning, i.e., a partitioner, receives the output of the redundancy adding encoder and partitions the second number of output bits into the two portions of output bits. Means for transmitting transmit the output bits of the first portion via a first channel and the output bits of the second portion via a second channel, wherein the second channel is spatially different from the first channel.

In accordance with another aspect of the present invention, an apparatus for receiving information comprises a receiver for receiving the first portion of bits via a first channel and the second portion of bits via a second channel, a combiner for combining the first and the second portions and a decoder for decoding the coded bitstream by removing redundancy from the coded bitstream, the decoder using the first and second portions of bits combined by the combiner.

This inventive transmitter receiver concept provides the following advantages:

- two channels allow time and/or space diversity;
- the partitioner partitions rather than duplicates the output signal of the encoder into two portions of output bits;
- the combiner in the receiver combines rather than selects the signals received from both channels and feeds the combined signal into the channel decoder;
- the signals from both channels are used for decoding all the time;
- in the best case, in which the signal powers in both channels are identical, transmitter power used for transmitting via each channel can be halved at least, thus, halving system costs with respect to the system illustrated in FIG. **7**; and
- when the transmitter powers are not changed, the signal quality output by the channel decoder can be considerably improved.

BRIEF DESCRIPTION OF THE DRAWINGS

The foregoing and other objects, features and advantages of the invention will become more readily apparent from the following detailed description of preferred embodiments which proceeds with reference to the drawings.

FIG. **1** shows a principle overview of a transmission receiving system in accordance with the present invention, comprising an inventive transmitter and an inventive receiver.

FIG. **2** shows a more detailed block diagram of the transmission receiving system shown in FIG. **1**, in which time and space diversity are embodied.

FIG. **3** shows a detailed block diagram of an inventive transmitter section.

FIG. **4** shows an input bit sequence and an output bit pattern of a convolutional encoder used in an inventive transmitter section.

FIG. **5** shows a detailed view of an inventive receiver section.

FIG. **6** shows a generalised block diagram of a prior art transmitting receiving system.

FIG. **7** shows a block diagram of a transmitter receiver system implementing time and space diversity, in which the output of the transmitter encoder is duplicated and a channel selection is performed in the receiver.

DESCRIPTION OF PREFERRED EMBODIMENTS

In FIG. **1** a general block diagram of an inventive apparatus for transmitting **100** and an inventive apparatus for receiving **200** is illustrated. The transmitting apparatus **100** comprises a bitstream source **110**, a redundancy adding encoder **120** and a partitioner **130**. The bitstream source **110** may be an MPEG encoder as described above. The encoder **120** is generally a redundancy adding encoder for generating an encoded bitstream on its output, wherein the encoder **120** is arranged to output, for a first number of input bits, a second number of output bits, the second number of output bits having at least twice as many output bits as the first number of input bits. This means that the encoder **120** implements a code rate equal to or less than 1/2. As it is known in the art, the code rate is defined by the number of input bits divided by the number of output bits produced by the encoder based on the number of input bits. In other words, a code rate 1/2 means that for each input bit, two output bits are produced. Analogously, a code rate of 1/3 means that for each input bit, three output bits are produced. Similarly, a code rate of 3/8 means that for three input bits, eight output bits are produced.

The code rate of the encoder **120** is set to be smaller than 1/2, such that the second number of output bits can be sub-divided into two portions of output bits, such that each portion of output bits individually allows the retrieval of information represented by the first number of input bits This means that a decoder **220** located in the receiving apparatus is able to retrieve information represented by the bitstream output by the bitstream source **110** when only one channel, i.e., channel 1 **300** or channel 2 **400** provides a useful signal, whereas the other channel has faded totally.

Another feature of the encoder **120** is that the first portion of output bit is coded based on the bitstream in a different way with respect to the second portion of output bits. In contrast to a simple repetition code in which redundancy is doubled by simply duplicating a signal to transmitted coded, the channel decoder **220** capabilities are enhanced, since the signal is transmitted over the channels **300** and **400** are derived from the bitstream output by the bitstream source **110** independently of each other. The partitioner **130** feeds means for transmitting, i.e., a transmitter, **140** for transmitting the first portion of output bits via the first channel **300** and the second portion of output bits via the second channel **400**. It is to be noted that both channels **300** and **400** are spatially different from each other.

As usual, a channel between the transmitter and the receiver is defined by the line of sight connection between the transmitter and the receiver. Thus, two channels are different from each other when a mobile receiver has moved with respect to a single transmitter, or when two transmitters exist positioned in different locations, e.g., orbital positions. In this case, it does not play any role whether the receiver is a mobile or a stationary receiver.

Thus, the transmitting means **140** may comprise one transmitter, e.g., one satellite and a delay stage, such that two different channels are created between the single transmitter and a mobile receiver, when the mobile receiver is at a first position and between the single transmitter and the mobile receiver when the mobile receiver has moved to a second position after the period defined by the delay stage in the transmitter. This concept is called time diversity for mobile receivers. Naturally, it is not possible to create two channels different from each other between a single stationary transmitter and a stationary receiver.

Alternatively, as it is described with reference to FIG. **2**, the transmitting means **140** comprise two transmitters positioned in different locations, to obtain space diversity.

The receiving apparatus **200** illustrated in FIG. **1** comprises receiving means **240***a* and **240***b*, the receiving means, i.e., the receiver, comprising a first receiver **240***a* for receiving the first portion of output bits transmitted via the first channel **300** and a second receiver **240***b* for receiving the second portion of output bits via the second channel **400**.

In accordance with the present invention, the output signals of the receiving means **240***a* and **240***b* are combined in a combiner **230** such that the output signals of both receivers are used in the channel decoder **220**.

FIG. **2** illustrates a transmission receiving system in accordance with the preferred embodiment of the present invention. The transmitting apparatus comprises, as already described in FIG. **1**, the bitstream source **110**, the encoder **120** generally termed as forward error correction, the partitioner **130** and transmitting means comprising a first satellite **140***a*, a second satellite **140***b* and a delay stage **140***c*.

The receiving apparatus comprises the channel decoder **220**, the combiner **230** and the receiving means (Rx) comprising the first and second receivers **240***a*, **240***b* and a delay stage **240***c*. The transmitting apparatus and the receiving apparatus are "connected" by the first channel **300** and the second channel **400**.

By using the delay stages **140***c* and **240***c*, which are positioned in opposite channels, time diversity is implemented in the transmission receiving system shown in FIG. **2**. Furthermore, by means of the provision of two transmitters, i.e., the first satellite **140***a* and the second satellite **140***b*, space diversity or spatial diversity is implemented into the inventive transmission receiving system.

With reference to FIG. **3**, a more detailed block diagram of the transmitting apparatus is described. The encoder **120** in the transmitting apparatus is implemented as a convolutional encoder in accordance with the present invention. As it is shown in FIG. **3**, the convolutional encoder comprises three generator polynominals, i.e., a first generator polynominal g1 **121**, a second generator polynominal g2 **122** and a third generator polynominal g3 **123**. Thus, the convolutional encoder **120** has a code rate of 1/3, since, for one input bit, the encoder produces three output bits. The transmitting apparatus shown in FIG. **3** further comprises a puncturing unit **125** that reduces the number of bits, i.e., the number of output bits, such that an even number of output bits to be transmitted over the first and second channel is obtained. The puncturing unit **125** is connected to the partitioner **130**, that, in accordance with the preferred embodiments of the present invention, comprises a parallel-to-serial converter and a demultiplexer to demultiplex the serial bitstream produced by the parallel-to-serial converter into two bitstreams. The block diagram in FIG. **3** further comprises the delay stage **140***c* of the transmitting means. The first transmitter and the second transmitter are not shown in FIG. **3**.

Thus, the first portion of output bits is transmitted via the first channel, whereas the second portion of output bits is delayed by the delay stage, transmitted via the second channel.

With reference to FIG. **4**, the functionality of the convolutional encoder **120**, the puncturing unit **125** and the partitioner **130** will be described. In FIG. **4**, an input bit sequence having bits **401**, **402** and **403** is illustrated. The convolutional encoder **120** will produce three parallel arranged output bits **411**, **412** and **413** for each input bit **401**, **402** and **403**. The notation of the output bits **411** to **413** relates to the channel, over which the respective bit is transmitted. Thus, bits termed E are transmitted over the early satellite, i.e., satellite **140***a* (FIG. **2**), whereas the bits termed L are transmitted over the late satellite, i.e., the satellite **140***b* (FIG. **2**), which input is delayed by the delay stage **140***c*. The bit termed X is not transmitted at all. This bit is discarded by the puncturing unit **125** to obtain a second number of output bits, which is an even number. In accordance with the preferred embodiment of the present invention, an even number of output bits to be transmitted by the transmitting means **140** (FIG. **1**) is required, since two channels exist and the number of bits transmitted over each channel are equal in the preferred embodiment. It has to be noted that equal numbers of bits in each channel are not essential for the present invention.

The output of the puncturing unit **125** is fed into a parallel-to-serial converter included in the combiner **130** (FIG. **3**) such that a serial bitstream, i.e., the second number of output bits, is obtained. The demultiplexer included in the partitioner **130** demultiplexes the serial bitstream output by the parallel-to-serial converter into two bitstreams, in order to produce the first portion **410** and the second portion **420** of output bits.

It has to be noted that the number of bits in each of the first and second portions **410** and **420** is larger than the first number of input bits **401**, **402** and **403** input into the convolutional encoder **120**. Thus, some redundancy still exists to be used by the channel decoder **220** (FIG. **2**), when one channel is totally lost, for instance, when a mobile receiver is under a bridge. In general, however, it is not required that the first and second portion of output bits comprise more bits than the first number of input bits, since both portions together still have a code rate of 1/2, whereas each portion of output bits **410**, **420** has a code rate of 1 when the convolutional encoder **120** has a code rate of 1/2.

Referring back to FIG. **4**, a convolutional encoder having a code rate of 3/8 is described. This means that for a first number of input bits, the first number being **3**, a second number of output bits, the second number being **8**, is produced.

Referring to FIG. **5**, a preferred embodiment of the receiving apparatus is described. Optionally, the receiving means comprises a QPSK demodulator **240***d* for receiving the first channel and a QPSK demodulator **240***e* for receiving the second channel. Naturally, the QPSK demodulators **240***d* and **240***e* only have to be provided when the transmitting apparatus has performed a QPSK modulation. The output of the QPSK demodulator is fed into the delay stage **240***c* and then into a multiplexer **230***a*. The output of the QPSK demodulator **240***e* is directly fed into the multiplexer **230***a*. Thus, the multiplexer **230***a* receives the first portion of output bits (**410** in FIG. **4**) and the second portion of output bits (**420** in FIG. **4**) for producing one single serial bitstream comprised of the first portion and the second portion. This continuous serial bitstream is input into a depuncturing unit

230*b* for undoing the puncturing carried out by the puncturing unit **125** (FIG. **3**).

Then, the depunctured bitstream, i.e., the combined bitstream output by the combiner that comprises the multiplexer **230***a* and the depuncturing unit **230***b*, is input into the channel decoder that, in accordance with the preferred embodiment of the present invention, comprises the Viterbi decoder **220***a* and a Reed-Solomon decoder **220***b*. Those skilled in the art will know that the Reed-Solomon decoder only has to be provided when a Reed-Solomon coding has been carried out in the transmitting apparatus. In accordance with the preferred embodiment of the present invention, the transmitting apparatus causes a concatenated forward error correction encoder having a convolutional encoder and a Reed-Solomon encoder. Thus, the receiving apparatus has to comprise a Viterbi decoder **220***a* and a Reed-Solomon decoder **220***b*. It is known in the art that convolutional encoders may create small burst errors. The Reed-Solomon encoder, however, is well suited for such burst errors.

In the following, the inventive transmitting receiving system illustrated in FIG. **1**, which makes use of an encoder having a code rate less than 1/2, and has a partitioner and a combiner is compared to the transmission receiving system shown in FIG. **7** that makes use of a duplicator and a channel decision controlled switch.

To ease the comparison of both systems, it is assumed that the encoders **120** (FIG. **1**) and **64** (FIG. **7**) comprise a convolutional encoder and a Reed-Solomon encoder. Furthermore, it is assumed that the convolutional encoder included in the redundancy adding encoder **120** of FIG. **1** implements a code rate of 3/8, whereas the convolutional encoder included in the redundancy adding encoder **64** of FIG. **7** encodes based on a code rate of 3/4. Since the transmitting apparatus shown in FIG. **7** transmits eight output bits for three input bits, i.e., the duplicator effectively doubles the output bits to be transmitted, it can be regarded as a redundancy adding encoder having a code rate of 3/8. The signals transmitted over the first and the second channels, however, are identical and identically derived from the bitstream output by the bitstream source **62**.

According to the code rate of 3/4 (convolutional coder only), for three information bits, four channel bits are transmitted over each satellite in the system of FIG. **7** Using two satellites, eight channel bits are transmitted for three information bits. Thus, it is clear that the system shown in FIG. **7** can be regarded as a system having a code rate of 3/8.

According to the literature and system simulation results, the following E_b/N_o performance can be assumed. It is noted that the term E_b/N_o represents the ratio of the energy per useful bit rate (unit factor: W/sec) to noise power density (unit facor: W/sec). Thus, the "unit factor" of E_b/N_o is 1. For QPSK, the C/N (=Power of transmitted signal/noise power within effective bandwidth) can be calculated by the following equation:

$$C/N=E_b/N_o+10\cdot\log(R)+3\ \mathrm{dB}(C/N \text{ and } E_b/N_o \text{ values in dB})$$

In this equation, R is the code rate. It is to be noted that the term C/N represents the "link margin". If the real C/N is higher than the link margin, a useful communication link is obtained. If the real C/N is lower than the C/N defined by the above-outlined equation, no satisfying communication link can be established.

The following Table gives the C/N in dB for three different code rates. The first line of the Table relates to the system shown in FIG. **7**, whereas the third line of the Table relates to the inventive system shown in FIG. **1**. The factor

(223/255) relates to the Reed-Solomon encoder. The factors 3/4, 1/2 and 3/8 relate to the convolutional coders "Rate".

TABLE

Code rate (convol. + Reed-Solomon)	Eb/NO [dB]	C/N [dB]
3/4 * (223/255) = 0.66	3.7	4.9
1/2 * (223/255) = 0.44	2.7	2.1
3/8 * (223/255) = 0.33	app. 2.4	0.6

In the following, the Table is explained. The required C/N value also applies to a system where the bitstream is de-multiplexed to two streams and transmitted using two QPSK modulators. The overall transmitted power is defined as:

$$C=C_{sat1}+C_{sat2}.$$

The noise power is defined as follows:

$$N=N_1+N_2.$$

It is assumed that the signal power is identical for satellites **1** and **2**, which is the best case for the inventive method. With $N_1=N_2$ and $C_{Sat1}=C_{Sat2}$ (the effective bandwidth is identical for both signals, i.e., channels), the following equation applies:

$$\frac{C}{N}=\frac{C_{Sat1}+C_{Sat2}}{N_1+N_2}=\frac{2C_{Sat1}}{2N_1}=\frac{C_{Sat1}}{N_1}=\frac{C_{Sat2}}{N_2}$$

Assuming the available signal power (=C) and the QPSK symbol rate are kept, the method in accordance with the present invention can give a gain of 4.3 dB compared to the required C/N, if one signal is decoded only (system of FIG. **7**). It is assumed that for other scenarios, i.e., both channels show different fading characteristics, the gain is lower. At least for the scenario C_{Sat1}/N or C_{Sat2}/N being greater than 4.9 dB, no gain is required. The output signal is error free in any case. The overall gain of the inventive transmission receiving system depends on the probability of the scenario. In other words, it is possible to receive the signal down to a C/N of 0.6 dB, which is a theoretical value that does not include implementation loss. If only one satellite signal is available, the required C/N_o is equal to 67 dBHz (not including implementation loss). The unit factor dBHz represents power divided by power density in logarithmical terms.

As it has been described with respect to FIGS. **3** and **4**, a convolutional encoder with a code rate of 1/3 is preferred The output of the convolutional enocoder is punctured to a code rate of 3/8 by not transmitting one channel bit out of 9. The output of the convolutional encoder and puncturing unit is converted into a serial form and demultiplexed. Four bits out of 8 are transmitted over satellite **1**, i.e., the first portion of output bits. The other four bits are transmitted over satellite **2**, i.e., the second portion of output bits. Optionally, an additional time interlever can be used.

As it has been outlined in the introductory portion, the polynominals g1, g2 and g3 describe the shift registers and modulo-2 adders or XOR gates which generate the convolutional code having a code rate of 1/3. The proposed polynominals are as follows:

g1=1100111 (binary)=147 (octal)

g2=1011101 (binary)=135 (octal)

g3=1110011 (binary)=163 (octal)

It should be noted that generator polynomials different from the above mentioned generator polynomials combined with certain puncturing schemes may be used as well (see J. Bibb Cain supra). However, the above given generator polynomials work very well in connection with the puncturing scheme described herein.

The receiver shown in FIG. **5** requires one Viterbi decoder only. The optimal combining with respect to the signal quality of the two signals is automatically performed by the Viterbi decoder. The Viterbi decoder performs maximum likelihood decoding using the channel state information, also called "metric". Algorithms known for Rician and Rayleigh channels can be adapted. If only one signal is available, i.e., one channel is faded totally, the input of the Viterbi decoder can be considered as a convolutional encoder having a code rate of 1/3 punctured to a code rate of 3/4. The equivalent puncturing scheme is:

For the early satellite

111

100

000

and for the late satellite

000

001

111

In accordance with a preferred embodiment of the present invention, a Viterbi decoder implementing a soft decision based on probabilities is used. Thus, the depuncturing unit inserts probabilities rather than actual bit states. Since the depuncturing unit does not have any information about the bits punctured by the puncturing unit in the transmitting apparatus, it inserts probabilities of 0.5 for the low and the high states of the bits.

Optionally, the combiner **230** (FIG. **1**) additionally comprises a channel estimator, that evaluates the signal to noise ratio of signals received from each channel. When the channel estimator determines low signal to noise ratios, it is adapted to insert 0.5 probabilities rather than the actual probabilities derived from the channel having a low signal to noise ratio. Thus, it can be ascertained that the maximum likelihood decoder is not misled by signals received via a channel having a low signal to noise ratio.

Although the preferred embodiment of the present invention has been described with respect to two channels, the inventive concept can also be applied to a transmission system comprising three or more channels. In the case of three channels, the code rate of the channel encoder **120** (FIG. **1**) is to be 1/3 or less. Additionally, the partitioner produces three portions of output bits rather than two portions of output bits. In this case, the transmitting apparatus as described in FIG. **3** can be used. However, no puncturing unit is necessary and the demultiplexer multiplexes three bitstreams, one for each of the three channels. After reading this specification, it is obvious for those skilled in the art that the present invention even can be extended to four or more channels.

Although the preferred embodiment of the present invention uses a convolutional encoder which is optionally extended by means of a Reed-Solomon encoder, other redundancy adding encoders can be adapted. These redundancy adding encoders, however, have to produce two portions of output bits that are coded differently with respect to each other, such that a "real" code rate of, for example, 3/8 in contrast to a doubled 3/4 code rate can be obtained.

The delay imposed by the different delay stages can be set in accordance with the real environment. Normally, a delay of four seconds is regarded as appropriate. However, other delay values can be adapted. It is to be noted, however, that high delay values result in high memory capacities for the transmitter and the receiver.

What is claimed is:

1. An apparatus for transmitting information, comprising:

a bitstream source for providing a bitstream representing the information;

a redundancy adding encoder for generating an encoded bitstream based on the bitstream provided by the bit-stream source wherein the encoder is arranged to output, for a first number of input bits, a second number of output bits, the second number of output bits having at least twice as many output bits as the first number of input bits, and wherein the second number of output bits includes two portions of output bits, each portion of output bits individually allowing the retrieval of information represented by the first number of input bits, and the first portion of output bits being coded based on the bitstream in a different way with respect to the second portion of output bits;

a partitioner for partitioning the second number of output bits into the two portions of output bits;

a transmitter for transmitting the output bits of the first portion via a first channel and the output bits of the second portion via a second channel, the second channel being spatially different from the first channel;

the transmitter being a single transmitter;

the first channel being defined by the single transmitter and a first position of a mobile receiver;

the second channel being defined by the single transmitter and a second position of the mobile receiver; and

the transmitter further includes delay means for delaying the second portion of output bits transmitted via the second channel such that time diversity is obtained.

2. An apparatus for transmitting information, comprising:

a bitstream source for providing a bitstream representing the information;

a redundancy adding encoder for generating an encoded bitstream based on the bitstream provided by the bit-stream source wherein the encoder is arranged to output, for a first number of input bits, a second number of output bits, the second number of output bits having at least twice as many output bits as the first number of input bits, and wherein the second number of output bits includes two portions of output bits, each portion of output bits individually allowing the retrieval of information represented by the first number of input bits, and the first portion of output bits being coded based on the bitstream in a different way with respect to the second portion of output bits;

a partitioner for partitioning the second number of output bits into the two portions of output bits; and

means for transmitting the output bits of the first portion via a first channel and the output bits of the second portion via a second channel, the second channel being spatially different from the first channel;

the means for transmitting including a first transmitter and a second transmitter spaced apart from the first transmitter;

the first channel being defined by the first transmitter and the receiver; and

the second channel being defined by the second transmitter and the receiver such that space diversity is obtained.

3. The apparatus of claim 2, in which:

the first and second transmitters include two satellites in different orbital positions, such that the first channel is defined by an uplink connection from earth to the first satellite and a downlink connection from the first satellite to a receiver on earth, and such that the second channel is defined by a uplink connection from earth to the second satellite and a downlink connection from the second satellite to the receiver on earth.

4. The apparatus of claim 2, in which:

one transmitter includes a satellite; and

the other transmitter includes a terrestrial sender such that terrestrial diversity is obtained.

5. The apparatus of claim 2, in which the transmitter further includes delay means for delaying the second portion of output bits transmitted via the second channel such that time diversity is obtained.

6. An apparatus for transmitting information, comprising:

a bitstream source for providing a bitstream representing the information;

a redundancy adding encoder for generating an encoded bitstream based on the bitstream provided by the bit-stream source wherein the encoder is arranged to output, for a first number of input bits, a second number of output bits, the second number of output bits having at least twice as many output bits as the first number of input bits, and wherein the second number of output bits includes two portions of output bits, each portion of output bits individually allowing the retrieval of information represented by the first number of input bits, and the first portion of output bits being coded based on the bitstream in a different way with respect to the second portion of output bits, the redundancy adding encoder including a convolutional encoder for obtaining a code rate less than or equal to 0.5, wherein the code rate is the ratio of the first number of input bits to the second number of output bits, the convolutional encoder combining a current input bit to be encoded with at least one of a certain number of preceding input bits;

a partitioner for partitioning the second number of output bits into the two portions of output bits; and

a transmitter for transmitting the output bits of the first portion via a first channel and the output bits of the second portion via a second channel, the second channel being spatially different from the first channel.

7. The apparatus of claim 6, in which:

the certain number of preceding bits is 6; and

the convolutional encoder comprises three generator polynominals g_1, g_2 and g_3 having the following binary form:

g_1=1100111,

g_2=1011101, and

g_3=1110011.

8. The apparatus of claim 6 which further comprises a puncturing unit operative to discard at least one predetermined bit of the encoded bitstream such that the second number of output bits is an even number, wherein the first and second portions of output bits comprise the same number of output bits.

9. An apparatus for transmitting information, comprising:

a bitstream source for providing a bitstream representing the information;

a redundancy adding encoder for generating an encoded bitstream based on the bitstream provided by the bitstream source wherein the encoder is arranged to output, for a first number of input bits, a second number of output bits, the second number of output bits having at least twice as many output bits as the first number of input bits, and wherein the second number of output bits includes two portions of output bits, each portion of output bits individually allowing the retrieval of information represented by the first number of input bits, and the first portion of output bits being coded based on the bitstream in a different way with respect to the second portion of output bits;

a partitioner for partitioning the second number of output bits into the two portions of output bits; and

a transmitter for transmitting the output bits of the first portion via a first channel and the output bits of the second portion via a second channel, the second channel being spatially different from the first channel;

the redundancy adding encoder is operative to code the bitstream provided by the bitstream source in a bit-by-bit fashion;

the partitioner includes a parallel storage for storing a predetermined amount of output bits of the convolutional encoder;

a parallel-to-serial converter for producing a serial stream of the stored bits to be partitioned into the first and second portion of output bits is provided; and

a de-multiplexer for performing the partition of the serial stream of output bits into the first and second portions is provided.

10. An apparatus for receiving information, the information being represented by an encoded bitstream, the encoded bitstream being encoded such that its redundancy is at least doubled with respect to a bitstream from which the encoded bitstream is derived, and that, for a first number of bits of the bitstream, the encoded bitstream comprises a second number of bits, the second number of bits having at least twice as many bits as the first number, and wherein the second number of bits includes two portions of bits, each portion of bits individually allowing the retrieval of information represented by the first number of bits, and the first portion of the bits being encoded in a different way with respect to the second portion of bits, the apparatus comprising:

receiving means for receiving the first portion of bits via a first channel and the second portion of bits via a second channel, the first and the second channels being spatially different from each other;

a combiner for combining the first and the second portions the combiner including a depuncturing unit for performing a depuncturing operation on the first and second portions of bits to compensate for a puncturing operation performed in a transmitter; and

a decoder for decoding the coded bitstream by removing redundancy from the coded bitstream, the decoder using the first and second portions of bits combined by the combiner.

11. The apparatus of claim **10**, in which the receiving means further includes delay means for delaying the portion of bits received via one channel to compensate for a delay imposed on the portion of bits received via the other channel.

12. The apparatus of claim **11**, in which the combiner includes a multiplexer for multiplexing first and second portions into a form suitable for the decoder.

13. An apparatus for receiving information, the information being represented by an encoded bitstream, the encoded bitstream being encoded such that its redundancy is at least doubled with respect to a bitstream from which the encoded bitstream is derived, and that, for a first number of bits of the bitstream, the encoded bitstream comprises a second number of bits, the second number of bits having at least twice as many bits as the first number, and wherein the second number of bits includes two portions of bits, each portion of bits individually allowing the retrieval of information represented by the first number of bits, and the first portion of the bits being encoded in a different way with respect to the second portion of bits, the apparatus comprising:

receiving means for receiving the first portion of bits via a first channel and the second portion of bits via a second channel, the first and the second channels being spatially different from each other;

a combiner for combining the first and the second portions;

a decoder for decoding the coded bitstream by removing redundancy from the coded bitstream, the decoder using the first and second portions of bits combined by the combiner,

the decoder comprising a soft decision decoder processing probabilities in that a received bit represents a high or low state rather than an actual wave form characteristic of the received bitstream; and

depuncturing means for compensating for a puncturing operation in a transmitter is provided and attributes to a bit to be depunctured equal probabilites for the high and low states.

14. The apparatus of claim **13** in which the decoder includes a Viterbi decoder performing maximum likelihood decoding using the state information of the first and second channels.

15. The apparatus of claim **14** in which the decoder further comprises a Reed-Solomon decoder fed by the Viterbi decoder for undoing a Reed-Solomon encoding performed in the transmitter.

16. The apparatus of claim **13** in which:

the decoder comprises a signal to noise ratio evaluating means for determining a channel having a low signal to noise ratio; and

a bit replacing means for replacing the bits of a portion of bits received via a channel having a low signal to noise ratio by values equivalent to a lower reliability for the high and low states.

17. The apparatus of claim **13** in which the receiving means comprises, for each channel, a QPSK demodulator for providing the first and the second portions of bits.

18. A method of transmitting information, comprising the following steps:

providing a bitstream representing the information;

generating a redundancy added encoded bitstream based on the bitstream provided in the step of providing, wherein for a first number of input bits, a second number of output bits is generated, the second number of output bits having at least twice as many output bits as the first number of input bits, and wherein the second number of output bits includes two portions of output bits, each portion of output bits individually allowing the retrieval of information represented by the first number of input bits, and the first portion of output bits being coded based on the bitstream in a different way with respect to the second portion of output bits;

partitioning the second number of output bits into the two portions of output bits; and

transmitting by use of a single transmitter the output bits of the first portion via a first channel and the output bits of the second portion via a second channel, the second channel being spatially different from the first channel;

the first channel being defined by the single transmitter and a first position of a mobile receiver;

the second channel being defined by the single transmitter and a second position of the mobile receiver; and

the step of transmitting comprising the following sub-steps: transmitting the first portion of output bits; and delaying the second portion of output bits before transmitting via the second channel such that time diversity is obtained.

19. A method of transmitting information, comprising the following steps:

providing a bitstream representing the information;

generating a redundancy added encoded bitstream based on the bitstream provided in the step of providing, wherein for a first number of input bits, a second number of output bits is generated, the second number of output bits having at least twice as many output bits as the first number of input bits, and wherein the second number of output bits includes two portions of output bits, each portion of output bits individually allowing the retrieval of information represented by the first number of input bits, and the first portion of output bits being coded based on the bitstream in a different way with respect to the second portion of output bits;

partitioning the second number of output bits into the two portions of output bits; and transmitting the output bits of the first portion via a first channel and the output bits of the second portion via a second channel, the second channel being spatially different from the first channel

the step of transmitting being carried out by a first transmitter and a second transmitter spaced apart from the first transmitter;

the first channel being defined by the first transmitter and the receiver; and

the second channel being defined by the second transmitter and the receiver such that space diversity is obtained.

20. The method of claim **19**, in which:

the first and second transmitters include two satellites in different orbital positions, such that the first channel is defined by an uplink connection from earth to the first satellite and a downlink connection from the first satellite to a receiver on earth, and such that the second channel is defined by a uplink connection from earth to the second satellite and a downlink connection from the second satellite to the receiver on earth.

21. The method of claim **19**, in which:

one transmitter includes a satellite; and

the other transmitter includes a terrestrial sender such that terrestrial diversity is obtained.

22. The method of claim **19**, in which the step of transmitting further includes the following substeps: delaying the second portion of output bits transmitted via the second channel such that time diversity is obtained.

23. A method of transmitting information, comprising the following steps:

providing a bitstream representing the information;

generating a redundancy added encoded bitstream based on the bitstream provided in the step of providing, wherein for a first number of input bits, a second number of output bits is generated, the second number of output bits having at least twice as many output bits as the first number of input bits, and wherein the second number of output bits includes two portions of output bits, each portion of output bits individually allowing the retrieval of information represented by the first number of input bits, and the first portion of output bits being coded based on the bitstream in a different way with respect to the second portion of output bits;

the generating step being carried out by means of a convolutional encoder for obtaining a code rate less than or equal to 0.5, wherein the code rate is the ratio of the first number of input bits to the second number of output bits, the convolutional encoder combining a current input bit to be encoded with at least one of a certain number of preceding input bits;

partitioning the second number of output bits into the two portions of output bits; and

transmitting the output bits of the first portion via a first channel and the output bits of the second portion via a second channel, the second channel being spatially different from the first channel.

24. The method of claim **23**, in which:

the certain number of preceding bits is 6; and

the convolutional encoder comprises three generator polynominals g_1, g_2 and g_3 having the following binary form:

g_1=1100111,

g_2=1011101, and

g_3=1110011.

25. The method of claim **23** which further comprises the step of puncturing to discard at least one predetermined bit of the encoded bitstream such that the second number of output bits is an even number, wherein the first and second portions of output bits comprise the same number of output bits.

26. A method of transmitting information, comprising the following steps:

providing a bitstream representing the information;

generating a redundancy added encoded bitstream based on the bitstream provided in the step of providing, by coding the bitstream provided by the bitstream source in a bit-by-bit fashion, wherein for a first number of input bits, a second number of output bits is generated, the second number of output bits having at least twice as many output bits as the first number of input bits, and wherein the second number of output bits includes two portions of output bits, each portion of output bits individually allowing the retrieval of information represented by the first number of input bits, and the first portion of output bits being coded based on the bitstream in a different way with respect to the second portion of output bits;

producing a serial stream of the stored bits to be partitioned into the first and second portion of output bits by the step of parallel-to-serial converting;

partitioning the second number of output bits into the two portions of output bits by de-multiplexing; and

transmitting the output bits of the first portion via a first channel and the output bits of the second portion via a second channel, the second channel being spatially different from the first channel.

27. A method of receiving information, the information being represented by an encoded bitstream, the encoded bitstream being encoded such that its redundancy is at least doubled with respect to a bitstream from which the encoded bitstream is derived, and that, for a first number of bits of the bitstream, the encoded bitstream comprises a second number of bits, the second number of bits having at least twice as many bits as the first number, and wherein the second number of bits includes two portions of bits, each portion of bits individually allowing the retrieval of information represented by the first number of bits, and the first portion of the bits being encoded in a different way with respect to the second portion of bits, the method comprising the following steps:

receiving the first portion of bits via a first channel and the second portion of bits via a second channel, the first and the second channels being spatially different from each other;

combining the first and the second portions;

performing a depuncturing operation on the first and second portions of bits to compensate for a puncturing operation performed in a transmitter; and

decoding the coded bitstream by removing redundancy from the coded bitstream, wherein the first and second portions of bits combined in the step of combining are used in the step of decoding.

28. The method of claim **27**, in which the step of receiving further includes the step of delaying the portion of bits received via one channel to compensate for a delay imposed on the portion of bits received via the other channel.

29. The method of claim **27**, in which the step of combining includes the step of multiplexing the first and second portions into a form suitable for the step of decoding.

30. The method of claim **27**, in which the step of receiving comprises, for each channel, the step of QPSK demodulating for providing the first and the second portions of bits.

31. A method of receiving information, the information being represented by an encoded bitstream, the encoded bitstream being encoded such that its redundancy is at least doubled with respect to a bitstream from which the encoded bitstream is derived, and that, for a first number of bits of the bitstream, the encoded bitstream comprises a second number of bits, the second number of bits having at least twice as many bits as the first number, and wherein the second number of bits includes two portions of bits, each portion of bits individually allowing the retrieval of information represented by the first number of bits, and the first portion of

the bits being encoded in a different way with respect to the second portion of bits, the method comprising the following steps:

receiving the first portion of bits via a first channel and the second portion of bits via a second channel, the first and the second channels being spatially different from each other;

combining the first and the second portions; and

decoding the coded bitstream by removing redundancy from the coded bitstream by soft decision decoding to process probabilities such that a received bit represents a high or low state rather than an actual wave form characteristic of the received bitstream, wherein the first and second portions of bits combined in the step of combining are used in the step of decoding; and

a step of depuncturing for compensating for a puncturing operation in a transmitter is carried out and attributes to a bit to be depunctured equal probabilites for the high and low states.

32. The method of claim **31** in which the step of decoding includes a Viterbi decoder performing maximum likelihood decoding using the state information of the first and second channels.

33. The method of claim **32** in which the step of decoding further comprises a Reed-Solomon decoder fed by the Viterbi decoder for undoing a Reed-Solomon encoding performed in the transmitter.

34. The method of claim **31**, in which:

the step of decoding includes the step of signal to noise ratio evaluating for determining a channel having a low signal to noise ratio; and

the step of bit replacing for replacing the bits of a portion of bits received via a channel having a low signal to noise ratio by equal probabilities for the high and low states is carried out.

35. The method of claim **31**, in which the step of receiving comprises, for each channel, the step of QPSK demodulating for providing the first and second portions of bits.

* * * * *

Exhibit B

US006931084B1

(12) **United States Patent**
Eberlein et al.

(10) **Patent No.: US 6,931,084 B1**
(45) **Date of Patent: Aug. 16, 2005**

(54) **DIFFERENTIAL CODING AND CARRIER RECOVERY FOR MULTICARRIER SYSTEMS**

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(73) Assignee: **Fraunhofer-Gesellschaft zur Foerderung der angewandten Forschung e.V.**, Munich (DE)

(*) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 0 days.

(21) Appl. No.: **09/673,266**

(22) PCT Filed: **Apr. 14, 1998**

(86) PCT No.: **PCT/EP98/02167**

§ 371 (c)(1), (2), (4) Date: **Nov. 29, 2000**

(87) PCT Pub. No.: **WO99/53664**

PCT Pub. Date: **Oct. 21, 1999**

(51) **Int. Cl.**[7] **H03D 1/04**; H03D 1/06; H03K 5/01; H03K 6/04; H04B 1/10; H04L 1/00

(52) **U.S. Cl.** **375/346**

(58) **Field of Search** 375/346; 370/32.1; 379/410

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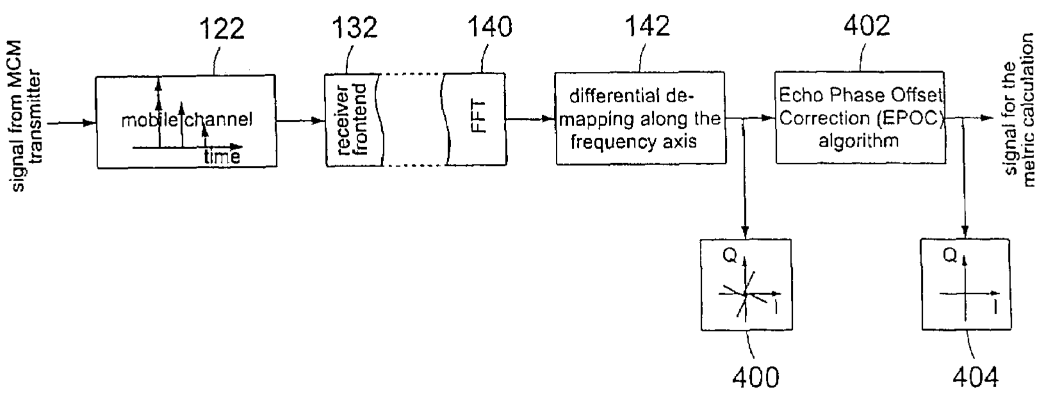
Primary Examiner—Stephen Chin
Assistant Examiner—Harry Vartanian
(74) *Attorney, Agent, or Firm*—Roylance, Abrams, Berdo & Goodman, L.L.P.

(57) **ABSTRACT**

A method of performing an echo phase offset correction in a multi-carrier demodulation system involves the step of differential phase decoding phase shifts based on a phase difference between simultaneous carriers having different frequencies. An echo phase offset is determined for each decoded phase shift by eliminating phase shift uncertainties related to the transmitted information from the decoded phase shift. The echo phase offsets are averaged in order to generate an averaged offset. Finally, each decoded phase shift is corrected based on the averaged offset.

16 Claims, 7 Drawing Sheets



Fraunhofer Ex 2002-39
Sirius XM v Fraunhofer, IPR2018-00689

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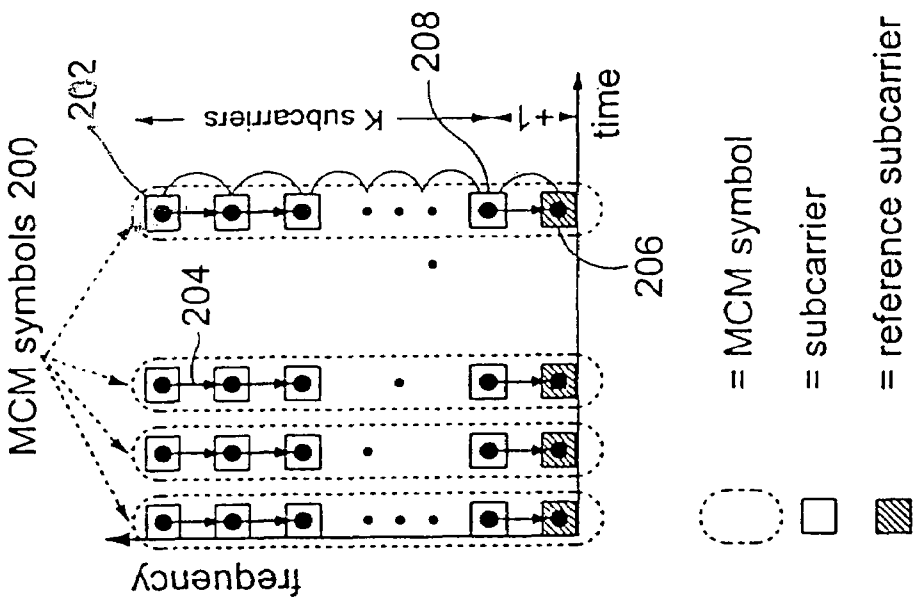
MCM symbols 200
202
K subcarriers
208
time
frequency
204
206
= MCM symbol
= subcarrier
= reference subcarrier

FIG.1

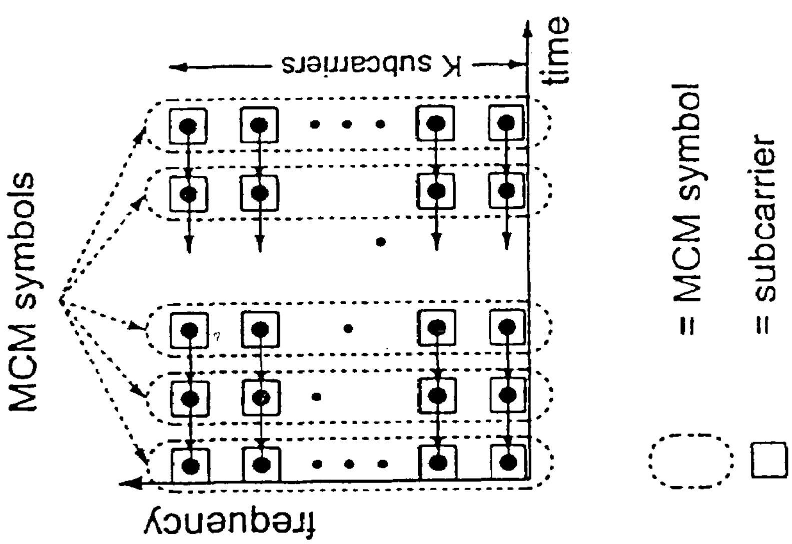
MCM symbols
K subcarriers
time
frequency
= MCM symbol
= subcarrier

FIG.8

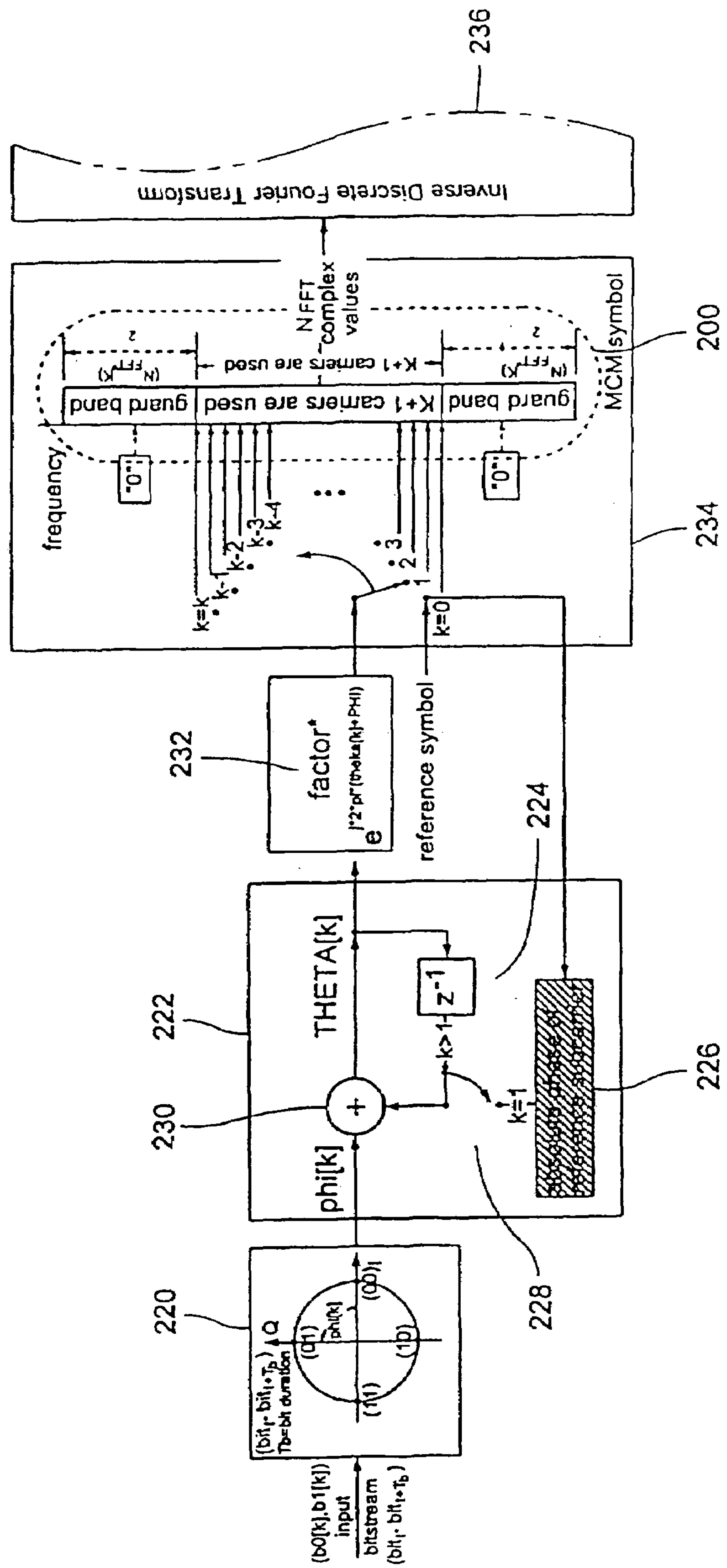
236
Inverse Discrete Fourier Transform
NFFT complex values
MCM symbol
200
guard band
K+1 carriers are used
frequency
"0"
234
232
factor* $e^{j*2*pi*(theta[k]+PHI)}$
reference symbol
k=0
224
THETA[k]
z^{-1}
222
230
phi[k]
k>1
k=1
226
228
220
input bitstream
Q

FIG.2

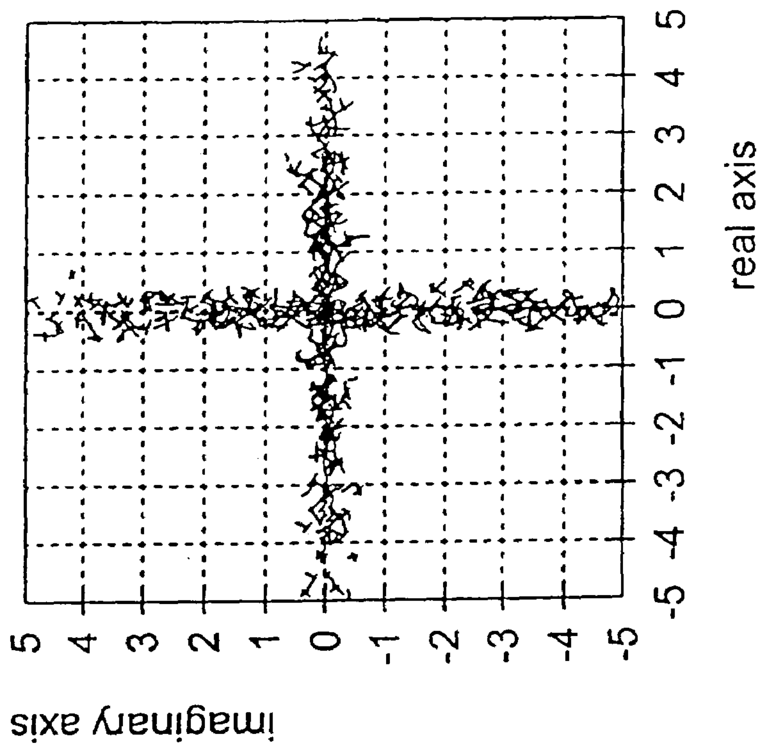
real axis
imaginary axis
-5 -4 -3 -2 -1 0 1 2 3 4 5
5 4 3 2 1 0 -1 -2 -3 -4 -5

FIG.3B

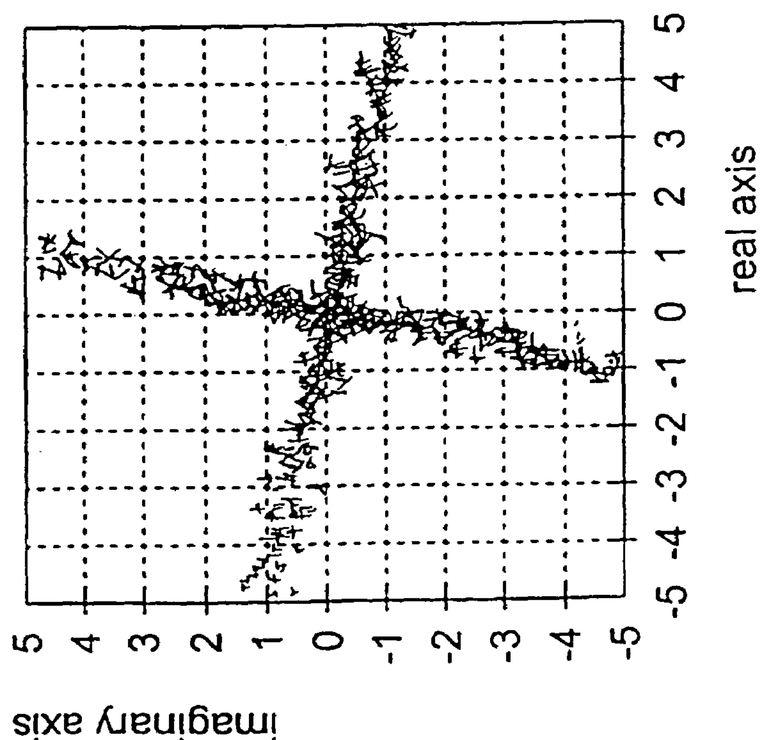
real axis
imaginary axis
-5 -4 -3 -2 -1 0 1 2 3 4 5
5 4 3 2 1 0 -1 -2 -3 -4 -5

FIG.3A

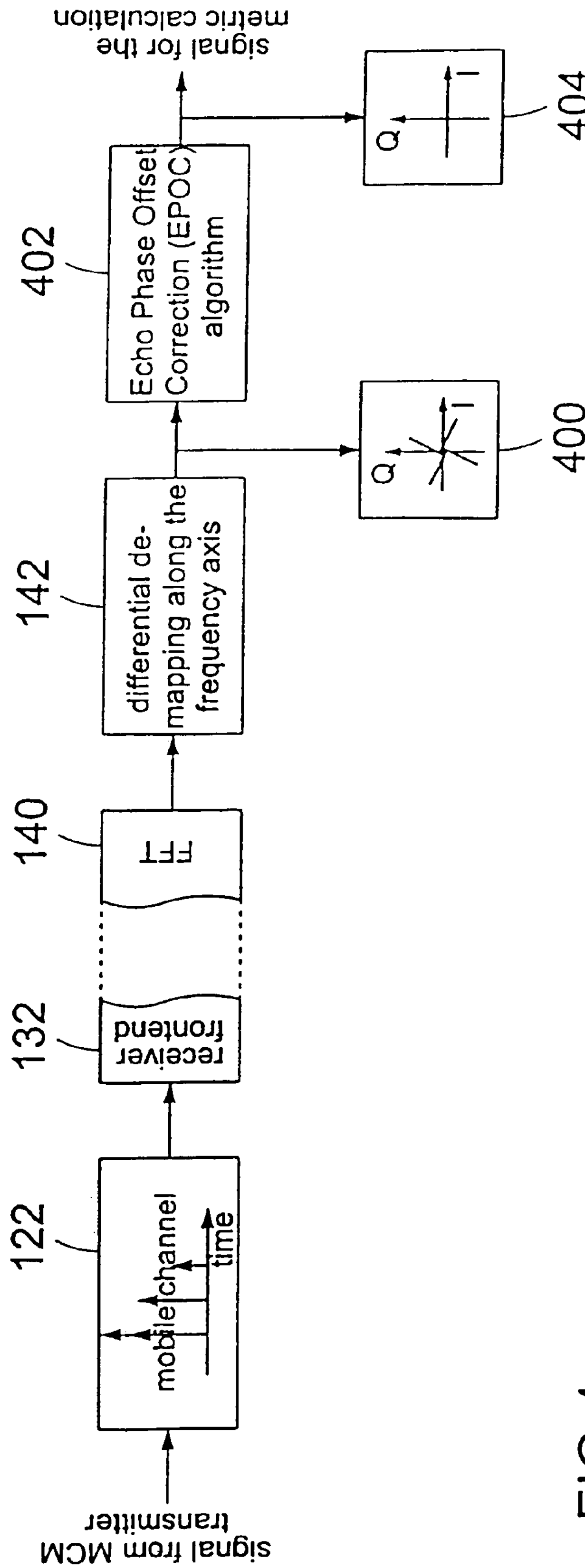
signal from MCM transmitter
122
mobile channel
time
132
receiver frontend
140
FFT
142
differential de-mapping along the frequency axis
400
Q
I
402
Echo Phase Offset Correction (EPOC) algorithm
404
Q
I
signal for the metric calculation

FIG.4

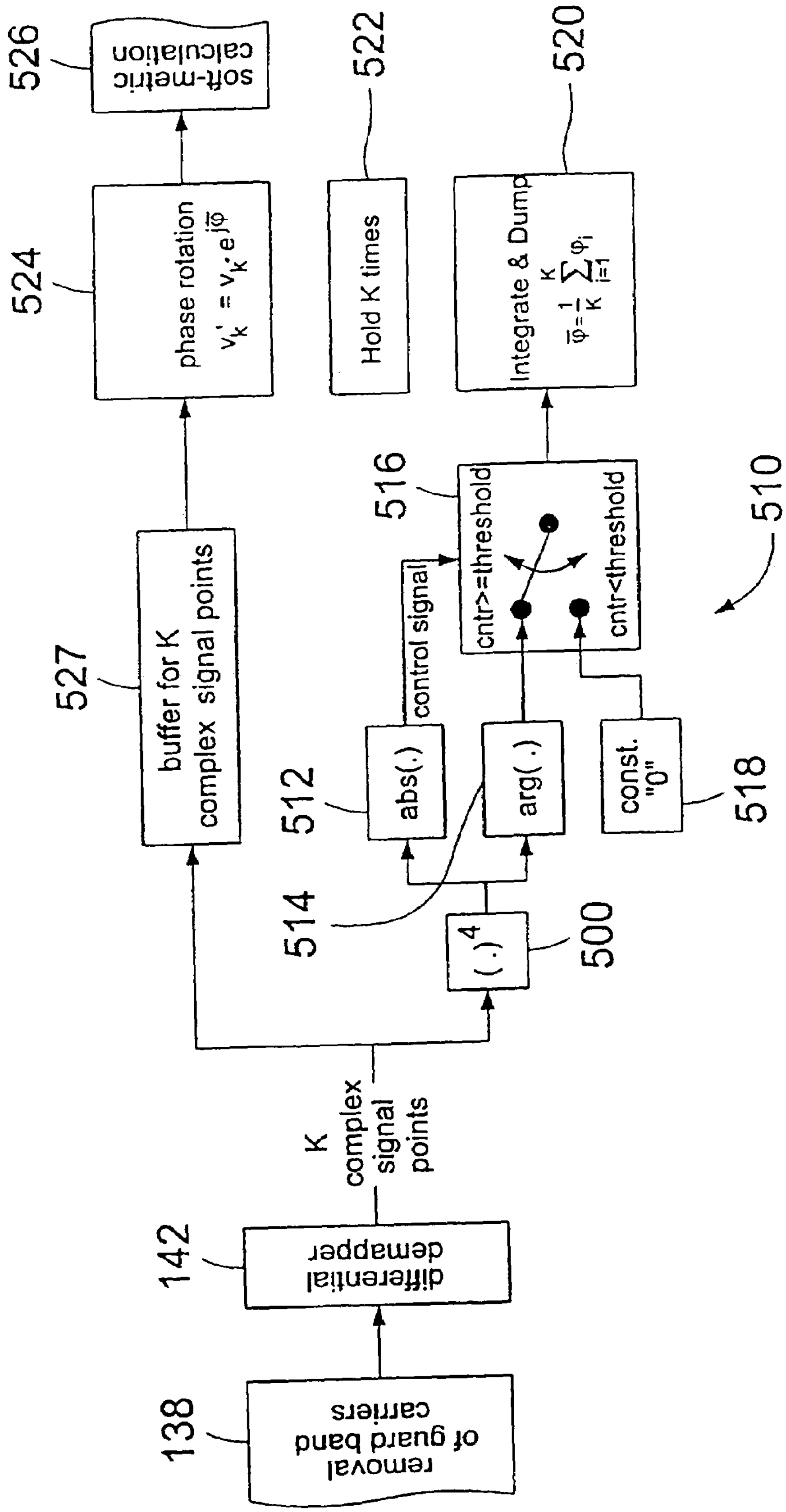
138
removal of guard band carriers
142
differential demapper
K complex signal points
500
(.)4
514
512
abs(.)
arg(.)
const. "0"
518
control signal
516
cntr>=threshold
cntr<threshold
510
527
buffer for K complex signal points
524
phase rotation
v'k = vk · e^(jφ̄)
522
Hold K times
520
Integrate & Dump
φ̄ = (1/K) Σ(i=1..K) φi
526
soft-metric calculation

FIG.5

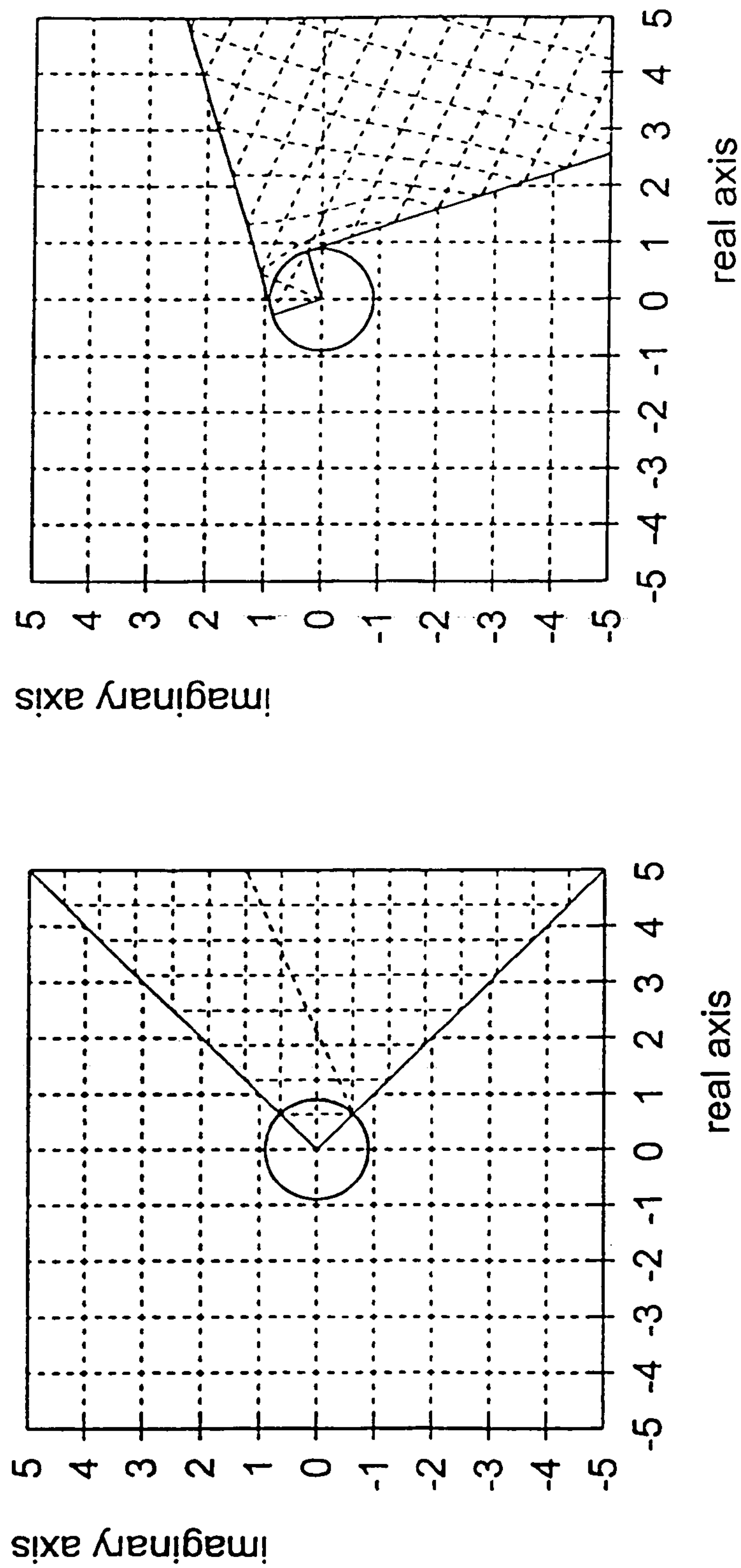

FIG.6

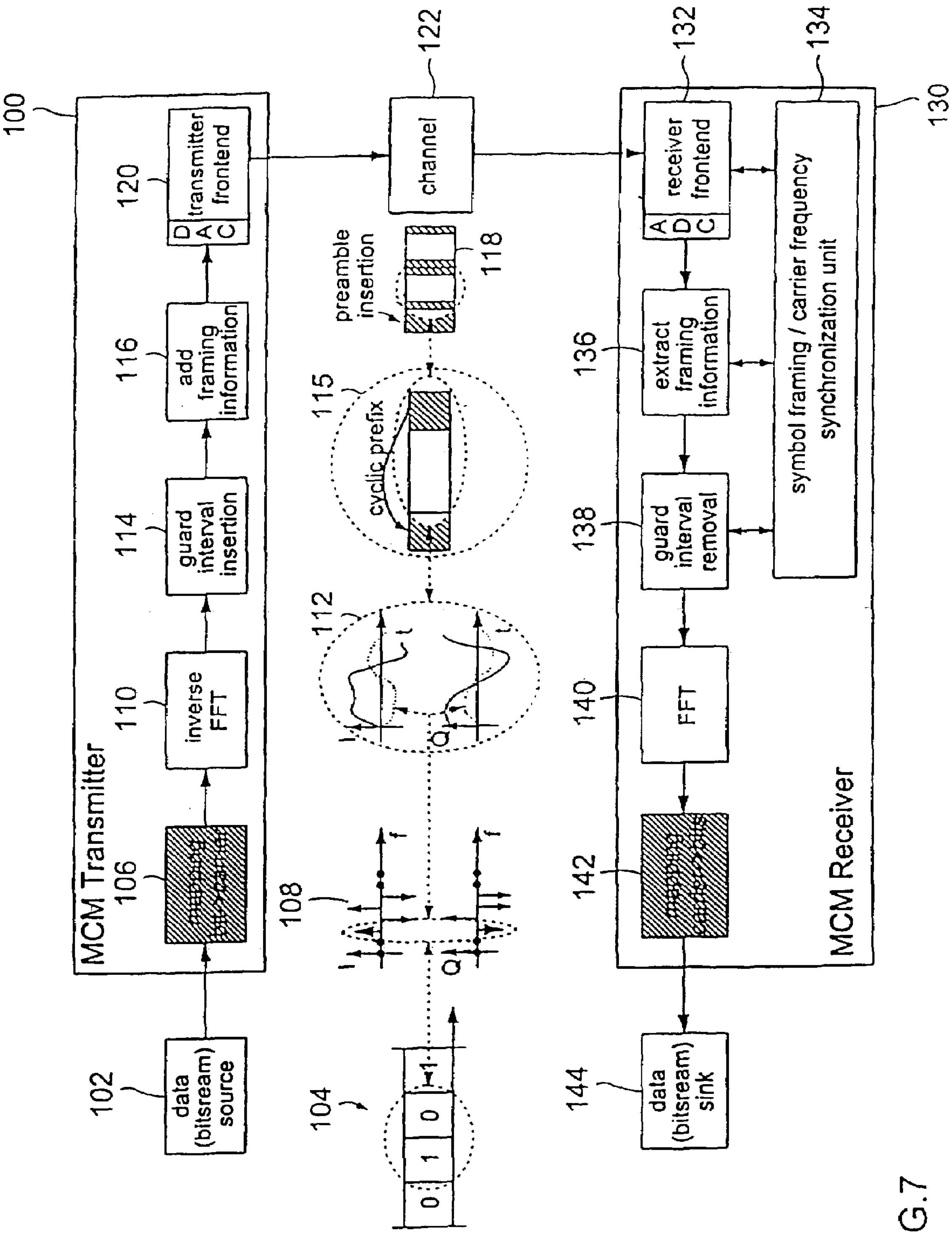
100
MCM Transmitter
102
data (bitsream) source
106
110
inverse FFT
114
guard interval insertion
116
add framing information
120
D A C transmitter frontend
104
0 1 0 -1
108
I
Q
f
112
I
Q
t
115
cyclic prefix
preamble insertion
118
122
channel
130
MCM Receiver
132
A D C receiver frontend
134
symbol framing / carrier frequency synchronization unit
136
extract framing information
138
guard interval removal
140
FFT
142
144
data (bitsream) sink

FIG.7

DIFFERENTIAL CODING AND CARRIER RECOVERY FOR MULTICARRIER SYSTEMS

FIELD OF THE INVENTION

The present invention relates to methods and apparatus for performing modulation and de-modulation in multi-carrier modulation systems (MCM systems) and, in particular, to methods and apparatus for differential mapping and de-mapping of information onto carriers of multi-carrier modulation symbols in such systems. Furthermore, the present invention relates to methods and apparatus for performing an echo phase offset correction when decoding information encoded onto carriers of multi-carrier modulation symbols in multi-carrier modulation systems.

BACKGROUND OF THE INVENTION

The present invention generally relates to broadcasting of digital data to mobile receivers over time-variant multipath channels. More specifically, the present invention is particularly useful in multipath environments with low channel coherence time, i.e. rapidly changing channels. In preferred embodiments, the present invention can be applied to systems implementing a multicarrier modulation scheme. Multi-carrier modulation (MCM) is also known as orthogonal frequency division multiplexing (OFDM).

In a MCM transmission system binary information is represented in the form of a complex spectrum, i.e. a distinct number of complex subcarrier symbols in the frequency domain. In the modulator a bitstream is represented by a sequence of spectra. Using an inverse Fourier-transform (IFFT) a MCM time domain signal is produced from this sequence of spectra.

FIG. **7** shows a MCM system overview. At **100** a MCM transmitter is shown. A description of such a MCM transmitter can be found, for example, in William Y. Zou, Yiyan Wu, "COFDM: AN OVERVIEW", IEEE Transactions on Broadcasting, vol. 41, No. 1, March 1995.

A data source **102** provides a serial bitstream **104** to the MCM transmitter. The incoming serial bitstream **104** is applied to a bit-carrier mapper **106** which produces a sequence of spectra **108** from the incoming serial bitstream **104**. An inverse fast Fourier transform (FFT) **110** is performed on the sequence of spectra **108** in order to produce a MCM time domain signal **112**. The MCM time domain signal forms the useful MCM symbol of the MCM time signal. To avoid intersymbol interference (ISI) caused by multipath distortion, a unit **114** is provided for inserting a guard interval of fixed length between adjacent MCM symbols in time. In accordance with a preferred embodiment of the present invention, the last part of the useful MCM symbol is used as the guard interval by placing same in front of the useful symbol. The resulting MCM symbol is shown at **115** in FIG. **7**.

A unit **116** for adding a reference symbol for each predetermined number of MCM symbols is provided in order to produce a MCM signal having a frame structure. Using this frame structure comprising useful symbols, guard intervals and reference symbols it is possible to recover the useful information from the MCM signal at the receiver side.

The resulting MCM signal having the structure shown at **118** in FIG. **7** is applied to the transmitter front end **120**. Roughly speaking, at the transmitter front end **120**, a digital/analog conversion and an up-converting of the MCM signal is performed. Thereafter, the MCM signal is transmitted through a channel **122**.

Following, the mode of operation of a MCM receiver **130** is shortly described referring to FIG. **7**. The MCM signal is received at the receiver front end **132**. In the receiver front end **132**, the MCM signal is down-converted and, furthermore, a digital/analog conversion of the down-converted signal is performed. The down-converted MCM signal is provided to a frame synchronization unit **134**. The frame synchronization unit **134** determines the location of the reference symbol in the MCM symbol. Based on the determination of the frame synchronization unit **134**, a reference symbol extracting unit **136** extracts the framing information, i.e. the reference symbol, from the MCM symbol coming from the receiver front end **132**. After the extraction of the reference symbol, the MCM signal is applied to a guard interval removal unit **138**.

The result of the signal processing performed so far in the MCM receiver are the useful MCM symbols. The useful MCM symbols output from the guard interval removal unit **138** are provided to a fast Fourier transform unit **140** in order to provide a sequence of spectra from the useful symbols. Thereafter, the sequence of spectra is provided to a carrier-bit mapper **142** in which the serial bitstream is recovered. This serial bitstream is provided to a data sink **144**.

As it is clear from FIG. **7**, every MCM transmitter **100** must contain a device which performs mapping of the transmitted bitstreams onto the amplitudes and/or phases of the sub-carriers. In addition, at the MCM receiver **130**, a device is needed for the inverse operation, i.e. retrieval of the transmitted bitstream from the amplitudes and/or phases of the sub-carriers.

For a better understanding of MCM mapping schemes, it is preferable to think of the mapping as being the assignment of one ore more bits to one or more sub-carrier symbols in the time-frequency plane. In the following, the term symbol or signal point is used for the complex number which represents the amplitude and/or phase modulation of a subcarrier in the equivalent baseband. Whenever all complex numbers representing all subcarrier symbols are designated, the term MCM symbol is used.

DESCRIPTION OF PRIOR ART

In principle, two methods for mapping the bitstream into the time-frequency plane are used in the prior art:

A first method is a differential mapping along the time axis. When using differential mapping along the time axis one or more bits are encoded into phase and/or amplitude shifts between two subcarriers of the same center frequency in adjacent MCM symbols. Such an encoding scheme is shown in FIG. **8**. The arrows depicted between the subcarrier symbols correspond to information encoded in amplitude and/or phase shifts between two subcarrier symbols.

A system applying such a mapping scheme is defined in the European Telecommunication standard ETS 300 401 (EU147-DAB). A system compliant to this standard uses Differential Quadrature Phase Shift Keying (DQPSK) to encode every two bits into a 0, 90, 180 or 270 degrees phase difference between two subcarriers of the same center frequency which are located in MCM symbols adjacent in time.

A second method for mapping the bitstream into the time-frequency plane is a non-differential mapping. When using non-differential mapping the information carried on a sub-carrier is independent of information transmitted on any other subcarrier, and the other subcarrier may differ either in frequency, i.e. the same MCM symbol, or in time, i.e. adjacent MCM symbols. A system applying such a mapping scheme is defined in the European Telecommunication standard ETS 300 744 (DVB-T). A system compliant to this standard uses 4, 16 or 64 Quadrature Amplitude Modulation (QAM) to assign bits to the amplitude and phase of a subcarrier.

The quality with which transmitted multi-carrier modulated signals can be recovered at the receiver depends on the properties of the channel. The most interesting property when transmitting MCM signals is the time interval at which a mobile channel changes its characteristics considerably. The channel coherence time T_c is normally used to determine the time interval at which a mobile channel changes its characteristics considerably. T_c depends on the maximum Doppler shift as follows:

$$f_{Doppler,max} = v \cdot f_{carrier}/c \qquad \text{(Eq.1)}$$

with

v: speed of the mobile receiver in [m/s]

fcarrier: carrier frequency of the RF signal [Hz]

c: speed of light ($3 \cdot 10^8$ m/s)

The channel coherence time T_c is often defined to be

$$T_c|_{50\%} = \frac{9}{16\pi f_{Doppler,max}} \text{ or } T_c|_{2nd\ Def.} = \frac{9}{16\pi f^2_{Doppler,max}} \qquad \text{(Eq. 2)}$$

It becomes clear from the existence of more than one definition, that the channel coherence time T_c is merely a rule-of-thumb value for the stationarity of the channel. As explained above, the prior art time-axis differential mapping requires that the mobile channel be quasi stationary during several MCM symbols periods, i.e. required channel coherence time Tc>>MCM symbol period. The prior art non-differential MCM mapping only requires that the mobile channel be quasi stationary during one symbol interval, i.e. required channel coherence time>MCM symbol period.

Thus, both prior art mapping schemes have specific disadvantages. For differential mapping into time axis direction the channel must be quasi stationary, i.e. the channel must not change during the transmission of two MCM symbols adjacent in time. If this requirement is not met, the channel induced phase and amplitude changes between MCM symbols will yield an increase in bit error rate.

With non-differential mapping exact knowledge of the phase of each subcarrier is needed (i.e. coherent reception). For multipath channels, coherent reception can only be obtained if the channel impulse response is known. Therefore, a channel estimation has to be part of the receiver algorithm. The channel estimation usually needs additional sequences in the transmitted waveform which do not carry information. In case of rapidly changing channels, which necessitate update of the channel estimation at short intervals, the additional overhead can quickly lead to insufficiency of non-differential mapping.

P. H. Moose: "Differentially Coded Multi-Frequency Modulation for Digital Communications", SIGNAL PROCESSING THEORIES AND APPLICATIONS, 18.–21. September 1990, pages 1807–1810, Amsterdam, NL, teaches a differentially coded multi-frequency modulation for digital communications. A multi-frequency differential modulation is described in which symbols are differentially encoded within each baud between adjacent tones. At the receiver, following a digital Fourier transform (DFT), the complex product between the DFT coefficient of digital frequency k and the complex conjugate of the DFT coefficient of digital frequency k−1 is formed. Thereafter, the result is multiplied by appropriate terms such that the differentially encoded phase bits are realigned to the original constellations. Thus, the constellation following the differential decoding must correspond to the original constellation.

SUMMARY OF THE INVENTION

It is an object of the present invention to provide methods and devices for performing an echo phase offset correction in a multi-carrier demodulation system.

In accordance with a first aspect, the present invention provides a method of performing an echo phase offset correction in a multi-carrier demodulation system, comprising the steps of:

differential phase decoding phase shifts based on a phase difference between simultaneous carriers having different frequencies;

determining an echo phase offset for each decoded phase shift by eliminating phase shift uncertainties related to the transmitted information from the decoded phase shift;

averaging the echo phase offsets in order to generate an averaged offset; and

correcting each decoded phase shift based on the averaged offset.

In accordance with a second aspect, the present invention provides a method of performing an echo phase offset correction in a multi-carrier demodulation system, comprising the steps of:

differential phase decoding phase shifts based on a phase difference between simultaneous carriers having different frequencies, the phase shifts defining signal points in a complex plane;

pre-rotating the signal points into the sector of the complex plane between −45° and +45°;

determining parameters of a straight line approximating the location of the pre-rotated signal points in the complex plane;

determining a phase offset based on the parameters; and

correcting each decoded phase shift based on the phase offset.

In accordance with a third aspect, the present invention provides an echo phase offset correction device for a multi-carrier demodulation system, comprising:

a differential phase decoder for decoding phase shifts based on a phase difference between simultaneous carriers having different frequencies;

means for determining an echo phase offset for each decoded phase shift by eliminating phase shift uncertainties related to the transmitted information from the decoded phase shift;

means for averaging the echo phase offsets in order to generate an averaged offset; and

means for correcting each decoded phase shift based on the averaged offset.

In accordance with a fourth aspect, the present invention provides an echo phase offset correction device for a multi-carrier demodulation system, comprising:

a differential phase decoder for decoding phase shifts based on a phase difference between simultaneous carriers having different frequencies, the phase shifts defining signal points in a complex plane;

means for pre-rotating the signal points into the sector of the complex plane between −45° and +45°;

means for determining parameters of a straight line approximating the location of the pre-rotated signal points in the complex plane;

means for determining a phase offset based on the parameters; and

means for correcting each decoded phase shift based on the phase offset.

The present invention provides methods and devices for performing an echo phase offset correction, suitable for multi-carrier (OFDM) digital broadcasting over rapidly changing multipath channels, comprising differential encoding of the data along the frequency axis such that there is no need for channel stationarity exceeding one multicarrier symbol.

When using the mapping process along the frequency axis it is preferred to make use of a receiver algorithm that will correct symbol phase offsets that can be caused by channel echoes.

The mapping scheme along the frequency axis for multi-carrier modulation renders the transmission to a certain extent independent of rapid changes in the multipath channel without introducing a large overhead to support channel estimation. Especially systems with high carrier frequencies and/or high speeds of the mobile carrying the receiving unit can benefit from such a mapping scheme.

Thus, the mapping scheme of a differential encoding along the frequency axis does not exhibit the two problems of the prior art systems described above. The mapping scheme is robust with regard to rapidly changing multipath channels which may occur at high frequencies and/or high speeds of mobile receivers.

The controlled respective parameters of the subcarriers are the phases thereof, such that the information is differentially phase encoded.

In accordance with the mapping described above, mapping is also differential, however, not into time axis direction but into frequency axis direction. Thus, the information is not contained in the phase shift between subcarriers adjacent in time but in the phase shift between subcarriers adjacent in frequency. Differential mapping along the frequency axis has two advantages when compared to prior art mapping schemes.

Because of differential mapping, no estimation of the absolute phase of the subcarriers is required. Therefore, channel estimation and the related overhead are not necessary. By choosing the frequency axis as direction for differentially encoding the information bitstream, the requirement that the channel must be stationary during several MCM symbols can be dropped. The channel only has to remain unchanged during the current MCM symbol period. Therefore, like for non-differential mapping it holds that

required channel coherence time≧MCM symbol period.

The present invention provides methods and apparatus for correction of phase distortions that can be caused by channel echoes. As described above, differential mapping into frequency axis direction solves problems related to the stationarity of the channel. However, differential mapping into frequency axis direction may create a new problem. In multipath environments, path echoes succeeding or preceding the main path can lead to systematic phase offsets between sub-carriers in the same MCM symbol. In this context, the main path is thought of being the path echo with the highest energy content. The main path echo will determine the position of the FFT window in the receiver of an MCM system.

According to the present invention, the information will be contained in a phase shift between adjacent subcarriers of the same MCM symbol. If not corrected for, the path echo induced phase offset between two subcarriers can lead to an increase in bit error rate. Therefore, application of the MCM mapping scheme presented in this invention will preferably be used in combination with a correction of the systematic subcarrier phase offsets in case of a multipath channel.

The introduced phase offset can be explained from the shifting property of the Discrete Fourier Transform (DFT):

$$x[((n-m))_N] \overset{DFT}{\longleftrightarrow} X[k]e^{-j\frac{2\pi}{N}km} \quad \text{(Eq. 3)}$$

with

$x[n]$: sampled time domain signal $(0 \le n \le N-1)$

$X[k]$: DFT transformed frequency domain signal $(0 \le k \le N-1)$

N: length of DFT

$(\ldots)_N$: cyclic shift of the DFT window in the time

m: length of DFT-Shift in the time domain

Equation 3 shows, that in a multipath channel, echoes following the main path will yield a subcarrier dependent phase offset. After differential demapping in the frequency axis direction at the receiver, a phase offset between two neighboring symbols remains. Because the channel induced phase offsets between differentially demodulated symbols are systematic errors, they can be corrected by an algorithm.

In the context of the following specification, algorithms which help correcting the phase shift are called Echo Phase offset correction (EPOC) algorithms. Two such algorithms are described as preferred embodiments for the correction of phase distortions that can be caused by channel echoes. These algorithms yield a sufficient detection security for MCM frequency axis mapping even in channels with echoes close to the limits of the guard interval.

In principle, an EPOC algorithm must calculate the echo induced phase offset from the signal space constellation following the differential demodulation and subsequently correct this phase offset.

BRIEF DESCRIPTION OF THE DRAWINGS

In the following, preferred embodiments of the present invention will be explained in detail on the basis of the drawings enclosed, in which:

FIG. **1** shows a schematic view representing a mapping scheme used according to the invention;

FIG. **2** shows a functional block diagram of an embodiment of a mapping device;

FIGS. **3**A and **3**B show scatter diagrams of the output of an differential de-mapper of a MCM receiver for illustrating the effect of an echo phase offset correction;

FIG. **4** shows a schematic block diagram for illustrating the position and the functionality of an echo phase offset correction unit;

FIG. **5** shows a schematic block diagram of an embodiment of an echo phase offset correction device according to the present invention;

FIG. **6** shows schematic views for illustrating a projection performed by another embodiment of an echo phase offset correction device according to the present invention;

FIG. **7** shows a schematic block diagram of a generic multi-carrier modulation system; and

FIG. **8** shows a schematic view representing a prior art differential mapping scheme.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

In a preferred embodiment thereof, the present invention is applied to a MCM system as shown in FIG. **7**. With respect to this MCM system, the present invention relates to the bit-carrier mapper **106** of the MCM transmitter **100** and the carrier-bit mapper **142** of the MCM receiver **130**, which are depicted with a shaded background in FIG. **7**.

An preferred embodiment of an inventive mapping scheme used by the bit-carrier mapper **106** is depicted in FIG. **1**. A number of MCM symbols **200** is shown in FIG. **1**. Each MCM symbol **200** comprises a number of sub-carrier symbols **202**. The arrows **204** in FIG. **1** illustrate information encoded between two sub-carrier symbols **202**. As can be seen from the arrows **204**, the bit-carrier mapper **106** uses a differential mapping within one MCM symbol along the frequency axis direction.

In the embodiment shown in FIG. **1**, the first sub-carrier (k=0) in an MCM symbol **200** is used as a reference sub-carrier **206** (shaded) such that information is encoded between the reference sub-carrier and the first active carrier **208**. The other information of a MCM symbol **200** is encoded between active carriers, respectively.

Thus, for every MCM symbol an absolute phase reference exists. In accordance with FIG. **1**, this absolute phase reference is supplied by a reference symbol inserted into every MCM symbol (k=0). The reference symbol can either have a constant phase for all MCM symbols or a phase that varies from MCM symbol to MCM symbol. A varying phase can be obtained by replicating the phase from the last subcarrier of the MCM symbol preceding in time.

In FIG. **2** a preferred embodiment of a device for performing a differential mapping along the frequency axis is shown. Referring to FIG. **2**, assembly of MCM symbols in the frequency domain using differential mapping along the frequency axis according to the present invention is described.

FIG. **2** shows the assembly of one MCM symbol with the following parameters:

NFFT designates the number of complex coefficients of the discrete Fourier transform, number of subcarriers respectively.

K designates the number of active carriers. The reference carrier is not included in the count for K.

According to FIG. **2**, a quadrature phase shift keying (QPSK) is used for mapping the bitstream onto the complex symbols. However, other M-ary mapping schemes (MPSK) like 2-PSK, 8-PSK, 16-QAM, 16-APSK, 64-APSK etc. are possible.

Furthermore, for ease of filtering and minimization of aliasing effects some subcarriers are not used for encoding information in the device shown in FIG. **2**. These subcarriers, which are set to zero, constitute the so-called guard bands on the upper and lower edges of the MCM signal spectrum.

At the input of the mapping device shown in FIG. **2**, complex signal pairs b**0**[k], b**1**[k] of an input bitstream are received. K complex signal pairs are assembled in order to form one MCM symbol. The signal pairs are encoded into the K differential phase shifts phi[k] needed for assembly of one MCM symbol. In this embodiment, mapping from Bits to the 0, 90, 180 and 270 degrees phase shifts is performed using Gray Mapping in a quadrature phase shift keying device **220**.

Gray mapping is used to prevent that differential detection phase errors smaller than 135 degrees cause double bit errors at the receiver.

Differential phase encoding of the K phases is performed in a differential phase encoder **222**. At this stage of processing, the K phases phi[k] generated by the QPSK Gray mapper are differentially encoded. In principal, a feedback loop **224** calculates a cumulative sum over all K phases. As starting point for the first computation (k=0) the phase of the reference carrier **226** is used. A switch **228** is provided in order to provide either the absolute phase of the reference subcarrier **226** or the phase information encoded onto the preceding (i.e. z^{-1}, where z^{-1} denotes the unit delay operator) subcarrier to a summing point **230**. At the output of the differential phase encoder **222**, the phase information theta [k] with which the respective subcarriers are to be encoded is provided. In preferred embodiments of the present invention, the subcarriers of a MCM symbol are equally spaced in the frequency axis direction.

The output of the differential phase encoder **222** is connected to a unit **232** for generating complex subcarrier symbols using the phase information theta[k]. To this end, the K differentially encoded phases are converted to complex symbols by multiplication with

$$\text{factor}*e^{j*[2*pi*(theta[k]+PHI)]} \qquad \text{(Eq.4)}$$

wherein factor designates a scale factor and PHI designates an additional angle. The scale factor and the additional angle PHI are optional. By choosing PHI=45° a rotated DQPSK signal constellation can be obtained.

Finally, assembly of a MCM symbol is effected in an assembling unit **234**. One MCM symbol comprising N_{FFT} subcarriers is assembled from N_{FFT}-K−1 guard band symbols which are "zero", one reference subcarrier symbol and K DQPSK subcarrier symbols. Thus, the assembled MCM symbol **200** is composed of K complex values containing the encoded information, two guard bands at both sides of the NFFT complex values and a reference subcarrier symbol.

The MCM symbol has been assembled in the frequency domain. For transformation into the time domain an inverse discrete Fourier transform (IDFT) of the output of the assembling unit **234** is performed by a transformator **236**. In preferred embodiments of the present invention, the transformator **236** is adapted to perform a fast Fourier transform (FFT).

Further processing of the MCM signal in the transmitter as well as in the receiver is as described above referring to FIG. **7**.

At the receiver a de-mapping device **142** (FIG. **7**) is needed to reverse the operations of the mapping device described above referring to FIG. **2**. The implementation of the de-mapping device is straightforward and, therefore, need not be described herein in detail.

However, systematic phase shifts stemming from echoes in multipath environments may occur between subcarriers in the same MCM symbol. This phase offsets can cause bit errors when demodulating the MCM symbol at the receiver.

Thus, it is preferred to make use of an algorithm to correct the systematic phase shifts stemming from echoes in multipath environments. Preferred embodiments of echo phase offset correction algorithms are explained hereinafter referring to FIGS. **3** to **6**.

In FIGS. **3**A and **3**B, scatter diagrams at the output of a differential demapper of a MCM receiver are shown. As can be seen from FIG. **3**A, systematic phase shifts between subcarriers in the same MCM symbol cause a rotation of the demodulated phase shifts with respect to the axis of the complex coordinate system. In FIG. **3**B, the demodulated phase shifts after having performed an echo phase offset correction are depicted. Now, the positions of the signal points are substantially on the axis of the complex coordinate system. These positions correspond to the modulated phase shifts of 0°, 90°, 180° and 270°, respectively.

An echo phase offset correction algorithm (EPOC algorithm) must calculate the echo induced phase offset from the signal space constellation following the differential demodulation and subsequently-correct this phase offset.

For illustration purposes, one may think of the simplest algorithm possible which eliminates the symbol phase before computing the mean of all phases of the subcarriers. To illustrate the effect of such an EPOC algorithm, reference is made to the two scatter diagrams of subcarriers symbols contained in one MCM symbol in FIGS. **3**A and **3**B. This scatter diagrams have been obtained as result of an MCM simulation. For the simulation a channel has been used which might typically show up in single frequency networks. The echoes of this channel stretched to the limits of the MCM guard interval. The guard interval was chosen to be 25% of the MCM symbol duration in this case.

FIG. **4** represents a block diagram for illustrating the position and the functionality of an echo phase offset correction device in a MCM receiver. The signal of a MCM transmitter is transmitted through the channel **122** (FIGS. **4** and **7**) and received at the receiver frontend **132** of the MCM receiver. The signal processing between the receiver frontend and the fast Fourier transformator **140** has been omitted in FIG. **4**. The output of the fast Fourier transformator is applied to the de-mapper, which performs a differential de-mapping along the frequency axis. The output of the de-mapper are the respective phase shifts for the subcarriers. The phase offsets of this phase shifts which are caused by echoes in multipath environments are visualized by a block **400** in FIG. **4** which shows an example of a scatter diagram of the subcarrier symbols without an echo phase offset correction.

The output of the de-mapper **142** is applied to the input of an echo phase offset correction device **402**. The echo phase offset correction device **402** uses an EPOC algorithm in order to eliminate echo phase offsets in the output of the de-mapper **142**. The result is shown in block **404** of FIG. **4**, i.e. only the encoded phase shifts, 0°, 90°, 180° or 270° are present at the output of the correction device **402**. The output of the correction device **402** forms the signal for the metric calculation which is performed in order to recover the bitstream representing the transmitted information.

A first embodiment of an EPOC algorithm and a device for performing same is now described referring to FIG. **5**.

The first embodiment of an EPOC algorithm starts from the assumption that every received differentially decoded complex symbol is rotated by an angle due to echoes in the multipath channel. For the subcarriers equal spacing in frequency is assumed since this represents a preferred embodiment of the present invention. If the subcarriers were not equally spaced in frequency, a correction factor would have to be introduced into the EPOC algorithm.

FIG. **5** shows the correction device **402** (FIG. **4**) for performing the first embodiment of an EPOC algorithm.

From the output of the de-mapper **142** which contains an echo phase offset as shown for example in FIG. **3**A, the phase shifts related to transmitted information must first be discarded. To this end, the output of the de-mapper **142** is applied to a discarding unit **500**. In case of a DQPSK mapping, the discarding unit can perform a "$(.)^4$" operation. The unit **500** projects all received symbols into the first quadrant. Therefore, the phase shifts related to transmitted information is eliminated from the phase shifts representing the subcarrier symbols. The same effect could be reached with a modulo-4 operation.

Having eliminated the information related symbol phases in unit **500**, the first approach to obtain an estimation would be to simply compute the mean value over all symbol phases of one MCM symbol. However, it is preferred to perform a threshold decision before determining the mean value over all symbol phases of one MCM symbol. Due to Rayleigh fading some of the received symbols may contribute unreliable information to the determination of the echo phase offset. Therefore, depending on the absolute value of a symbol, a threshold decision is performed in order to determine whether the symbol should contribute to the estimate of the phase offset or not.

Thus, in the embodiment shown in FIG. **5**, a threshold decision unit **510** is included. Following the unit **500** the absolute value and the argument of a differentially decoded symbol is computed in respective computing units **512** and **514**. Depending on the absolute value of a respective symbol, a control signal is derived. This control signal is compared with a threshold value in a decision circuit **516**. If the absolute value, i.e. the control signal thereof, is smaller than a certain threshold, the decision circuit **516** replaces the angle value going into the averaging operation by a value equal to zero. To this end, a switch is provided in order to disconnect the output of the argument computing unit **514** from the input of the further processing stage and connects the input of the further processing stage with a unit **518** providing a constant output of "zero".

An averaging unit **520** is provided in order to calculate a mean value based on the phase offsets ϕ_i determined for the individual subcarrier symbols of a MCM symbol as follows:

$$\bar{\varphi} = \frac{1}{K}\sum_{i=1}^{K} \varphi_i \qquad \text{(Eq. 5)}$$

In the averaging unit **520**, summation over K summands which have not been set to zero in the unit **516** is performed. The output of the averaging unit **520** is provided to a hold unit **522** which holds the output of the averaging unit **520** K times. The output of the hold unit **522** is connected with a phase rotation unit **524** which performs the correction of the phase offsets of the K complex signal points on the basis of the mean value $\bar{\phi}$.

The phase rotation unit **524** performs the correction of the phase offsets by making use of the following equation:

$$v'_k = v_k \cdot e^{-j\bar{\phi}} \qquad \text{(Eq.6)}$$

In this equation, v'_k designates the K phase corrected differentially decoded symbols for input into the soft-metric calculation, whereas v_k designates the input symbols. As long as a channel which is quasi stationary during the duration of one MCM symbols can be assumed, using the mean value over all subcarriers of one MCM symbol will provide correct results.

A buffer unit **527** may be provided in order to buffer the complex signal points until the mean value of the phase offsets for one MCM symbol is determined. The output of the phase rotation unit **524** is applied to the further processing stage **526** for performing the soft-metric calculation.

With respect to the results of the above echo phase offset correction, reference is made again to FIGS. **3**A and **3**B. The two plots stem from a simulation which included the first embodiment of an echo phase offset correction algorithm described above. At the instant of the scatter diagram snapshot shown in FIG. **3**A, the channel obviously distorted the constellation in a way, that a simple angle rotation is a valid assumption. As shown in FIG. **3**B, the signal constellation can be rotated back to the axis by applying the determined mean value for the rotation of the differentially detected symbols.

A second embodiment of an echo phase offset correction algorithm is described hereinafter. This second embodiment can be preferably used in connection with multipath channels that have up to two strong path echoes. The algorithm of the second embodiment is more complex than the algorithm of the first embodiment.

What follows is a mathematical derivation of the second embodiment of a method for echo phase offset correction. The following assumptions can be made in order to ease the explanation of the second embodiment of an EPOC algorithm.

In this embodiment, the guard interval of the MCM signal is assumed to be at least as long as the impulse response h[q], q=0, 1, . . . , Qh−1 of the multipath channel.

At the transmitter every MCM symbol is assembled using frequency axis mapping explained above. The symbol of the reference subcarrier equals 1, i.e. 0 degree phase shift. The optional phase shift PHI equals zero, i.e. the DQPSK signal constellation is not rotated.

Using an equation this can be expressed as

$$a_k = a_{k-1} a_k^{inc} \qquad \text{(Eq. 7)}$$

with

k: index $k = 1, 2, \ldots, K$ of the active subcarrier;

$a_k^{inc} = e^{j\frac{\pi}{2}m}$: complex phase increment symbol; $m = 0, 1, 2, 3$ is the $QPSK$ symbol number which is derived from Gray encoding pairs of 2 Bits;

$a_0 = 1$: symbol of the reference subcarrier.

At the DFT output of the receiver the decision variables

$$e_k = a_k H_k \qquad \text{(Eq.8)}$$

are obtained with

$$H_k = \sum_{i=0}^{Q_h-1} h[i] \cdot e^{-j\frac{2\pi}{K}ki} \qquad \text{(Eq. 9)}$$

being the DFT of the channel impulse response h[q] at position k.

With $|a_k|^2=1$ the differential demodulation yields

$$v_k = e_k \cdot e_{k-1}^* = a_k^{inc} H_k H_{k-1}^* \qquad \text{(Eq.10)}$$

For the receiver an additional phase term ϕ_k is introduced, which shall be used to correct the systematic phase offset caused by the channel. Therefore, the final decision variable at the receiver is

$$v'_k = v_k \cdot e^{j\phi_k} = a_k^{inc} \cdot e^{j\phi_k} \cdot H_k \cdot H_{k-1}^* \qquad \text{(Eq.11)}$$

As can be seen from the Equation 11, the useful information a_k^{inc} is weighted with the product $e^{j\phi_k} \cdot H_k \cdot H_{k-1}^*$ (rotation and effective transfer function of the channel). This product must be real-valued for an error free detection. Considering this, it is best to choose the rotation angle to equal the negative argument of $H_k \cdot H_{k-1}^*$. To derive the desired algorithm for 2-path channels, the nature of $H_k \cdot H_{k-1}^*$ is investigated in the next section.

It is assumed that the 2-path channel exhibits two echoes with energy content unequal zero, i.e. at least two dominant echoes. This assumption yields the impulse response

$$h[q] = c_1 \delta_0[q] + c_2 \delta_0[q-q_0] \qquad \text{(Eq.12)}$$

with

c_1, c_2: complex coefficients representing the path echoes;

q_0: delay of the second path echo with respect to the first path echo;

δ_0: Dirac pulse; $\delta_0[k]=1$ for k=0
$\delta_0[k]=0$ else

The channel transfer function is obtained by applying a DFT (Eq.9) to Equation 12:

$$H_k = H\left(e^{j\frac{2\pi}{K}k}\right) = c_1 + c_2 \cdot e^{-j\frac{2\pi}{K}kq_0} \qquad \text{(Eq. 13)}$$

With Equation 13 the effective transfer function for differential demodulation along the frequency axis is:

$$\begin{aligned} H_k \cdot H_{k-1}^* &= \left(c_1 + c_2 e^{-j\frac{2\pi}{K}kq_0}\right) \cdot \left(c_1^* + c_2^* e^{+j\frac{2\pi}{K}(k-1)q_0}\right) \\ &= c_a + c_b \cos\left(\frac{\pi}{K} q_0 (2k-1)\right) \end{aligned} \qquad \text{(Eq. 14)}$$

Assuming a noise free 2-path channel, it can be observed from Equation 14 that the symbols on the receiver side are located on a straight line in case the symbol 1+j0 has been send (see above assumption). This straight line can be characterized by a point

$$c_a = |c_1|^2 + |c_2|^2 \cdot e^{-j\frac{2\pi}{K}q_0} \qquad \text{(Eq. 15)}$$

and the vector

$$c_b = 2c_1 c_2^* \cdot e^{-j\frac{\pi}{K}q_0} \qquad \text{(Eq. 16)}$$

which determines its direction.

With the above assumptions, the following geometric derivation can be performed. A more suitable notation for the geometric derivation of the second embodiment of an EPOC algorithm is obtained if the real part of the complex plane is designated as x=Re{z}, the imaginary part as y=Im{z}, respectively, i.e. z=x+jy. With this new notation, the straight line, on which the received symbols will lie in case of a noise-free two-path channel, is

$$f(x) = a + b \cdot x \quad \text{(Eq. 17)}$$

with

$$a = \mathrm{Im}\{c_a\} - \frac{\mathrm{Re}\{c_a\}}{\mathrm{Re}\{c_b\}} \cdot \mathrm{Im}\{c_b\} \quad \text{(Eq. 18)}$$

and

$$b = -\frac{\mathrm{Im}\{c_a\} - \frac{\mathrm{Re}\{c_a\}}{\mathrm{Re}\{c_b\}} \cdot \mathrm{Im}\{c_b\}}{\mathrm{Re}\{c_a\} - \frac{\mathrm{Im}\{c_a\}}{\mathrm{Im}\{c_b\}} \cdot \mathrm{Re}\{c_b\}} \quad \text{(Eq. 19)}$$

Additional noise will spread the symbols around the straight line given by Equations 17 to 19. In this case Equation 19 is the regression curve for the cluster of symbols.

For the geometric derivation of the second embodiment of an EPOC algorithm, the angle ϕ_k from Equation 11 is chosen to be a function of the square distance of the considered symbol from the origin:

$$\phi_k = f_K(|z|^2) \quad \text{(Eq.20)}$$

Equation 20 shows that the complete signal space is distorted (torsion), however, with the distances from the origin being preserved.

For the derivation of the algorithm of the second embodiment, $f_K(\cdot)$ has to be determined such that all decision variables v'_k (assuming no noise) will come to lie on the real axis:

$$\mathrm{Im}\left\{(x + jf(x)) \cdot e^{jf_K(|z|^2)}\right\} = 0 \quad \text{(Eq. 21)}$$

Further transformations of Equation 21 lead to a quadratic equation which has to be solved to obtain the solution for ϕ_k.

In case of a two-path channel, the echo phase offset correction for a given decision variable v_k is

$$v'_k = v_k \cdot e^{j\varphi_k} \quad \text{(Eq. 22)}$$

with

$$\varphi_k = \begin{cases} -a\tan\left(\dfrac{a + b\sqrt{|v_k|^2(1+b^2) - a^2}}{-ab + \sqrt{|v_k|^2(1+b^2) - a^2}}\right) & \text{for } |v_k|^2 \geq \dfrac{a^2}{1+b^2} \\ a\tan\left(\dfrac{1}{b}\right) & \text{for } |v_k|^2 < \dfrac{a^2}{1+b^2} \end{cases} \quad \text{(Eq. 23)}$$

From the two possible solutions of the quadratic equation mentioned above, Equation 23 is the one solution that cannot cause an additional phase shift of 180 degrees.

The two plots in FIG. **15** show the projection of the EPOC algorithm of the second embodiment for one quadrant of the complex plane. Depicted here is the quadratic grid in the sector $|\arg(z)| \leqq \pi/4$ and the straight line y=f(x)=a+b·x with a =−1.0 and b=0.5 (dotted line). In case of a noise-free channel, all received symbols will lie on this straight line if 1+j0 was send. The circle shown in the plots determines the

boarder line for the two cases of Equation 23. In the left part, FIG. **15** shows the situation before the projection, in the right part, FIG. **15** shows the situation after applying the projection algorithm. By looking on the left part, one can see, that the straight line now lies on the real axis with 2+j0 being the fix point of the projection. Therefore, it can be concluded that the echo phase offset correction algorithm according to the second embodiment fulfills the design goal.

Before the second embodiment of an EPOC algorithm can be applied, the approximation line through the received symbols has to be determined, i.e. the parameters a and b must be estimated. For this purpose, it is assumed that the received symbols lie in sector $|\arg(z)| \leqq \pi/4$, if 1+j0 was sent. If symbols other than 1+j0 have been sent, a modulo operation can be applied to project all symbols into the desired sector. Proceeding like this prevents the necessity of deciding on the symbols in an early stage and enables averaging over all signal points of one MCM symbol (instead of averaging over only ¼ of all signal points).

For the following computation rule for the EPOC algorithm of the second embodiment, x_i is used to denote the real part of the i-th signal point and y_i for its imaginary part, respectively (i=1, 2, . . . , K). Altogether, K values are available for the determination. By choosing the method of least squares, the straight line which has to be determined can be obtained by minimizing

$$(a, b) = \underset{(\tilde{a}, \tilde{b})}{\arg\min} \sum_{i=1}^{K} \left(y_i - (\tilde{a} + \tilde{b} \cdot x_i)\right)^2 \quad \text{(Eq. 24)}$$

The solution for Equation 24 can be found in the laid open literature. It is

$$b = \frac{\sum_{i=1}^{K} (x_i - \overline{x}) \cdot y_i}{\sum_{i=1}^{K} (x_i - \overline{x})^2}, \; a = \overline{y} - \overline{x} \cdot b \quad \text{(Eq. 25)}$$

with mean values

$$\overline{x} = \frac{1}{N} \sum_{i=1}^{K} x_i, \; \overline{y} = \frac{1}{N} \sum_{i=1}^{K} y_i \quad \text{(Eq. 26)}$$

If necessary, an estimation method with higher robustness can be applied. However, the trade-off will be a much higher computational complexity.

To avoid problems with the range in which the projection is applicable, the determination of the straight line should be separated into two parts. First, the cluster's centers of gravity are moved onto the axes, following, the signal space is distorted. Assuming that a and b are the original parameters of the straight line and α is the rotation angle, $f_K(\cdot)$ has to be applied with the transformed parameters

$$b' = \frac{b \cdot \cos(\alpha) - \sin(\alpha)}{\cos(\alpha) + b \cdot \sin(\alpha)}, \quad a' = a \cdot (\cos(\alpha) - b' \cdot \sin(\alpha)) \quad \text{(Eq.27)}$$

Besides the two EPOC algorithms explained above section, different algorithms can be designed that will, however, most likely exhibit a higher degree of computational complexity.

The new mapping method for Multicarrier Modulation schemes presented herein consists in principal of two important aspects. Differential mapping within one MCM symbol along the frequency axis direction and correction of the channel echo related phase offset on the subcarriers at the receiver side. The advantage of this new mapping scheme is its robustness with regard to rapidly changing multipath channels which may occur at high frequencies and/or high speeds of mobile receivers.

What is claimed is:

1. A method of performing an echo phase offset correction in a multi-carrier demodulation system, comprising the steps of:

differential phase decoding phase shifts based on a phase difference between simultaneous carriers having different frequencies;

determining an echo phase offset for each decoded phase shift by eliminating phase shift uncertainties related to the transmitted information from said decoded phase shift;

averaging said echo phase offsets in order to generate an averaged offset;

correcting each decoded phase shift based on said averaged offset; and

further comprising a step of comparing an absolute value of a symbol associated with a respective decoded phase shift with a threshold, wherein only phase shifts having associated therewith symbols having an absolute value exceeding said threshold are used in said step of averaging said echo phase offsets.

2. The method according to claim 1, wherein said step of differential phase decoding comprises the step of differential phase decoding phase shifts based on a phase difference between simultaneous carriers which are adjacent in the frequency axis direction.

3. The method according to claim 1, wherein said step of differential phase decoding comprises the step of differential phase decoding phase shifts based on phase differences between at least three simultaneous carriers which are equally spaced in the frequency axis direction.

4. A method of performing an echo phase offset correction in a multi-carrier demodulation system, comprising the steps of:

differential phase decoding phase shifts based on a phase difference between simultaneous carriers having different frequencies, said phase shifts defining signal points in a complex plane;

pre-rotating said signal points into the sector of said complex plane between −45° and +45°;

determining parameters a, b of a straight line approximating the location of said pre-rotated signal points in said complex plane;

determining a phase offset based on said parameters a, b; and

correcting each decoded phase shift based on said phase offset.

5. The method according to claim 4, wherein said simultaneous carriers are equally spaced in the frequency axis direction.

6. The method according to claim 4, wherein said step of determining said parameters a, b comprises a least squares method for selecting those parameters which minimize the deviations of said pre-rotated signal points from said straight line.

7. The method according to claim 6, wherein said parameters a, b are determined as follows:

$$b = \frac{\sum_{i=1}^{K}(x_i - \overline{x})\cdot y_i}{\sum_{i=1}^{K}(x_i - \overline{x})^2}, \quad a = \overline{y} - \overline{x}\cdot b$$

$$\overline{x} = \frac{1}{N}\sum_{i=1}^{K} x_i, \quad \overline{y} = \frac{1}{N}\sum_{i=1}^{K} y_i$$

wherein

x and y designate the coordinates of the signal points in the complex plane,

i is an index from 1 to N, and

K is the numer of signal points.

8. The method according to claim 7, wherein said phase offset ϕ_k is determined as follows:

$$\varphi_k = \begin{cases} -a\tan\left(\dfrac{a + b\sqrt{|v_k|^2(1+b^2) - a^2}}{-ab + \sqrt{|v_k|^2(1+b^2) - a^2}}\right) & \text{for } |v_k|^2 \geq \dfrac{a^2}{1+b^2} \\ a\tan\left(\dfrac{1}{b}\right) & \text{for } |v_k|^2 < \dfrac{a^2}{1+b^2} \end{cases}$$

wherein v_k is a given decision variable.

9. An echo phase offset correction device for a multicattier demodulation system, comprising:

a differential phase decoder for decoding phase shifts based on a phase difference between simultaneous carriers having different frequencies;

means for determining an echo phase offset for each decoded phase shift comprising means for eliminating phase shift uncertainties related to the transmitted information from said decoded phase shift;

means for averaging said echo phase offsets in order to generate an averaged offset;

means for correcting each decoded phase shift based on said averaged offset; and

means for comparing an absolute value of a symbol associated with a respective decoded phase shift with a threshold, wherein said means for averaging said phase offsets only uses phase shifts having associated therewith symbols having an absolute value exceeding said threshold.

10. The device according to claim 9, wherein said differential phase decoder is adapted for decoding said phase shifts based on a phase difference between simultaneous carriers which are adjacent in the frequency axis direction.

11. The device according to claim 9, wherein said differential phase decoder is adapted for decoding said phase shifts based on phase differences between at least three simultaneous carriers which are equally spaced in the frequency axis direction.

12. An echo phase offset correction device for a multicarrier demodulation system, comprising:

a differential phase decoder for decoding phase shifts based on a phase difference between simultaneous carriers having different frequencies, said phase shifts defining signal points in a complex plane;

means for pre-rotating said signal points into the sector of said complex plane between −45° and +45°;

means for determining parameters a, b of a straight line approximating the location of said pre-rotated signal points in said complex plane;

means for determining a phase offset based on said parameters a, b; and

means for correcting each decoded phase shift based on said phase offset.

13. The device according to claim **12**, wherein said differential phase decoder comprises means for decoding phase shifts of at least three simultaneous carriers which are equally spaced in the frequency axis direction.

14. The device according to claim **12**, wherein said means for determining said parameters a, b comprises means for performing a least squares method for selecting those parameters which minimize the deviations of said pre-rotated signal points from said straight line.

15. The device according to claim **14**, wherein said means for determining said parameters a, b calculates said parameters a, b as follows:

$$b = \frac{\sum_{i=1}^{K}(x_i - \overline{x})\cdot y_i}{\sum_{i=1}^{K}(x_i - \overline{x})^2}, \quad a = \overline{y} - \overline{x}\cdot b$$

-continued

$$\overline{x} = \frac{1}{N}\sum_{i=1}^{K} x_i, \quad \overline{y} = \frac{1}{N}\sum_{i=1}^{K} y_i$$

wherein

x and y designate the coordinates of the signal points in the complex plane,

i is an index from 1 to N, and

K is the number of signal points.

16. The device according to claim **15**, wherein said means for determining said phase offset ϕ_k calculates said phase offset ϕ_k as follows:

$$\varphi_k = \begin{cases} -a\tan\left(\dfrac{a + b\sqrt{|v_k|^2(1+b^2) - a^2}}{-ab + \sqrt{|v_k|^2(1+b^2) - a^2}}\right) & \text{for } |v_k|^2 \geq \dfrac{a^2}{1+b^2} \\ a\tan\left(\dfrac{1}{b}\right) & \text{for } |v_k|^2 < \dfrac{a^2}{1+b^2} \end{cases}$$

wherein v_k is a given decision variable.

* * * * *

Exhibit C

US006993084B1

(12) **United States Patent**
Eberlein et al.

(10) **Patent No.: US 6,993,084 B1**
(45) **Date of Patent: Jan. 31, 2006**

(54) **COARSE FREQUENCY SYNCHRONISATION IN MULTICARRIER SYSTEMS**

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(*) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 0 days.

(21) Appl. No.: **09/673,271**

(22) PCT Filed: **Apr. 14, 1998**

(86) PCT No.: **PCT/EP98/02169**

§ 371 (c)(1), (2), (4) Date: **Nov. 28, 2000**

(87) PCT Pub. No.: **WO99/53665**

PCT Pub. Date: **Oct. 21, 1999**

(51) **Int. Cl.**
H04L 27/28 (2006.01)

(52) **U.S. Cl.** **375/260**

(58) **Field of Classification Search** 375/132, 375/130, 135, 261, 326, 379, 295, 376, 324, 375/266, 268, 140, 260, 262, 271; 370/206, 370/517, 203, 204, 207, 208, 216, 343, 480, 370/210, 347, 510, 509, 470, 279, 209; 35/138
See application file for complete search history.

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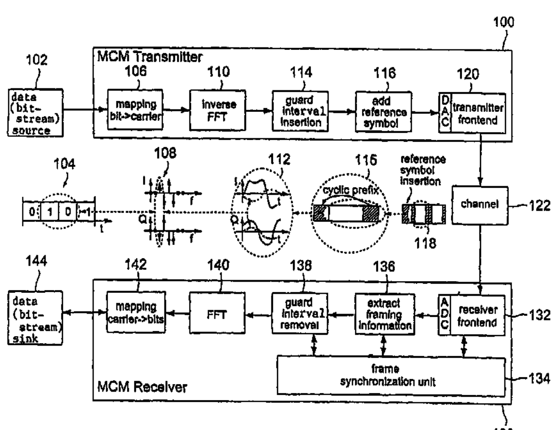
Primary Examiner—Emmanuel Bayard
(74) *Attorney, Agent, or Firm*—Roylance, Abrams, Berdo & Goodman, L.L.P.

(57) **ABSTRACT**

A method for generating a signal having a frame structure, each frame of the frame structure comprising at least one useful symbol, a guard interval associated to the at least one useful symbol and a reference symbol, comprises the steps of performing an amplitude modulation of a bit sequence such that the envelope of the amplitude modulated bit sequence defines a reference pattern of the reference symbol and inserting the amplitude modulated bit sequence into said signal as said reference symbol. A method for frame synchronization of a signal having such a frame structure comprises the steps of receiving the signal, down-converting the received signal, performing an amplitude-demodulation of the down-converted signal in order to generate an envelope, correlating the envelope with a predetermined reference pattern in order to detect a signal reference pattern of the reference symbol in the signal, and performing the frame synchronization based on the detection of the signal reference pattern.

46 Claims, 3 Drawing Sheets



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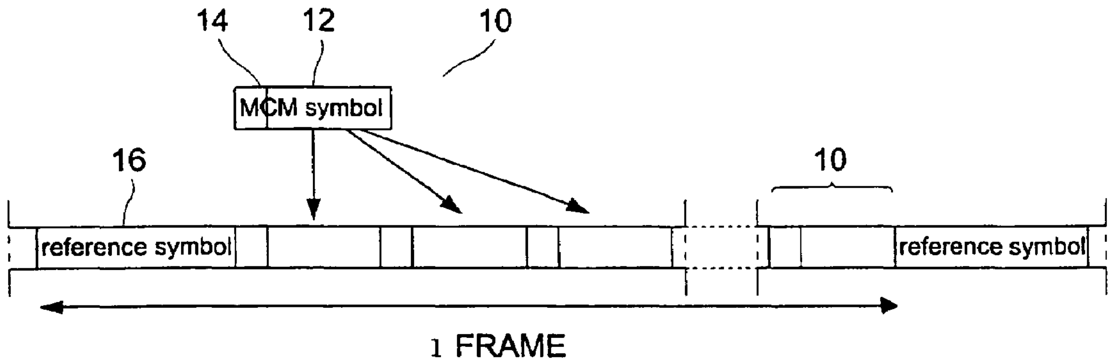
14
12
10
MCM symbol
16
10
reference symbol
reference symbol
1 FRAME

FIG.1

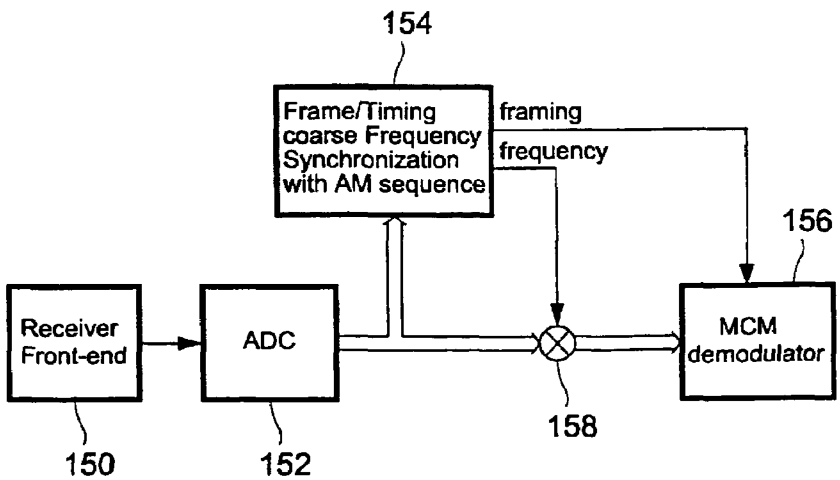
154
Frame/Timing coarse Frequency Synchronization with AM sequence
framing
frequency
156
Receiver Front-end
ADC
MCM demodulator
158
150
152

FIG.3

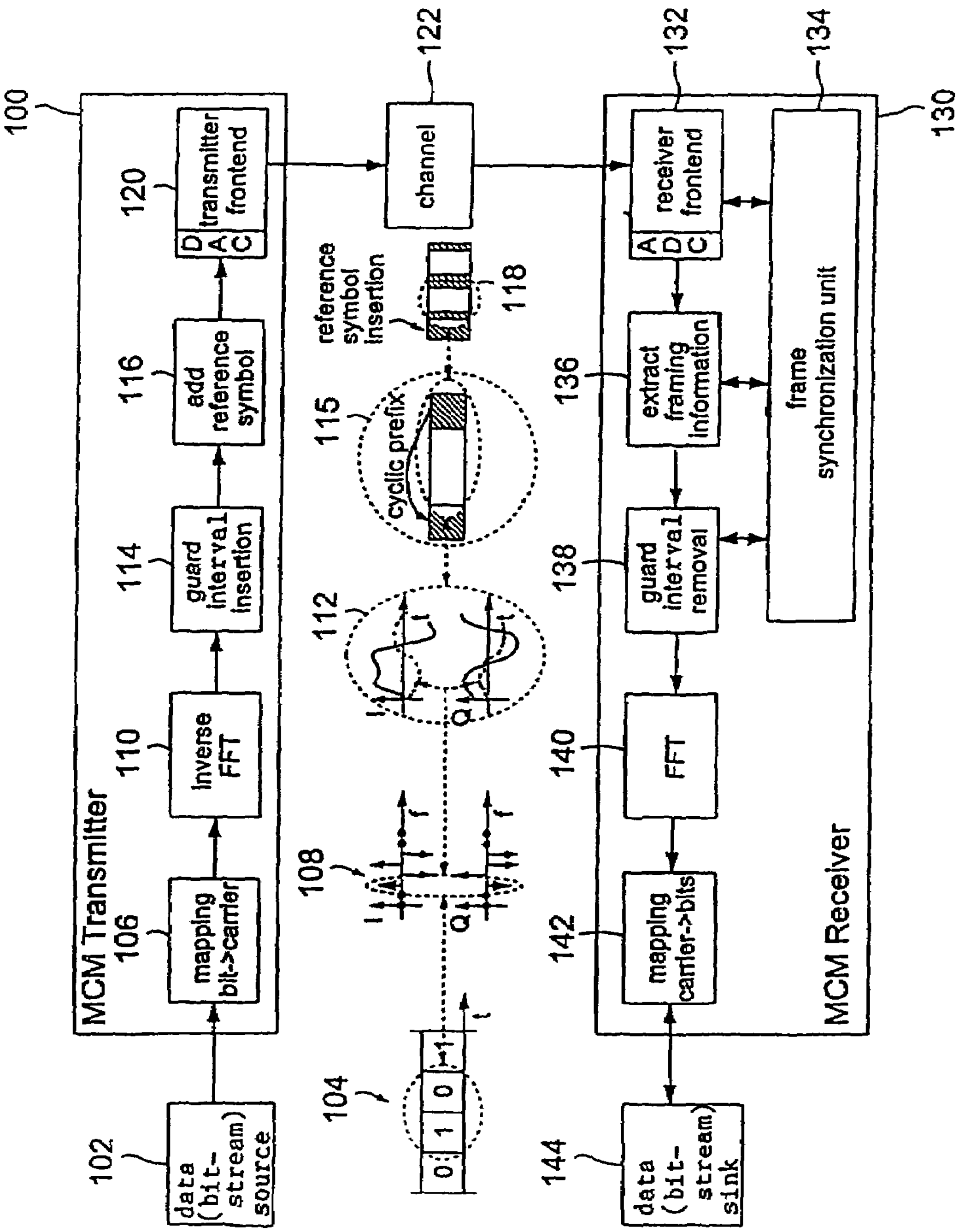
100
MCM Transmitter
102
data (bit-stream) source
106
mapping bit->carrier
110
inverse FFT
114
guard interval insertion
116
add reference symbol
120
DAC
transmitter frontend
122
channel
104
108
112
115
cyclic prefix
118
reference symbol insertion
130
MCM Receiver
132
ADC
receiver frontend
134
frame synchronization unit
136
extract framing information
138
guard interval removal
140
FFT
142
mapping carrier->bits
144
data (bit-stream) sink

FIG.2

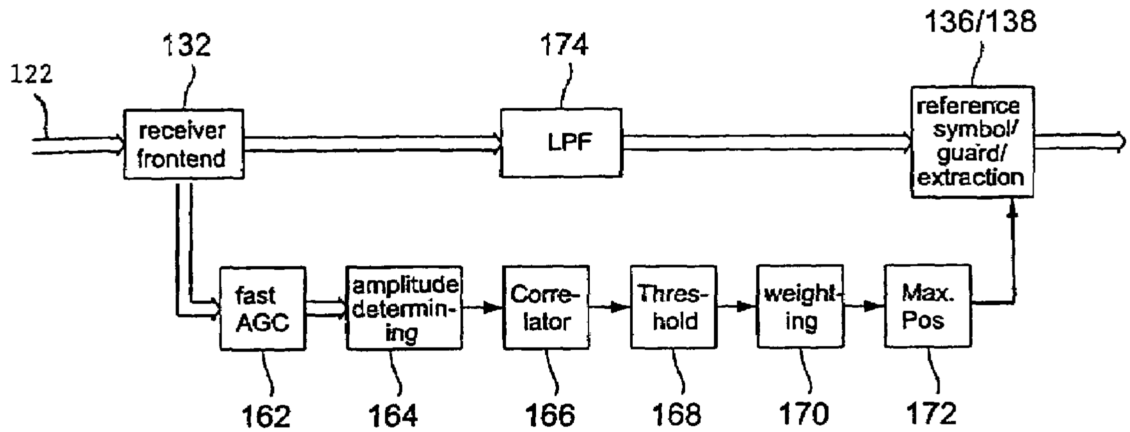
122
132
174
136/138
receiver frontend
LPF
reference symbol/ guard/ extraction
fast AGC
amplitude determin- ing
Corre- lator
Thres- hold
weight- ing
Max. Pos
162
164
166
168
170
172

FIG.4

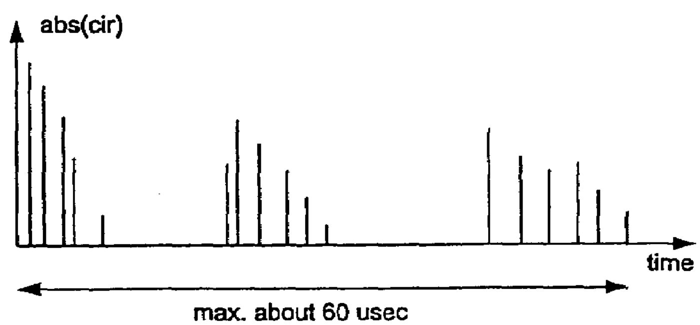
abs(cir)
time
max. about 60 usec

FIG.5

COARSE FREQUENCY SYNCHRONISATION IN MULTICARRIER SYSTEMS

FIELD OF THE INVENTION

The present invention relates to methods and apparatus for generating a signal having a frame structure, wherein each frame of the frame structure is composed of useful symbols, a guard interval associated to each useful symbol and one reference symbol. In addition, the present invention relates to methods and apparatus for frame synchronization of signals having the above structure.

The present invention is particularly useful in a MCM transmission system (MCM=Multi-carrier modulation) using an orthogonal frequency division multiplexing (OFDM) for digital broadcasting.

BACKGROUND OF THE INVENTION

In a MCM (OFDM) transmission system the binary information is represented in the form of a complex spectrum, i.e. a distinct number of complex subcarrier symbols in the frequency domain. In the modulator a bitstream is represented by a sequence of spectra. Using an inverse Fourier-transform (IFFT) a MCM time domain signal is produced from this sequence of spectra.

In case of a transmission of this described MCM signal via a multipath channel with memory, intersymbol interference (ISI) occurs due to multipath dispersion. To avoid ISI a guard interval of fixed length is added between adjacent MCM symbols in time. The guard interval is chosen as cyclic prefix. This means that the last part of a time domain MCM symbol is placed in front of the symbol to get a periodic extension. If the fixed length of the chosen guard interval is greater than the maximum multipath delay, ISI will not occur.

In the receiver the information which is in the frequency and time domain (MCM) has to be recovered from the MCM time domain signal. This is performed in two steps. Firstly, optimally locating the FFT window, thus eliminating the guard interval in front of each MCM time domain symbol. Secondly, performing a Fourier Transform of the sequence of useful time samples thus obtained.

As a result a sequence of spectral symbols is thus recovered. Each of the symbols contains a distinct number of information carrying subcarrier symbols. Out of these, the information bits are recovered using the inverse process of the modulator.

Performing the above described method, the following problem occurs in the receiver. The exact position of the guard interval and hence the position of the original useful parts of the time domain MCM symbols is generally unknown. Extraction of the guard interval and the subsequent FFT-transform of the resulting useful part of the time signal is not possible without additional information. To provide this additional information, a known (single carrier) sequence in the form of a (time domain) reference symbol is inserted into the time signal. With the knowledge about the positions of the reference symbols in the received signal, the exact positions of the guard intervals and thus the interesting information carrying time samples are known.

The periodical insertion of the reference symbol results in a frame structure of the MCM signal. This frame structure of a MCM signal is shown in FIG. **1**. One frame of the MCM signal is composed of a plurality of MCM symbols **10**. Each MCM symbol **10** is formed by an useful symbol **12** and a guard interval **14** associated therewith. As shown in FIG. **1**, each frame comprises one reference symbol **16**.

A functioning synchronization in the receiver, i.e. frame, frequency, phase, guard interval synchronization is necessary for the subsequent MCM demodulation. Consequently, the first and most important task of the base band processing in the receiver is to find and synchronize to the reference symbol.

DESCRIPTION OF THE PRIOR ART

Most prior art methods for frame synchronization have been developed for single carrier transmission over the AWGN channel (AWGN=Additive White Gaussian Noise). These prior art methods based on correlation are, without major changes, not applicable for transmission over multipath fading channels with large frequency offsets or MCM transmission systems that use, for example, an orthogonal frequency division multiplexing.

For MCM transmission systems particular frame synchronization methods have been developed.

Warner, W. D., Leung C.: OFDM/FM Frame Synchronization for Mobile Radio Data Communication, IEEE Trans. On Vehicular Technology, vol. VT-**42**, August 1993, pp. 302 to 313, teaches the insertion of reference symbols in the form of tones in parallel with the data into the MCM symbol. The reference symbols occupy several carriers of the MCM signal. In the receiver, the synchronization carriers are extracted in the frequency domain, after a FFT transform (FFT=fast Fourier transform) using a correlation detector. In the presence of large frequency offsets, this algorithm becomes very complex because several correlators must be implemented in parallel.

A further prior art technique is to insert a periodic reference symbol into the modulated MCM signal. This reference symbol is a CAZAC sequence (CAZAC=Constant Amplitude Zero Autocorrelation). Such techniques are taught by: Classen, F., Meyr, H.: Synchronization algorithms for an OFDM system *Vehic. Technology Conference,* 1997; Schmidl, T. M., Cox, D. C.: Low-Overhead, Low-Complexity [Burst] Synchronization for OFDM Transmission, *Proc. IEEE Int. Conf. on Commun.,* 1996. In such systems, the receiver's processor looks for a periodic repetition. For these algorithms coarse frequency synchronization has to be achieved prior to or at least simultaneously with frame synchronization.

Van de Beek, J, Sandell, M., Isaksson, M, Börjesson, P.: Low-Complex Frame Synchronization in OFDM Systems, *Proc. of the ICUPC,* 1995, avoid the insertion of additional reference symbols or pilot carriers and use instead the periodicity in the MCM signal which is inherent in the guard interval and the associated cyclical extension. This method is suitable only for slowly varying fading channels and small frequency offsets.

U.S. Pat. No. 5,191,576 relates to a method for the diffusion of digital data designed to be received notably by mobile receivers moving in an urban environment. In this method, the header of each frame of a broadcast signal having a frame structure has a first empty synchronization symbol and a second unmodulated wobbled signal forming a two-stage analog synchronization system. The recovery of the synchronization signal is achieved in an analog way, without prior extraction of a clock signal at the binary level.

EP 0631406 A relates to data signals, COFDM signals, for example, and to methods and apparatus for diffusing said signals. The COFDM signals comprises a sequence of symbols, each symbol having an useful portion and a guard interval. Two symbols of a COFDM signal are provided as synchronization symbols. One of the two symbols is a zero symbol, whereas the other thereof is a synchronization symbol which is formed by an unmodulated multiplex of the carrier frequencies having a constant envelope. Beside the two symbols as synchronization symbols, it is taught in EP 0631406 A to modulate the pilot frequency of the data signal with a reference signal which carries the synchronization information. This reference signal modulated on the pilot frequency of the data signal can be used by a MABLR demodulator.

WO 98/00946 A relates to a system for a timing and frequency synchronization of OFDM signals. Two OFDM training symbols are used to obtain full synchronization in less than two data frames. The OFDM training symbols are placed into the OFDM signal, preferably at least once every frame. The first OFDM training symbol is produced by modulating the even-numbered OFDM sub-carriers whereas the odd-numbered OFDM sub-carriers are suppressed. Thus, in accordance with WO 98/00946 A, the first OFDM training symbol is produced by modulating the even-numbered carriers of this symbol with a first predetermined PN sequence.

Moose: "A technique for orthogonal frequency division multiplexing frequency offset correction", IEEE TRANSACTIONS ON COMMUNICATIONS, Vo. 42, No. 10, October 1994, pages 2908 to 2914, teaches methods for correcting frequency offsets in OFDM digital communications. The methods involve repetition of a data symbol and comparison of the phases of each of the carriers between the successive symbols. The phase shift of each of the carriers between the repeated symbols is due to the frequency offset since the modulation phase values are not changed in the repeated symbols.

Keller; Hanzo: "Orthogonal frequency division multiplex synchronization techniques for wireless local area networks", IEEE INTERNATIONAL SYMPOSIUM ON PERSONAL, INDOOR AND MOBILE RADIO COMMUNICATIONS, Oct. 15, 1996, pages 963 to 967, teach frequency acquisition, frequency tracking, symbol synchronization and frame synchronization techniques. Regarding the frame synchronization, it is taught to use a reference symbol which consists of repetitive copies of a synchronization pattern of pseudo-random samples. The frame synchronization is achieved by autocorrelation techniques using the periodic synchronization segments such that for the synchronization algorithms proposed no a priori knowledge of the synchronization sequences is required.

The methods for frame synchronization available up to date require either prior achieved frequency synchronization or become very complex when the signal in the receiver is corrupted by a large frequency offset.

If there is a frequency offset in the receiver, as can easily be the case when a receiver is powered-on and the frequency synchronization loop is not yet locked, problems will occur. When performing a simple correlation there will only be noise at the output of the correlator, i.e. no maximum can be found if the frequency offset exceeds a certain bound. The size of the frequency offset depends on the length (time) of the correlation to be performed, i.e. the longer it takes, the smaller the allowed frequency offset becomes. In general, frequency offset increases implementation complexity.

Frequency offsets occur after power-on or later due to frequency deviation of the oscillators used for down-conversion to baseband. Typical accuracies for the frequency of a free running local oscillator (LO) are at ±50 ppm of the carrier frequency. With a carrier frequency in S-band (e.g. 2.34 GHz) there will be a maximum LO frequency deviation

of above 100 kHz (117.25 kHz). A deviation of this magnitude puts high demands on the above methods.

In the case of multipath impaired transmission channel, a correlation method yields several correlation maxima in addition to the distinct maximum for an AWGN channel. The best possible frame header position, i.e. the reference symbol, has to be selected to cope with this number of maxima. In multipath channels, frame synchronization methods with correlations can not be used without major changes. Moreover, it is not possible to use data demodulated from the MCM system, because the demodulation is based on the knowledge of the position of the guard interval and the useful part of the MCM symbol.

SUMMARY OF THE INVENTION

It is an object of the present invention to provide a method and an apparatus for generating a signal having a frame structure that allow a frame synchronization after the signals have been transmitted even in the case of a carrier frequency offset or in the case of a transmission via a multipath fading channel.

It is a further object of the present invention to provide a method and an apparatus for frame synchronization of a signal having a frame structure even in the case of a carrier frequency offset.

In accordance with a first aspect, the present invention provides a method for generating a signal having a frame structure, each frame of the frame structure comprising at least one useful symbol, a guard interval associated to the at least one useful symbol and a reference symbol, the method comprising the steps of performing an amplitude modulation of a bit sequence, the envelope of the amplitude modulated bit sequence defining the reference pattern of the reference symbol and inserting the amplitude modulated bit sequence into said signal as said reference symbol.

In accordance with a second aspect, the present invention provides a method for generating a multi-carrier modulated signal having a frame structure, each frame of the frame structure comprising at least one useful symbol, a guard interval associated to the at least one useful symbol and a reference symbol, the method comprising the steps of:

providing a bitstream;

mapping bits of the bitstream to carriers in order to provide a sequence of spectra;

performing an inverse Fourier transform in order to provide multi-carrier modulated symbols;

associating a guard interval to each multi-carrier modulated symbol;

generating the reference symbol by performing an amplitude modulation of a bit sequence, the envelope of the amplitude modulated bit sequence defining the reference pattern of the reference symbol;

associating the reference symbol to a predetermined number of multi-carrier modulated symbols and associated guard intervals in order to define the frame; and

inserting said amplitude modulated bit sequence into said signal as said reference symbol.

In accordance with a third aspect, the present invention provides a method for frame synchronization of a signal having a frame structure, each frame of the frame structure comprising at least one useful symbol, a guard interval associated with the at least one useful symbol and a reference symbol, the method comprising the steps of:

receiving the signal;

down-converting the received signal;

performing an amplitude-demodulation of the down-converted signal in order to generate an envelope;

correlating the envelope with a predetermined reference pattern in order to detect the signal reference pattern of the reference symbol in the signal; and

performing the frame synchronization based on the detection of the signal reference pattern.

In accordance with a fourth aspect, the present invention provides a method for frame synchronization of a multi-carrier modulated signal having frame structure, each frame of the frame structure comprising at least one useful symbol, a guard interval associated to the at least one useful symbol and a reference symbol, the method comprising the steps of:

receiving the multi-carrier modulated signal;

down-converting the received multi-carrier modulated signal;

performing an amplitude-demodulation of the down-converted multi-carrier modulated signal in order to generate an envelope;

correlating the envelope with a predetermined reference pattern in order to detect the signal reference pattern of the reference symbol in the multi-carrier modulated signal;

performing the frame synchronization based on the detection of the signal reference pattern;

extracting the reference symbol and the at least one guard interval from the down-converted received multi-carrier modulated signal based on the frame synchronization;

performing a Fourier transform in order to provide a sequence of spectra from the at least one useful symbol;

de-mapping the sequence of spectra in order to provide a bitstream.

In accordance with a fifth aspect, the present invention provides an apparatus for generating a signal having a frame structure, each frame of the frame structure comprising at least one useful symbol, a guard interval associated to the at least one useful symbol and a reference symbol, the apparatus comprising an amplitude modulator for performing an amplitude modulation of a bit sequence, the envelope of the amplitude modulated bit sequence defining the reference pattern of the reference symbol; and

means for inserting the amplitude modulated bit sequence into said signal as said reference symbol.

In accordance with a sixth aspect, the present invention provides an apparatus for generating a multi-carrier modulated signal having a frame structure, each frame of the frame structure comprising at least one useful symbol, a guard interval associated to the at least one useful symbol and a reference symbol, the apparatus comprising:

means for providing a bitstream;

means for mapping bits of the bitstream to carriers in order to provide a sequence of spectra;

means for performing an inverse Fourier transform in order to provide multi-carrier modulated symbols;

means for associating a guard interval to each multi-carrier modulated symbol;

means for generating the reference symbol by an amplitude modulator for performing an amplitude modulation of a bit sequence, the envelope of the amplitude modulated bit sequence defining the reference pattern of the reference symbol;

means for associating the reference symbol to a predetermined number of multi-carrier modulated symbols and associated guard intervals in order to define the frame; and

means for inserting the amplitude modulated bit sequence into said signal as said reference symbol.

In accordance with a seventh aspect, the present invention provides an apparatus for frame synchronization of a signal having a frame structure, each frame of the frame structure comprising at least one useful symbol, a guard interval associated to the at least one useful symbol and a reference symbol, the apparatus comprising:

receiving means for receiving the signal;

a down-converter for down-converting the received signal;

an amplitude-demodulator for performing an amplitude demodulation of the down-converted signal in order to generate an envelope;

a correlator for correlating the envelope with a predetermined reference pattern in order to detect the signal reference pattern of the reference symbol in the signal; and

means for performing the frame synchronization based on the detection of the signal reference pattern.

In accordance with a eighth aspect, the present invention provides an apparatus for frame synchronization of a multi-carrier modulated signal having a frame structure, each frame of the frame structure comprising at least one useful symbol, a guard interval associated to the at least one useful symbol and a reference symbol, the apparatus comprising:

a receiver for receiving the multi-carrier modulated signal;

a down-converter for down-converting the received multi-carrier modulated signal;

an amplitude-demodulator for performing an amplitude-demodulation of the down-converted multi-carrier modulated signal in order to generate an envelope;

a correlator for correlating the envelope with a predetermined reference pattern in order to detect the signal reference pattern of the reference symbol in the multi-carrier modulated signal;

means for performing the frame synchronization based on the detection of the signal reference pattern;

means for extracting the reference symbol and the at least one guard interval from the down-converted received multi-carrier modulated signal based on the frame synchronization in order to generate the at least one useful symbol;

means for performing a Fourier transform in order to provide a sequence of spectra from the at least one useful symbol; and

means for de-mapping the sequence of spectra in order to provide a bitstream.

The present invention provides a novel structure of the reference symbol along with a method to determine the position of the reference symbol and thus the start of a frame in a signal having a frame structure as shown for example in FIG. 1.

The invention relates to a method for finding frame headers independently of other synchronization information and thus for positioning the FFT windows correctly. This includes the extraction of a guard interval. The method is based on the detection of a known reference symbol of the frame header in the reception signal, e.g. in the digital complex baseband. The new frame synchronization will be performed as the first synchronization task.

Synchronization to the reference symbol, i.e. the frame header is the first step to initiate radio reception. The reference symbol is structured to accomplish this. The information contained in the reference symbol must therefore be independent of other synchronization parameters, e.g. frequency offset. For this reason, in accordance with the present invention, the form of the reference symbol selected is an amplitude modulated sequence (AM sequence) in the complex baseband. Thus, the information contained in the reference symbol is only that given in the amplitude and not that in the phase. Note that the phase information will be corrupted by a possible frequency offset. In preferred embodiments of the present invention, the AM information is constructed from a bit sequence with special features. The information sequence is selected in a way which makes it easy and secure to find it in the time domain. A bit sequence with good autocorrelation properties is chosen. Good autocorrelation properties means a distinct correlation maximum in a correlation signal which should be as white as possible.

A pseudo random bit sequence (PRBS) having good autocorrelation properties meets the above requirements.

Using the envelope of the signal to carry bit information offers additional flexibility. First it has to be decided which envelope values should correspond to the binary values of 0 and 1. The parameters are mean amplitude and modulation rate. Attention should be paid to selecting the mean amplitude of the reference symbol (performance) identically to the mean amplitude of the rest of the frame. This is due to the amplitude normalization (AGC; AGC=Automatic Gain Control) performed in the receiver. It is also possible to select the mean amplitude of the reference symbol higher than the mean signal amplitude, but then care has to be taken that the time constant of the AGC (1/sensitivity) is selected high enough to secure that the strong (boosted) signal of the reference symbol does not influence the AGC control signal and thus attenuate the signal following the reference symbol.

Another degree of freedom can be characterized as modulation degree d. This parameter is responsible for the information density of the modulating signal mod(t) formed out of the binary sequence bin(t) as follows: mod(t)=bin(t/d). This modulation degree can be chosen as free parameter fixed by an integer or real relation to the sampling rate. It is appropriate to choose the modulation degree d as an integer value because of the discrete values of the binary sequence:

d = 1: mod (m)	= bin (m)	
d = 2: mod (m)	= bin (m/2)	for m even
	= bin_int (m/2)	for m odd
d = 3: mod (m)	= bin (m/3)	for m = 0, ±3, ±6, ±9,
	bin_int (m/3)	else

The signal values bin_int(m/d) are computed from the binary sequence bin(m) by ideal interpolation (between the discrete integer values m) with the factor of d. This is similar to an ideal sampling rate expansion (with sin(x)/x interpolation), but the sampling rate remains, only less bits of the binary sequence bin(m) correspond to the resulting interpolated sequence mod(m). This parameter m indicates the discrete time.

With increasing m the modulating signal mod(t) is expanded in time relative to the basic binary sequence, this results in a bandwidth compression of the resulting AM spectrum with regard to the basic binary sequence. A time expansion by a factor 2 results in a bandwidth compression by the same factor 2. In addition to the bandwidth compression, a further advantage of a higher modulation degree d is a reduced complexity of the search method in the receiver due to the fact that only each dth sample has a corresponding binary value. Choosing the factor d=1 is not preferred since this would result in aliasing due to disregard of the sampling theorem. For this reason, in a preferred embodiment of the present invention d is chosen to be 2.

The choice of length and repetition rate of the reference symbol is, on the one hand, dominated by the channel properties, e.g. the channel's coherence time. On the other hand the choice depends on the receiver requirements concerning mean time for initial synchronization and mean time for resynchronization after synchronization loss due to a channel fade.

In the receiver, the first step after the down-conversion of the received signal is to perform an amplitude-demodulation of the down-converted signal in order to generate an envelope, i.e. in order to determine the amplitude of the signal. This envelope is correlated with a replica reference pattern in order to detect the signal reference pattern of the reference symbol in the signal. In the case of a AWGN channel, the result of this correlation will be a white noise signal with zero mean value and with a clearly visible (positive) maximum. In the case of a multipath channel, several maxima will occur in the correlation signal computed by this correlation. In the former case, the location of the reference symbol is determined based on the signal maximum, whereas in the latter case a weighting procedure is performed in order to find out the maximum corresponding to the location of the reference symbol.

Thus, the present invention shows how to find a reference symbol by a detection method which is simple. Furthermore, the present invention can be used for one-carrier or multi-carrier systems. The present invention is particularly useful in multi-carrier modulation systems using an orthogonal frequency division multiplexing, for example in the field of digital broadcasting. The synchronization methods according to the present invention are independent of other synchronization steps. Since the information needed for the synchronization is contained in the envelope of the preamble, i.e. the reference symbol, the reference symbol is independent of possible frequency offsets. Thus, a derivation of the correct down sampling timing and the correct positioning of the FFT window can be achieved. The reference symbol of the present invention can be detected even if the frequency synchronization loop is not yet locked or even in the case of a carrier frequency offset. The frame synchronization method in accordance with the present invention is preferably performed prior to other and without knowledge of other synchronization efforts.

BRIEF DESCRIPTION OF THE DRAWINGS

In the following, preferred embodiments of the present invention will be explained in detail on the basis of the drawings enclosed, in which:

FIG. **1** shows a schematic view of a signal having a frame structure;

FIG. **2** shows a block diagram of a MCM system to which the present invention can be applied;

FIG. **3** shows a schematic block diagram of a frame and frequency synchronization system in a MCM receiver;

FIG. **4** shows a schematic diagram of an apparatus for frame synchronization; and

FIG. **5** shows a typical channel impulse response of a single frequency network in S-band.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

Although the present invention is explained mainly referring to a MCM system, it is obvious that the present invention can be used in connection with different signal transmissions that are based on different kinds of modulation.

FIG. **2** shows a MCM system overview on the basis of which the present invention will be described in detail. At **100** a MCM transmitter is shown that substantially corresponds to a prior art MCM transmitter except for the kind of the reference symbol being added to each frame of a MCM signal. A description of such a MCM transmitter can be found, for example, in William Y. Zou, Yiyan Wu, "COFDM: AN OVERVIEW", IEEE Transactions on Broadcasting, vol. 41, No. **1**, March 1995.

A data source **102** provides a serial bitstream **104** to the MCM transmitter. The incoming serial bitstream **104** is applied to a bit-carrier mapper **106** which produces a sequence of spectra **108** from the incoming serial bitstream **104**. An inverse fast Fourier transform (FFT) **110** is performed on the sequence of spectra **108** in order to produce a MCM time domain signal **112**. The MCM time domain signal forms the useful MCM symbol of the MCM time signal. To avoid intersymbol interference (ISI) caused by multipath distortion, a unit **114** is provided for inserting a guard interval of fixed length between adjacent MCM symbols in time. In accordance with a preferred embodiment of the present invention, the last part of the useful MCM symbol is used as the guard interval by placing same in front of the useful symbol. The resulting MCM symbol is shown at **115** in FIG. **2** and corresponds to the MCM symbol **10** depicted in FIG. **1**. signal transmitted through the channel **122** is received at the receiver front end **132**. The down-converted MCM signal is sampled at the receiver front end **132** and is, in the preferred embodiment, provided to a fast running automatic gain control (time constant<MCM symbol duration) in order to eliminate fast channel fluctuations (channel coherence time=MCM symbol duration). The fast AGC **162** is used in addition to the normally slow AGC in the signal path, in the case of transmission over a multipath channel with long channel impulse response and frequency selective fading. The fast AGC adjusts the average amplitude range of the signal to the known average amplitude of the reference symbol. The so processed symbol is provided to an amplitude determining unit **164**.

The amplitude determining unit **164** can use the simple alpha$_{max+}$ beta$_{min-}$ method in order to calculate the amplitude of the signal. This method is described for example in Palachels A.: DSP-mP Routine Computes Magnitude, EDN, Oct. 26, 1989; and Adams, W. T., and Bradley, J.: Magnitude Approximations for Microprocessor Implementation, IEEE Micro, Vol. 3, No. 5, October 1983.

The output signal of the amplitude determining unit **164** is applied to a correlator **166**. In the correlator **166**, a cross correlation between the amplitude signal output from the amplitude determining unit **164** and a known ideal amplitude information is computed. The known ideal amplitude information is stored in the correlator. For both, the amplitude and the known ideal amplitude information, their amplitudes are symmetrically to zero relative to their average amplitude.

In the ideal AWGN case, the result will be a white noise signal with zero mean value and with a clearly visible positive maximum. In this ideal AWGN case, the position of the single maximum is evaluated in a maximum position unit **172**. On the basis of this evaluation, the reference symbol and the guard intervals are extracted from the MCM signal in a combined reference symbol/guard extraction unit **136**/**138**. Although these units are shown as a combined unit **136**/**138** in FIG. **4**, it is clear that separate units can be provided. The MCM signal is transmitted from the RF front end **150** to the reference symbol/guard extraction unit **136**/**138** via a low pass filter **174**.

In the case of time spreading encountered in a multipath channel, several maxima corresponding to the number of clusters in the channel impulse response occur in the output signal of the correlator. A schematic view of three such clusters located in a time window of maximum about 60 microseconds is shown in FIG. **5**. Out of the several maxima caused by the time spreading encountered in a multipath channel, the best one has to be selected as the position of the frame header, i.e. the reference symbol. Therefore, a threshold unit **168** and a weighting unit **170** are provided between the correlator **166** and the maximum position unit **172**. The threshold unit **168** is provided to remove maxima having an amplitude below a predetermined threshold. The weighting unit **164** is provided in order to perform a weighting procedure on the remaining maxima such that the maximum corresponding to the reference symbol can be determined. An examplary weighting procedure performed in the weighting unit **170** is as follows.

The first significant maximum is considered to be the best one. The output signal of the correlator is observed from the first detected maximum onwards for the maximum length of the channel impulse response and an amplitude weighting function is applied to the signal. Because the actual channel impulse response length is unknown, the following fact can be remembered. During system design, the length of the channel impulse response has to be investigated. In a MCM system, the guard interval shall be equal or longer than the maximum expected channel impulse response. For this reason, the part (interval with l_l samples, l_l corresponding to the maximum expected channel impulse response, i.e. the guard interval length) of the correlation output signal starting with the first maximum,

$$I_{k0}(n)=r(k_0+n),\ 0\leq n\leq l_l-1 \qquad \text{(Eq. 1)}$$

with k_0 being the position of the first maximum, will be examined to find the best frame start position. The above signal part is weighted with the function

$$W(n)=10^{-\frac{\text{weight\_dB}}{10}\frac{n}{l_l-1}} \qquad \text{(Eq. 2)}$$

The position (n_{max}) of the maximum in the resulting signal interval

$$I_{k,weighted}(n)=[r(k_0+n)W(n)]=\left[r(k_0+n)10^{-\frac{\text{weight\_dB}}{10}\frac{n}{l_l-1}}\right] \qquad \text{(Eq. 3)}$$

$$0\leq n\leq l_l-1$$

will be chosen as best frame start position.

r(k) designates the output signal of the correlator (**166**) at the time k. The signal is present with a clock frequency which is determined by the multiplication: oversampling factor*subcarrier symbol frequency. The parameter k designates the discrete time in sample clocks. This signal is windowed with information from the threshold unit **168**. An interval having the length of l_I values is extracted from the signal r(k). The first value being written into the interval is the correlation start value at the time k_0, at which the output value $r(k_0)$ exceeds the threshold value of the threshold unit **168** for the first time. The interval with the windowed signal is designated by the term $I(k_0)$. The parameter n designates the relative time, i.e. position, of a value inside the interval.

Using the described weighting operation, the earlier correlation maxima are more likely to be chosen as right frame start position. A later coming maximum will only be chosen as frame start position, if the value of the maximum is significantly higher than the earlier one. This operation is applicable especially for MCM, because here it is better to detect the frame start positions some samples too early than some samples too late. Positioning the frame start some samples too early leads to positioning the FFT window a little bit into the guard interval, this contains information of the same MCM symbol and therefore leads to little effects. If the frame start position is detected some samples too late, then the FFT window includes some samples of the following guard interval.

This leads to a more visible degradation, because the following guard interval contains information of the following MCM symbol (ISI occurs).

It is important to know that the first visible correlation maximum after receiver power-on does not necessarily correspond to the first CIR (channel impulse response) cluster. It is possible that it is corresponding to a later cluster, see FIG. **5**. For this reason during power-on one should wait for a second frame start before starting demodulation.

It is clear that amplitude determining methods different from the described alpha_{max+} beta_{min-} method can be used. For simplification, it is possible to reduce the amplitude calculation to a detection as to whether the current amplitude is above or below the average amplitude. The output signal then consists of a −1/+1 sequence which will be correlated with a known bit sequence, also in −1/+1 values. This correlation can easily be performed using a simple integrated circuit (IC).

In addition, an oversampling of the signal received at the RF front end can be performed. For example, the received signal can be expressed with two times oversampling.

This oversampled signal is passed to a fast running AGC to eliminate fast channel fluctuations before the amplitude of the signal is calculated. The amplitude information will be hard quantized. Values larger than the mean amplitude, mean amplitude is 1, will be expressed as +1, values smaller than the mean amplitude will be expressed as −1. This −1/+1 signal is passed to the correlator that performs a cross correlation between the quantized signal and the stored ideal amplitude values of the reference symbol:

amp__sto(k)=2*bin(k/4),

if k=2(oversampling factor)*2(interpolation factor)* 1,2,3 . . . 92

(92 for 184 reference symbol and interpolation factor 2)

amp__sto(k)=0, else, k<=2(oversampling factor)* plary weighting procedure performed in the weighting unit **170** is as follows.

The first significant maximum is considered to be the best one. The output signal of the correlator is observed from the first detected maximum onwards for the maximum length of the channel impulse response and an amplitude weighting function is applied to the signal. Because the actual channel impulse response length is unknown, the following fact can be remembered. During system design, the length of the channel impulse response has to be investigated. In a MCM

system, the guard interval shall be equal or longer than the maximum expected channel impulse response. For this reason, the part (interval with ll samples, ll corresponding to the maximum expected channel impulse response, i.e. the guard interval length) of the correlation output signal starting with the first maximum,

$$I_{k0}(n)=r(k_0+n),\ 0\#\ n\ \#\ l_I-1 \qquad \text{(Eq.1)}$$

with k_0 being the position of the first maximum, will be examined to find the best frame start position. The above signal part is weighted with the function

$$W(n) = 10^{-\frac{\text{weight\_dB}}{10}\frac{n}{l_I-1}} \qquad \text{(Eq. 2)}$$

(Eq.2)

The position (n_{max}) of the maximum in the resulting signal interval

$$I_{k0,weighted}(n) = [r(k_0+n)W(n)] = \left[r(k_0+n)10^{-\frac{\text{weight\_dB}}{10}\frac{n}{l_I-1}}\right] \qquad \text{(Eq. 3)}$$

$$0\ \#\ n\ \#\ l_I\ 1$$

(Eq.3)

will be chosen as best frame start position.

r(k) designates the output signal of the correlator (**166**) at the time k. The signal is present with a clock frequency which is determined by the multiplication: oversampling factor*subcarrier symbol frequency. The parameter k designates the discrete time in sample clocks. This signal is windowed with information from the threshold unit **168**. An interval having the length of l_I values is extracted from the signal r(k). The first value being written into the interval is the correlation start value at the time k_0, at which the output value $r(k_0)$ exceeds the threshold value of the threshold unit **168** for the first time. The interval with the windowed signal is designated by the term $I(k_0)$. The parameter n designates the relative time, i.e. position, of a value inside the interval.

Using the described weighting operation, the earlier correlation maxima are more likely to be chosen as right frame start position. A later coming maximum will only be chosen as frame start position, if the value of the maximum is significantly higher than the earlier one. This operation is applicable especially for MCM, because here it is better to detect the frame start positions some samples too early than some samples too late. Positioning the frame start some samples too early leads to positioning the FFT window a little bit into the guard interval, this contains information of the same MCM symbol and therefore leads to little effects. If the frame start position is detected some samples too late, then the FFT window includes some samples of the following guard interval. This leads to a more visible degradation, because the following guard interval contains information of the following MCM symbol (ISI occurs).

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It is clear that amplitude determining methods different from the described alpha_{+} beta_{min-} method can be used. For simplification, it is possible to reduce the amplitude calculation to a detection as to whether the current amplitude is above or below the average amplitude. The output signal then consists of a −1/+1 sequence which will be correlated with a known bit sequence, also in −1/+1 values. This correlation can easily be performed using a simple integrated circuit (IC).

In addition, an oversampling of the signal received at the RF front end can be performed. For example, the received signal can be expressed with two times oversampling.

This oversampled signal is passed to a fast running AGC to eliminate fast channel fluctuations before the amplitude of the signal is calculated. The amplitude information will be hard quantized. Values larger than the mean amplitude, mean amplitude is 1, will be expressed as +1, values smaller than the mean amplitude will be expressed as −1. This −1/+1 signal is passed to the correlator that performs a cross correlation between the quantized signal and the stored ideal amplitude values of the reference symbol:

amp__sto(k)=2*bin(k/4),

if k=2(oversampling factor)*2(interpolation factor)*1,2,3 . . . 92

(92 for 184 reference symbol and interpolation factor 2)

amp__sto(k)=0, else, k<=2(oversampling factor)

*2(interpolation factor)*92

(first part of amp__sto=[0 0 0 −1 0 0 0 1 0 0 0 1 0 0 0 −1 0 . . .]).

With this algorithm a correlation maximum of 92 is achievable.

Again, the maxima in the correlator output signal correspond to different frame start positions due to different multipath clusters. In this signal with various maxima the best frame start position has to be chosen. This is done in the following steps: The output of the correlator is given to a threshold detection. If the signal first time exceeds the threshold (a threshold of 50 has proved to be applicable) the best position search algorithm is initialized. The correlator output signal in the interval following the threshold exceeding value will be weighted with the weighting function, see above. The position of the resulting maximum in the weighted signal will be chosen as best frame start position. With the knowledge about the best frame start position the guard interval extraction and the following MCM demodulation will be performed.

Some more efforts can be carried out to increase frame synchronization accuracy. These methods will be explained in the following.

A postprocessing of the frame start decision is performed in order a) to increase the reliability of the frame synchronization; b) to secure that no frame start position is disregarded; and c) to optimize the frame start position in case of varying CIR cluster positions.

Using information of other frame start positions. It is known that in front of each frame a reference symbol is inserted into the signal. If the position of the currently detected frame start has changed significantly regarding the last detected frame start, demodulation of the two frames in total and completely independent from each other is possible. It is also possible to buffer the last signal frame and to perform the required shift of the frame start position step by step with the MCM symbols of the frame. This results in an interpolative positioning of the single MCM symbols including simultaneous asynchronous guard interval extraction for the different MCM symbols.

Such an interpolative positioning of the FFT window is also possible if one frame start position is missing, i.e. the frame start has not been detected. If one frame start position is missing the guard interval extraction can be performed the same way as in the frame before without large performance degradation. This is due to the normally only slowly varying CIR cluster positions, but only if the signal strength is good enough. Stopping demodulation and waiting for the next detected frame start position is also imaginable but not desirable because of the long interrupt.

What follows is an example of a reference symbol of 184 samples (subcarrier symbols) as provided by the inventive apparatus for generating a signal having a frame structure.

The underlying binary sequence of length 92 is:

```
bin = [0 1 1 0 1 1 0 1 0 1 1 0 1 0 1 0
       0 0 1 1 1 0 0 0 0 0 0 0 0 1 1 0
       1 1 1 1 1 0 0 0 1 1 1 0 0 0 0 0
       0 0 1 1 1 0 1 1 1 0 0 1 1 0 1 1
       1 0 1 1 0 1 0 1 0 1 1 0 1 1 0 1
       1 0 1 0 0 0 0 1 0 1 1 0]
```

The modulated binary sequence is:

```
i__q =   [0.5 1.5 1.5 0.5 1.5 1.5 0.5 1.5 0.5 1.5 1.5 0.5 1.5
          0.5 1.5 0.5 0.5 0.5 1.5 1.5 1.5 0.5 0.5 0.5 0.5 0.5
          0.5 0.5 0.5 1.5 1.5 0.5 1.5 1.5 1.5 1.5 1.5 0.5 0.5
          0.5 1.5 1.5 1.5 0.5 0.5 0.5 0.5 0.5 0.5 0.5 1.5 1.5
          1.5 0.5 1.5 1.5 1.5 0.5 0.5 1.5 1.5 0.5 1.5 1.5 1.5
          0.5 1.5 1.5 0.5 1.5 0.5 1.5 0.5 1.5 1.5 0.5 1.5 1.5
          0.5 1.5 0.5 1.5 0.5 0.5 0.5 0.5 1.5 0.5 1.5 1.5 0.5]
```

This modulated binary sequence i__q is interpolated in order to produce an interpolated sequence i__q__int:

```
i__q__int =   [0.5000 1.0635 1.5000 1.7195 1.5000 0.8706 0.5000
               0.8571 1.5000 1.7917 1.5000 0.8108 0.5000 1.0392
               1.5000 1.0392 0.5000 0.8108 1.5000 1.7984 1.5000
               0.8108 0.5000 1.0460 1.5000 0.9997 0.5000 0.9603
               1.5000 1.1424 0.5000 0.3831 0.5000 0.4293 0.5000
               0.9997 1.5000 1.5769 1.5000 1.5769 1.5000 1.0065
               0.5000 0.3899 0.5000 0.5325 0.5000 0.4931 0.5000
               0.4999 0.5000 0.4931 0.5000 0.5325 0.5000 0.3967
               0.5000 0.9603 1.5000 1.7522 1.5000 0.8571 0.5000
               0.8965 1.5000 1.6422 1.5000 1.4669 1.5000 1.4737
               1.5000 1.6096 1.5000 0.9929 0.5000 0.4226 0.5000
               0.4226 0.5000 0.9997 1.5000 1.5769 1.5000 1.5769
               1.5000 1.0065 0.5000 0.3899 0.5000 0.5325 0.5000
               0.4931 0.5000 0.4931 0.5000 0.5325 0.5000 0.3899
               0.5000 1.0065 1.5000 1.5701 1.5000 1.6096 1.5000
               0.8965 0.5000 0.8965 1.5000 1.6096 1.5000 1.5633
               1.5000 1.0392 0.5000 0.2867 0.5000 0.9929 1.5000
               1.7454 1.5000 0.8571 0.5000 0.9033 1.5000 1.6028
               1.5000 1.6028 1.5000 0.9033 0.5000 0.8503 1.5000
               1.7917 1.5000 0.8108 0.5000 1.0460 1.5000 0.9929
               0.5000 0.9929 1.5000 1.0460 0.5000 0.8108 1.5000
               1.7917 1.5000 0.8571 0.5000 0.8571 1.5000 1.7849
               1.5000 0.8571 0.5000 0.8571 1.5000 1.7917 1.5000
               0.8176 0.5000 1.0065 1.5000 1.1424 0.5000 0.3436
               0.5000 0.5788 0.5000 0.3436 0.5000 1.1424 1.5000
               1.0065 0.8312 1.5000 1.7263 1.5000 1.0635 0.5000
               0.0637]
```

*amp__int=i__q__int+j*i__q__int*

amp__int is the reference symbol inserted periodically into the signal after the guard interval insertion.

As it is clear from the above specification, the present invention provides methods and apparatus for generating a signal having a frame structure and methods and apparatus for frame synchronization when receiving such signals which are superior when compared with prior art systems. The frame synchronization algorithm in accordance with the present invention provides all of the properties shown in Table 1 in contrary to known frame synchronization procedures. Table 1 shows a comparison between the system in accordance with the present invention using an AM sequence as reference symbol and prior art systems (single carrier and MCM Eureka 147).

TABLE 1

	Single carrier (e.g. QPSK like WS)	MCM Eureka 147	MCM with AM sequence
Carrier offset allowed	no	yes	yes
Constant power achieved at Rx input	yes	no	yes
Coarse frequency offset estimation possible	no	no	yes
Coarse channel estimation possible (cluster estimation)	yes	no	yes

As can be seen from Table 1 different synchronization tasks and parameters can be derived using the frame synchronization with an AM sequence in accordance with the present invention. The frame synchronization procedure MCM Eureka 147 corresponds to the procedure described in U.S. Pat. No. 5,191,576.

What is claimed is:

1. A method for generating a signal having a frame structure, each frame of said frame structure comprising at least one useful symbol, a guard interval associated to said at least one useful symbol and a reference symbol, said method comprising the step of

performing an amplitude modulation of a bit sequence, an envelope of the amplitude modulated bit sequence defining a reference pattern of said reference symbol; and

inserting, in time domain, the reference symbol into said signal, wherein said reference symbol comprises a real part and an imaginary part, said real part and said imaginary part being equal and being formed by said amplitude modulated bit sequence.

2. The method according to claim **1**, wherein said signal is an orthogonal frequency division multiplexed signal.

3. The method according to claim **1**, wherein said amplitude modulation is performed such that a mean amplitude of said reference symbol substantially corresponds to a mean amplitude of the remaining signal.

4. The method according to claim **1**, wherein said bit sequence is a pseudo random bit sequence having good autocorrelation characteristics.

5. The method according to claim **1**, wherein a number of useful symbols in each frame is defined depending on channel properties of a channel through which the signal or a multi-carrier modulated signal is transmitted.

6. A method for generating a multi-carrier modulated signal having a frame structure, each frame of said frame structure comprising at least one useful symbol, a guard interval associated to said at least one useful symbol and a reference symbol, said method comprising the steps of:

providing a bitstream;

mapping bits of said bitstream to carriers in order to provide a sequence of spectra;

performing an inverse Fourier transform in order to provide multi-carrier modulated symbols;

associating a guard interval to each multi-carrier modulated symbol;

generating said reference symbol by performing an amplitude modulation of a bit sequence, an envelope of the amplitude modulated bit sequence defining a reference pattern of said reference symbol;

associating said reference symbol to a predetermined number of multi-carrier modulated symbols and associated guard intervals in order to define said frame; and

inserting, in time domain, said reference symbol into said signal, wherein said reference symbol comprises a real part and an imaginary part, said real part and said imaginary part being equal and being formed by said amplitude modulated bit sequence.

7. The method according to claim **6**, wherein said multi-carrier modulated signal is an orthogonal frequency division multiplex signal.

8. The method according to claim **6**, wherein said amplitude modulation is performed such that a mean amplitude of said reference symbol substantially corresponds to a mean amplitude of the remaining multi-carrier modulated signal.

9. A method for frame synchronization of a signal having a frame structure, each frame of said frame structure comprising at least one useful symbol, a guard interval associated with said at least one useful symbol and a reference symbol, said reference symbol comprising a real part and an imaginary part, said real part and said imaginary part being equal and being formed by an amplitude modulated bit sequence, said method comprising the steps of:

receiving said signal;

down-converting said received signal;

in time domain, performing an amplitude-demodulation of said down-converted signal in order to generate an envelope;

in time domain, correlating said envelope with a predetermined reference pattern in order to detect a signal reference pattern of said reference symbol in said signal; and

performing said frame synchronization based on the detection of said signal reference pattern.

10. The method according to claim **9**, further comprising the step of performing a fast automatic gain control of said received down-converted signal prior to the step of performing said amplitude-demodulation.

11. The method according to claim **9**, wherein the step of performing said amplitude-demodulation comprises the step of calculating an amplitude of said signal using the alpha_{max+} beta_{min-} method.

12. The method according to claim **9**, further comprising the steps of sampling respective amplitudes of said received down-converted signal and comparing said sampled amplitudes with a predetermined threshold in order to generate a bit sequence in order to perform said amplitude demodulation.

13. The method according to claim **12**, wherein the step of sampling respective amplitudes of said received down-converted signal further comprises the step of performing an over-sampling of said received down-converted signal.

14. The method according to claim **9**, further comprising the step of applying a result of the frame synchronization for a frame in said signal to at least one subsequent frame in said signal.

15. The method according to claim **9**, further comprising the step of detecting a location of said signal reference pattern based on an occurrence of a maximum of a correlation signal when correlating said envelope with said predetermined reference pattern.

16. The method according to claim **15**, further comprising the steps of:

weighting a plurality of maxima of said correlation signal such that a maximum occurring first is weighted stronger than any subsequently occurring maximum; and

detecting said location of said signal reference pattern based on the greatest one of said weighted maxima.

17. The method according to claim **16**, further comprising the step of:

disabling the step of performing said frame synchronization for a predetermined period of time after having switched-on a receiver performing said method for frame synchronization.

18. A method for frame synchronization of a multi-carrier modulated signal having frame structure, each frame of said frame structure comprising at least one useful symbol, a guard interval associated to said at least one useful symbol and a reference symbol, said reference symbol comprising a real part and an imaginary part, said real part and said imaginary part being equal and being formed by an amplitude modulated bit sequence, said method comprising the steps of:

receiving said multi-carrier modulated signal;

down-converting said received multi-carrier modulated signal;

in time domain, performing an amplitude-demodulation of said down-converted multi-carrier modulated signal in order to generate an envelope;

in time domain, correlating said envelope with a predetermined reference pattern in order to detect a signal reference pattern of said reference symbol in said multi-carrier modulated signal;

performing said frame synchronization based on the detection of said signal reference pattern;

extracting said reference symbol and said at least one guard interval from said down-converted received multi-carrier modulated signal based on said frame synchronization;

performing a Fourier transform in order to provide a sequence of spectra from said at least one useful symbol; and

de-mapping said sequence of spectra in order to provide a bitstream.

19. The method according to claim **18**, further comprising the step of performing a fast automatic gain control of said received down-converted multi-carrier modulated signal prior to the step of performing said amplitude-demodulation.

20. The method according to claim **18**, wherein the step of performing said amplitude-demodulation comprises the step of calculating an amplitude of said multi-carrier modulated signal using the alpha$_{max+}$ beta$_{min-}$ method.

21. The method according to claim **18**, further comprising the steps of sampling respective amplitudes of said received down-converted multi-carrier modulated signal and comparing said sampled amplitudes with a predetermined threshold in order to generate a bit sequence in order to perform said amplitude demodulation.

22. The method according to claim **21**, wherein the step of sampling respective amplitudes of said received down-converted multi-carrier modulated signal further comprises the step of performing an over-sampling of said received down-converted multi-carrier modulated signal.

23. The method according to claim **18**, further comprising the step of applying a result of the frame synchronization for a frame in said signal to at least one subsequent frame in said multi-carrier modulated signal.

24. An apparatus for generating a signal having a frame structure, each frame of said frame structure comprising at least one useful symbol, a guard interval associated to said at least one useful symbol and a reference symbol, said apparatus comprising:

an amplitude modulator for performing an amplitude modulation of a bit sequence, an envelope of the amplitude modulated bit sequence defining a reference pattern of said reference symbol; and

means for inserting, in time domain, the reference symbol into said signal, wherein said reference symbol comprises a real part and an imaginary part, said real part and said imaginary part being equal and being formed by said amplitude modulated bit sequence.

25. The apparatus according to claim **24**, wherein said signal is an orthogonal frequency division multiplexed signal.

26. The apparatus according to claim **24**, wherein a mean amplitude of said reference symbol substantially corresponds to a mean amplitude of the remaining signal.

27. The apparatus according to claim **24**, comprising means for determining a number of useful symbols in each frame depending on channel properties of a channel through which the signal or a multi-carrier modulated signal is transmitted.

28. An apparatus for generating a multi-carrier modulated signal having a frame structure, each frame of said frame structure comprising at least one useful symbol, a guard interval associated to said at least one useful symbol and a reference symbol, said apparatus comprising:

means for providing a bitstream;

means for mapping bits of said bitstream to carriers in order to provide a sequence of spectra;

means for performing an inverse Fourier transform in order to provide multi-carrier modulated symbols;

means for associating a guard interval to each multi-carrier modulated symbol;

means for generating said reference symbol comprising an amplitude modulator for performing an amplitude modulation of a bit sequence, an envelope of the amplitude modulated bit sequence defining a reference pattern of said reference symbol;

means for associating said reference symbol to a predetermined number of multi-carrier modulated symbols and associated guard intervals in order to define said frame; and

means for inserting, in time domain, the reference symbol into said signal, wherein said reference symbol comprises a real part and an imaginary part, said real part and said imaginary part being equal and being formed by said amplitude modulated bit sequence.

29. The apparatus according to claim **28**, wherein said multi-carrier modulated signal is an orthogonal frequency division multiplex signal.

30. The apparatus according to claim **28**, wherein said means for generating said reference symbol performs the amplitude modulation such that a mean amplitude of said reference symbol substantially corresponds to a mean amplitude of the remaining multi-carrier modulated signal.

31. The apparatus according to claim **28**, wherein said means for generating said reference symbol generates a pseudo random bit sequence having good autocorrelation characteristics as said bit sequence.

32. An apparatus for frame synchronization of a signal having a frame structure, each frame of said frame structure comprising at least one useful symbol, a guard interval associated to said at least one useful symbol and a reference symbol, said reference symbol comprising a real part and an imaginary part, said real part and said imaginary part being equal and being formed by an amplitude modulated bit sequence, said apparatus comprising:

receiving means for receiving said signal;

a down-converter for down-converting said received signal;

an amplitude-demodulator for performing, in time domain, an amplitude demodulation of said down-converted signal in order to generate an envelope;

a correlator for correlating, in time domain, said envelope with a predetermined reference pattern in order to detect a signal reference pattern of said reference symbol in said signal; and

means for performing said frame synchronization based on the detection of said signal reference pattern.

33. The apparatus according to claim **32**, further comprising means for performing a fast automatic gain control of said received down-converted signal preceding said amplitude-demodulator.

34. The apparatus according to claim **32**, wherein said amplitude-demodulator comprises means for calculating an amplitude of said signal using the alpha_{max+} beta_{min-} method.

35. The apparatus according to claim **32**, further comprising means for sampling respective amplitudes of said received down-converted signal, wherein said amplitude-demodulator comprises means for comparing said sampled amplitudes with a predetermined threshold in order to generate a bit sequence.

36. The apparatus according to claim **35**, wherein said means for sampling comprises means for over-sampling said received down-converted signal.

37. The apparatus according to claim **32**, further comprising means for applying a result of the frame synchronization for a frame in said signal to at least one subsequent frame in said signal.

38. The apparatus according to claim **32**, further comprising means for detecting a location of said signal reference pattern based on an occurrence of a maximum of a correlation signal output of said correlator.

39. The apparatus according to claim **38**, further comprising means for weighting a plurality of maxima of said correlation signal such that a maximum occurring first is weighted stronger than any subsequently occurring maximum; and

means for detecting said location of said signal reference pattern based on the greatest one of said weighted maxima.

40. The apparatus according to claim **39**, further comprising means for disabling said means for performing said frame synchronization for a predetermined period of time after having switched-on a receiver comprising said apparatus for frame synchronization.

41. An apparatus for frame synchronization of a multi-carrier modulated signal having a frame structure, each frame of said frame structure comprising at least one useful symbol, a guard interval associated to said at least one useful symbol and a reference symbol, said reference symbol comprising a real part and an imaginary part, said real part and said imaginary part being equal and being formed by an amplitude modulated bit sequence,

said apparatus comprising:

a receiver for receiving said multi-carrier modulated signal;

a down-converter for down-converting said received multi-carrier modulated signal;

an amplitude-demodulator for performing, in the time domain, an amplitude-demodulation of said down-converted multi-carrier modulated signal in order to generate an envelope;

a correlator for correlating, in the time domain, said envelope with a predetermined reference pattern in order to detect a signal reference pattern of said reference symbol in said multi-carrier modulated signal;

means for performing said frame synchronization based on the detection of said signal reference pattern;

means for extracting said reference symbol and said at least one guard interval from said down-converted received multi-carrier modulated signal based on said frame synchronization in order to generate said at least one useful symbol;

means for performing a Fourier transform in order to provide a sequence of spectra from said at least one useful symbol; and

means for de-mapping said sequence of spectra in order to provide a bitstream.

42. The apparatus according to claim **41**, further comprising means for performing a fast automatic gain control of said received down-converted multi-carrier modulated signal preceding said amplitude-demodulator.

43. The apparatus according to claim **41**, wherein said amplitude-demodulator comprises means for calculating an amplitude of said multi-carrier modulated signal using the alpha_{max+} beta_{min-} method.

44. The apparatus according to claim **41**, further comprising means for sampling respective amplitudes of said received down-converted multi-carrier modulated signal, wherein said amplitude-demodulator comprises means for comparing said sampled amplitudes with a predetermined threshold in order to generate a bit sequence.

45. The apparatus according to claim **44**, wherein said means for sampling comprises means for over-sampling said received down-converted multi-carrier modulated signal.

46. The apparatus according to claim **41**, further comprising means for applying a result of the frame synchronization for a frame in said multi-carrier modulated signal to at least one subsequent frame in said multi-carrier modulated signal.

* * * * *

Exhibit D

US007061997B1

(12) United States Patent
Eberlein et al.

(10) **Patent No.: US 7,061,997 B1**
(45) **Date of Patent: Jun. 13, 2006**

(54) **METHOD AND APPARATUS FOR FINE FREQUENCY SYNCHRONIZATION IN MULTI-CARRIER DEMODULATION SYSTEMS**

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(73) Assignee: **Fraunhofer-Gesellschaft zur Foerderung der angewandten Forschung e.V.**, Munich (DE)

(*) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 0 days.

(21) Appl. No.: **09/673,270**

(22) PCT Filed: **Apr. 14, 1998**

(86) PCT No.: **PCT/EP98/02184**

§ 371 (c)(1), (2), (4) Date: **Nov. 29, 2000**

(87) PCT Pub. No.: **WO99/53667**

PCT Pub. Date: **Oct. 21, 1999**

(51) **Int. Cl.**
H03D 3/22 (2006.01)
H01J 11/00 (2006.01)

(52) **U.S. Cl.** **375/332**; 370/210

(58) **Field of Classification Search** 375/332, 375/200; 370/203, 200, 207, 480, 210
See application file for complete search history.

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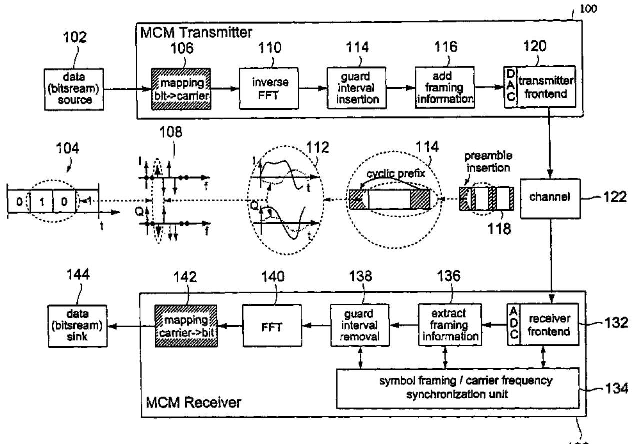
Primary Examiner—Stephen Chin
Assistant Examiner—Cicely Ware
(74) *Attorney, Agent, or Firm*—Roylance,Abrams,Berdo & Goodman, L.L.P.

(57) **ABSTRACT**

A method and an apparatus relating to a fine frequency synchronization compensating for a carrier frequency deviation from an oscillator frequency in a multi-carrier demodulation system of the type capable of carrying out a differential phase decoding of multi-carrier modulated signals, the signals comprising a plurality of symbols, each symbol being defined by phase differences between simultaneous carriers having different frequencies. A phase difference between phases of the same carrier in different symbols is determined. Thereafter, a frequency offset is determined by eliminating phase shift uncertainties related to the transmitted information from the phase difference making use of a M-PSK decision device. Finally, a feedback correction of the carrier frequency deviation is performed based on the determined frequency offset. Alternatively, an averaged frequency offset can be determined by averaging determined frequency offsets of a plurality of carriers. Then, the feedback correction of the frequency deviation is performed based on the averaged frequency offset.

7 Claims, 13 Drawing Sheets



Fraunhofer Ex 2002-74
Sirius XM v Fraunhofer, IPR2018-00689

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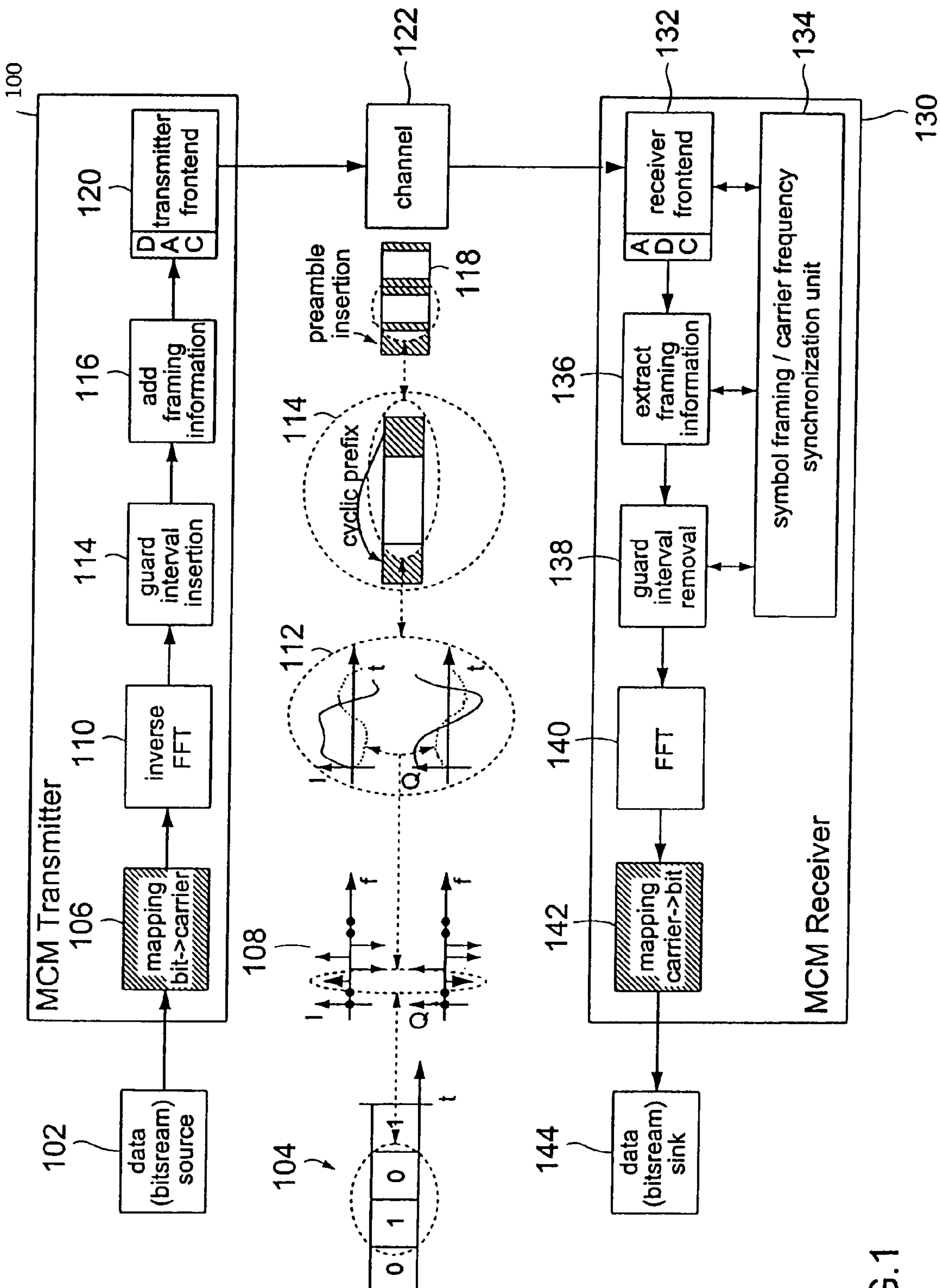

100
MCM Transmitter
102
data (bitsream) source
106
mapping bit->carrier
110
inverse FFT
114
guard interval insertion
116
add framing information
120
DAC
transmitter frontend
104
0
1
0
1
t
108
I
Q
f
112
t
114
cyclic prefix
preamble insertion
118
122
channel
130
MCM Receiver
132
receiver frontend
ADC
134
symbol framing / carrier frequency synchronization unit
136
extract framing information
138
guard interval removal
140
FFT
142
mapping carrier->bit
144
data (bitsream) sink

FIG.1

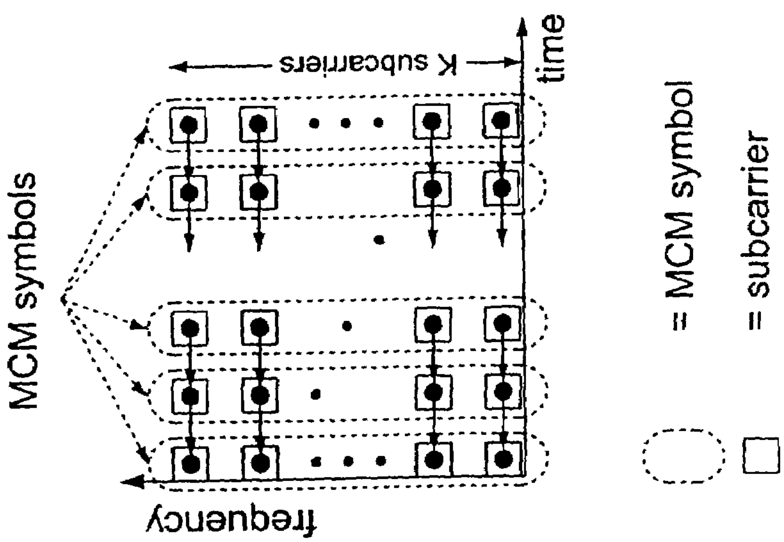

MCM symbols
K subcarriers
time
frequency
= MCM symbol
= subcarrier

FIG.2A

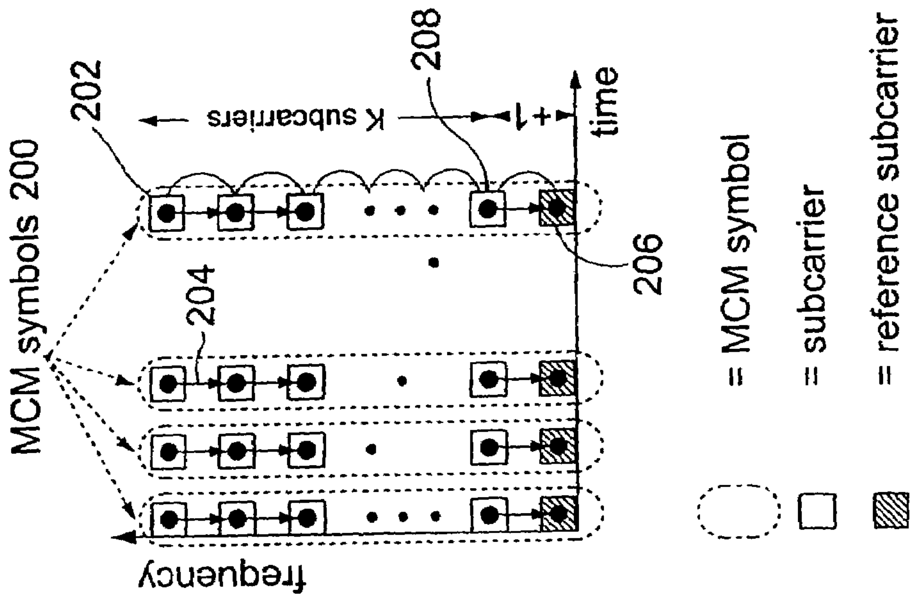

MCM symbols 200
202
208
K subcarriers
K+1
time
204
206
frequency
= MCM symbol
= subcarrier
= reference subcarrier

FIG.2B

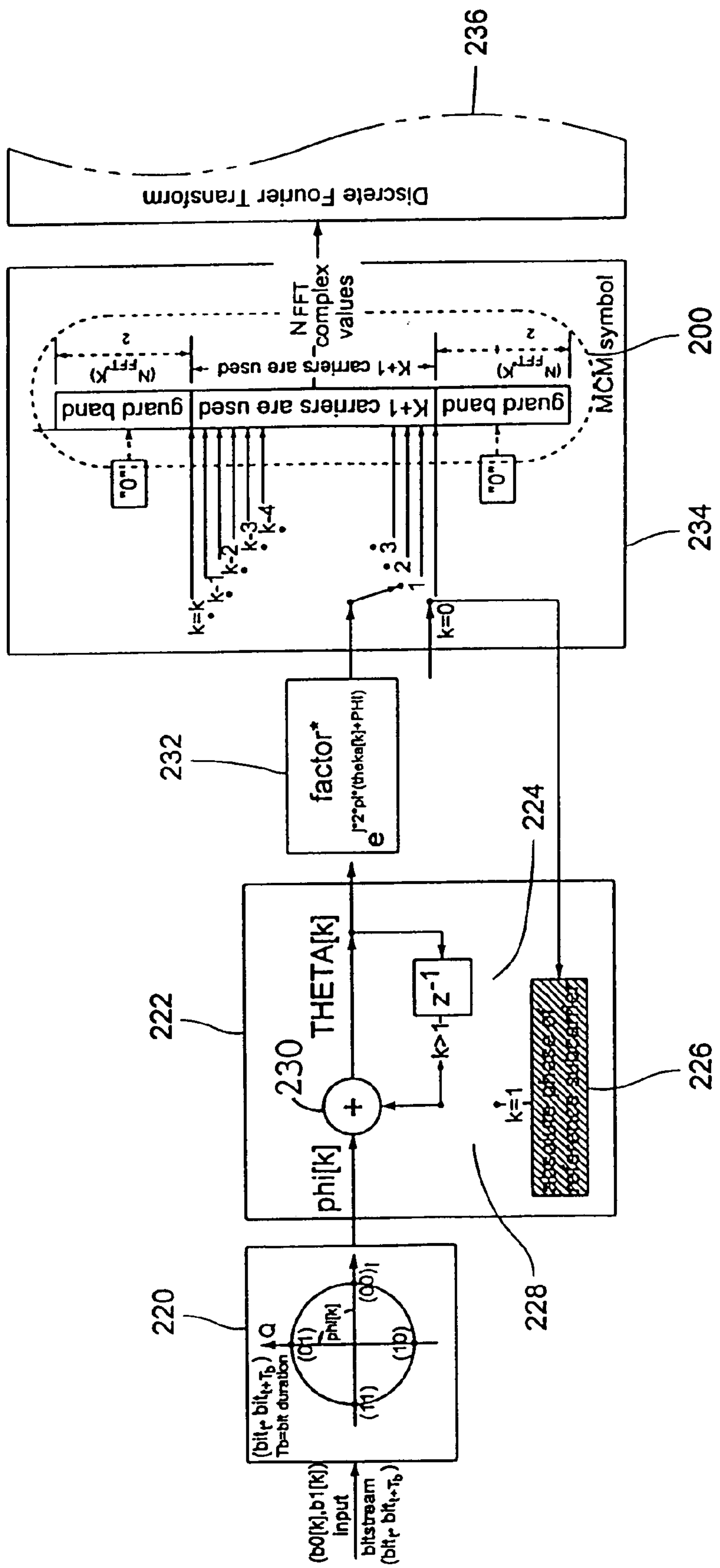

236
Discrete Fourier Transform
NFFT complex values
MCM symbol
200
guard band
K+1 carriers are used
guard band
(NFFT-K)/2
K+1 carriers are used
(NFFT-K)/2
"0"
"0"
k=k
k-1
k-2
k-3
k-4
3
2
1
k=0
234
232
factor* e^j*2*pi*(theta[k]+PHI)
224
222
THETA[k]
z^-1
k>1
230
+
phi[k]
k=1
226
228
220
Q
(01)
(00)
(11)
(10)
phi[k]
I
(bit_t, bit_t+Tb)
Tb=bit duration
(b0[k],b1[k])
input
bitstream
(bit_t, bit_t+Tb)

FIG.3

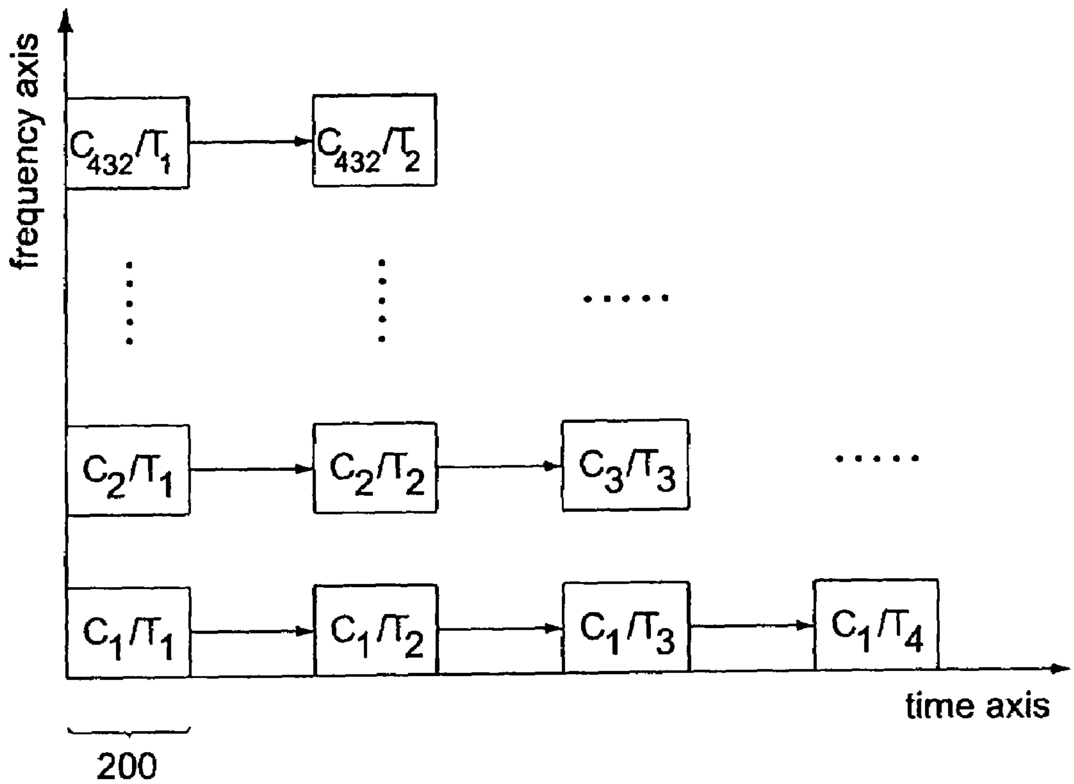

frequency axis
C432/T1
C432/T2
C2/T1
C2/T2
C3/T3
C1/T1
C1/T2
C1/T3
C1/T4
time axis
200

FIG.4

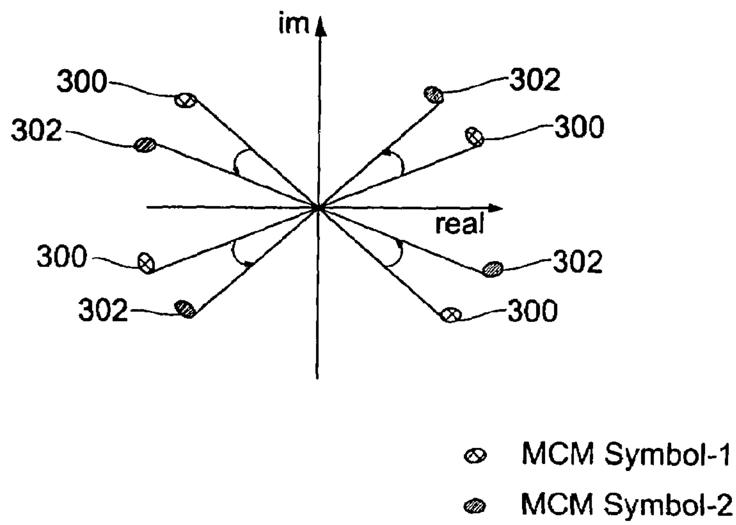

im
real
300
302
300
302
300
302
302
300
MCM Symbol-1
MCM Symbol-2

FIG.5

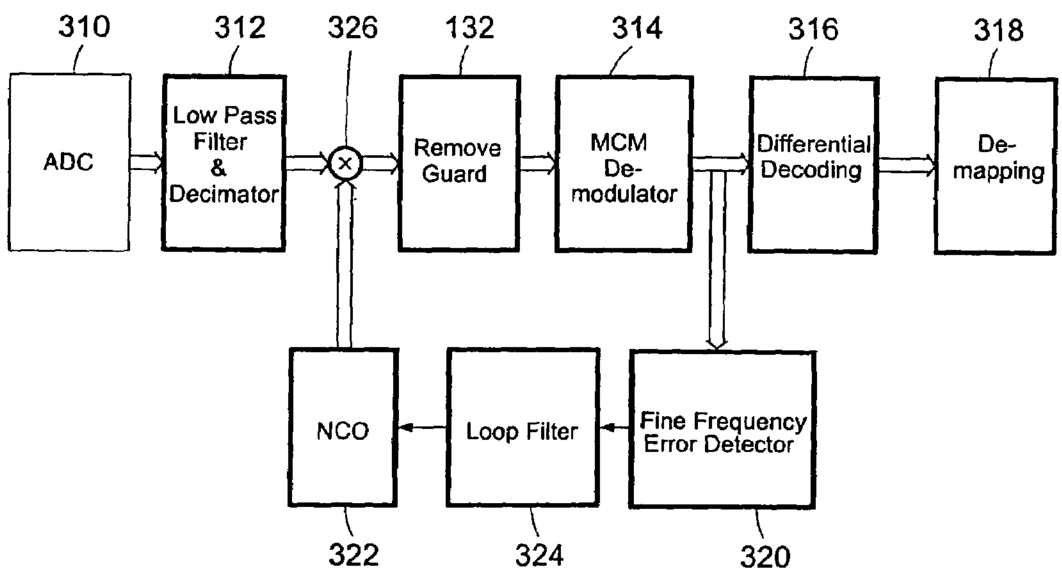

310
312
326
132
314
316
318
ADC
Low Pass Filter & Decimator
Remove Guard
MCM De-modulator
Differential Decoding
De-mapping
NCO
Loop Filter
Fine Frequency Error Detector
322
324
320

FIG.6

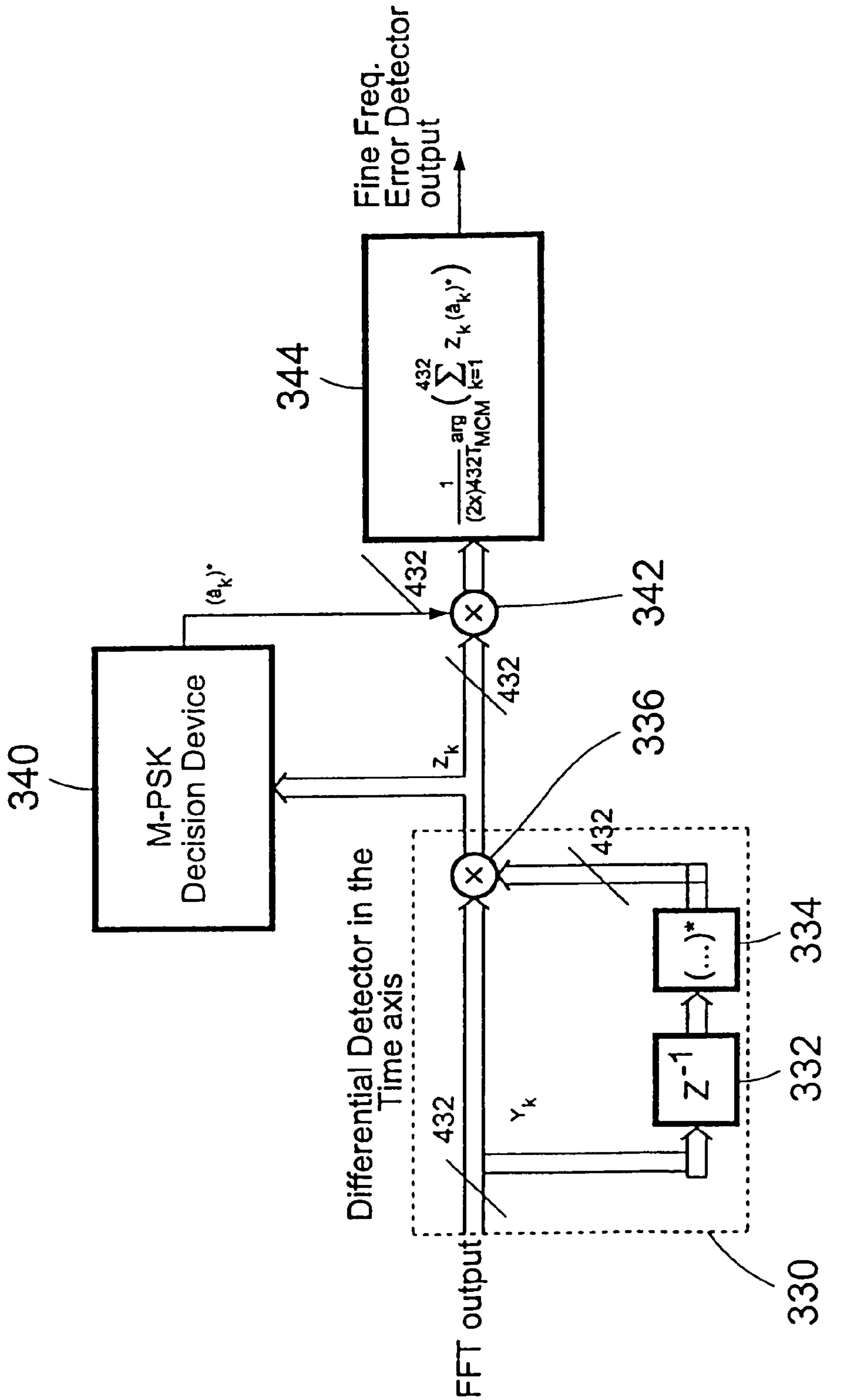
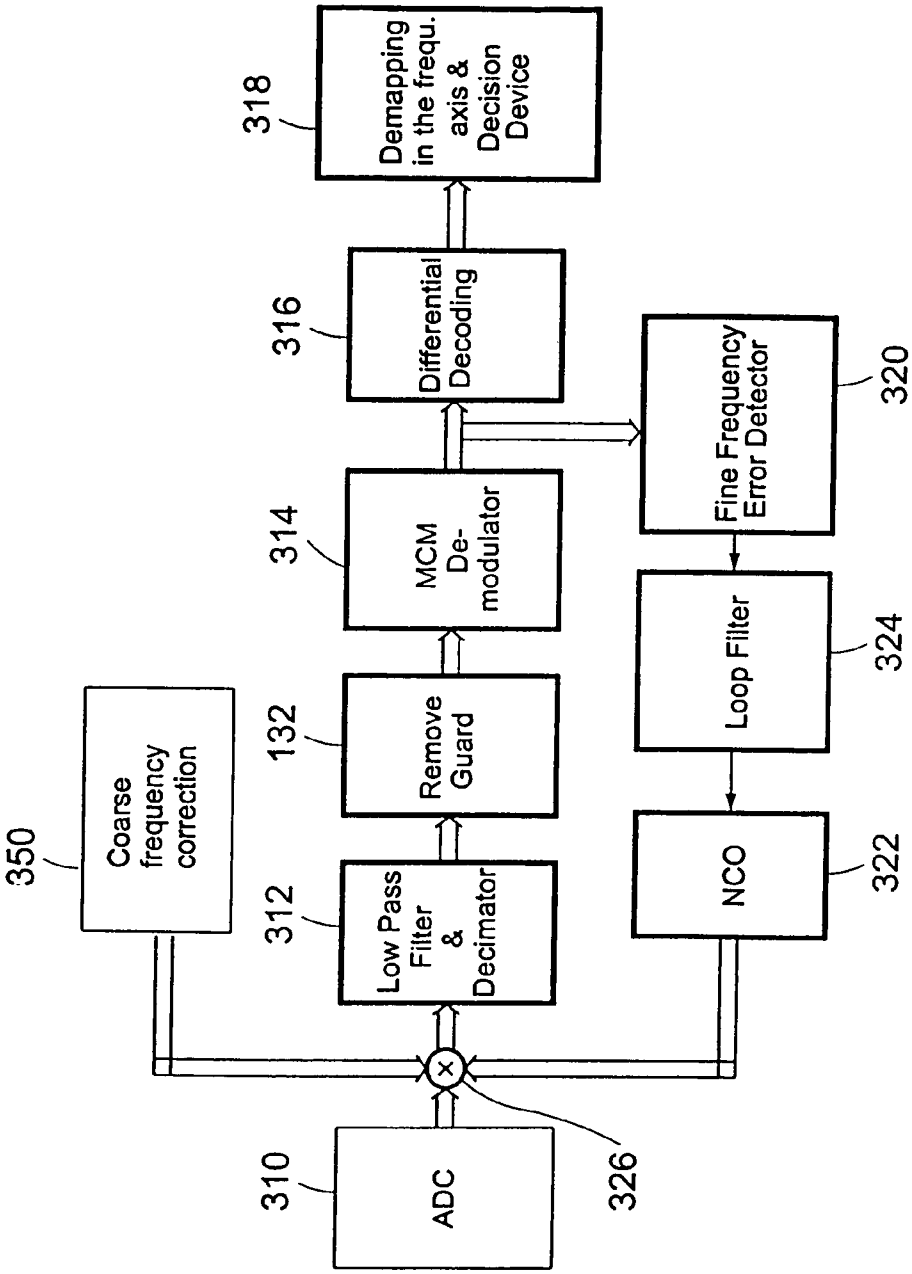

Fine Freq. Error Detector output
344
$\frac{1}{(2x)432T_{MCM}} \arg\left(\sum_{k=1}^{432} z_k (a_k)^*\right)$
432
$(a_k)^*$
342
340
M-PSK Decision Device
z_k
432
336
Differential Detector in the Time axis
432
Y_k
432
z^{-1}
$(\ldots)^*$
334
332
330
FFT output

FIG.7



318
Demapping in the frequ. axis & Decision Device
316
Differential Decoding
320
Fine Frequency Error Detector
314
MCM De-modulator
324
Loop Filter
132
Remove Guard
350
Coarse frequency correction
312
Low Pass Filter & Decimator
322
NCO
310
ADC
326

FIG.8

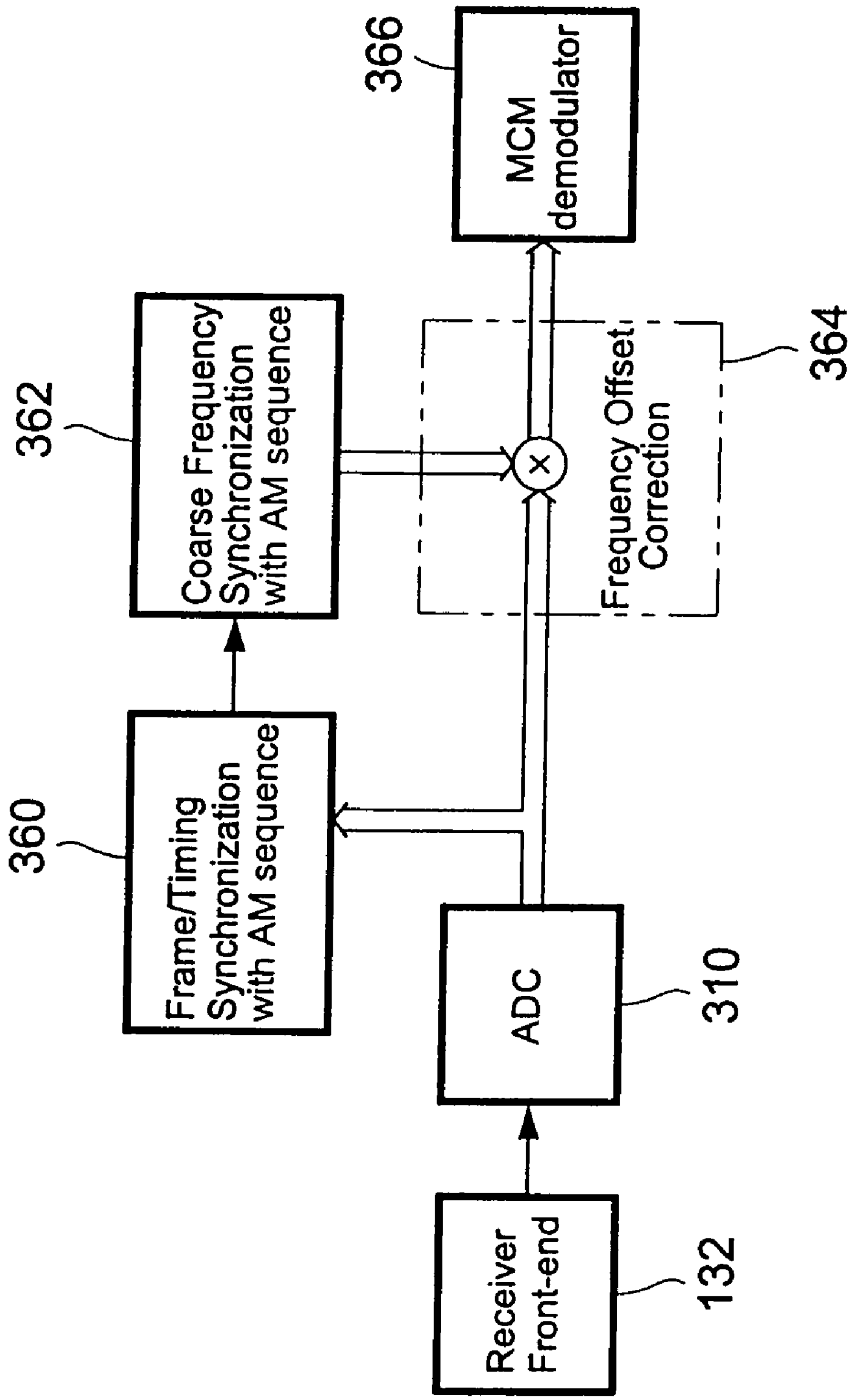

366
MCM demodulator
362
Coarse Frequency Synchronization with AM sequence
364
Frequency Offset Correction
360
Frame/Timing Synchronization with AM sequence
ADC
310
Receiver Front-end
132

FIG.9

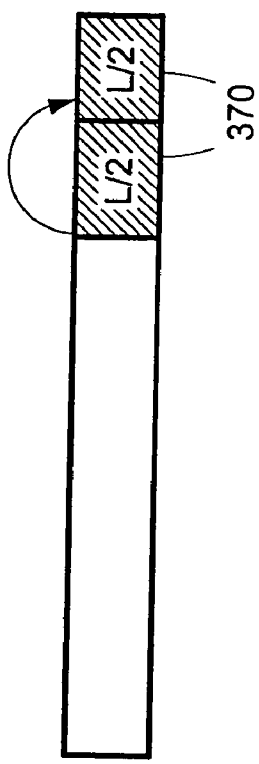

L/2
L/2
370

FIG.10

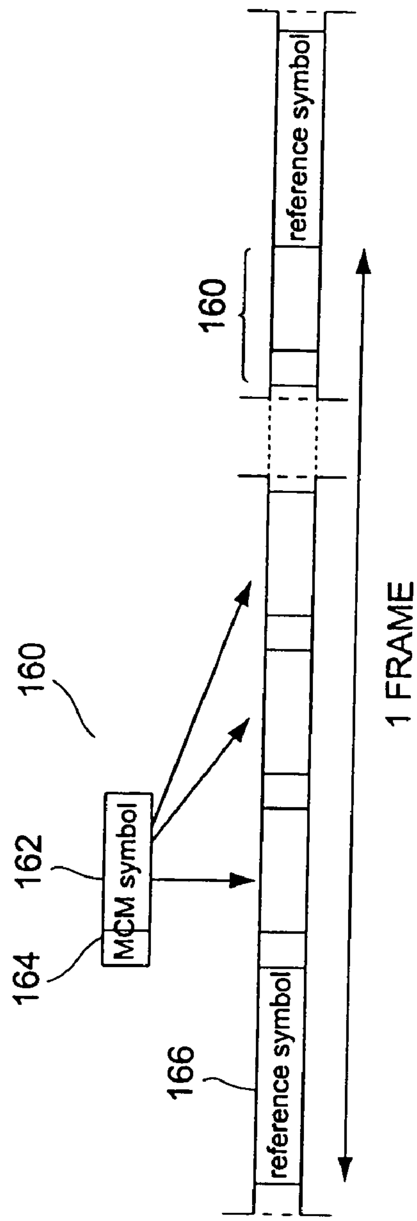

reference symbol
160
1 FRAME
160
MCM symbol
162
164
166
reference symbol

FIG.11

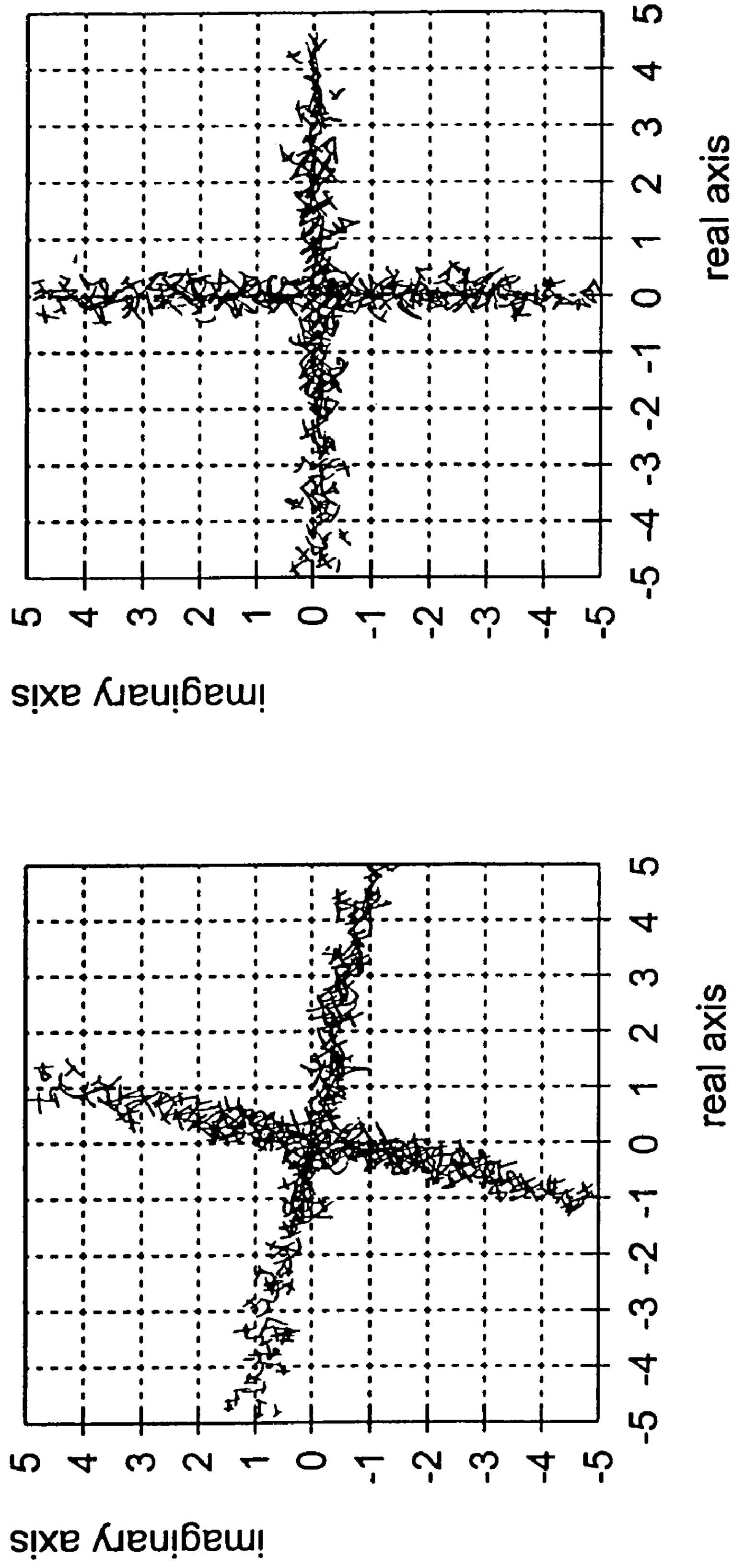


FIG.12

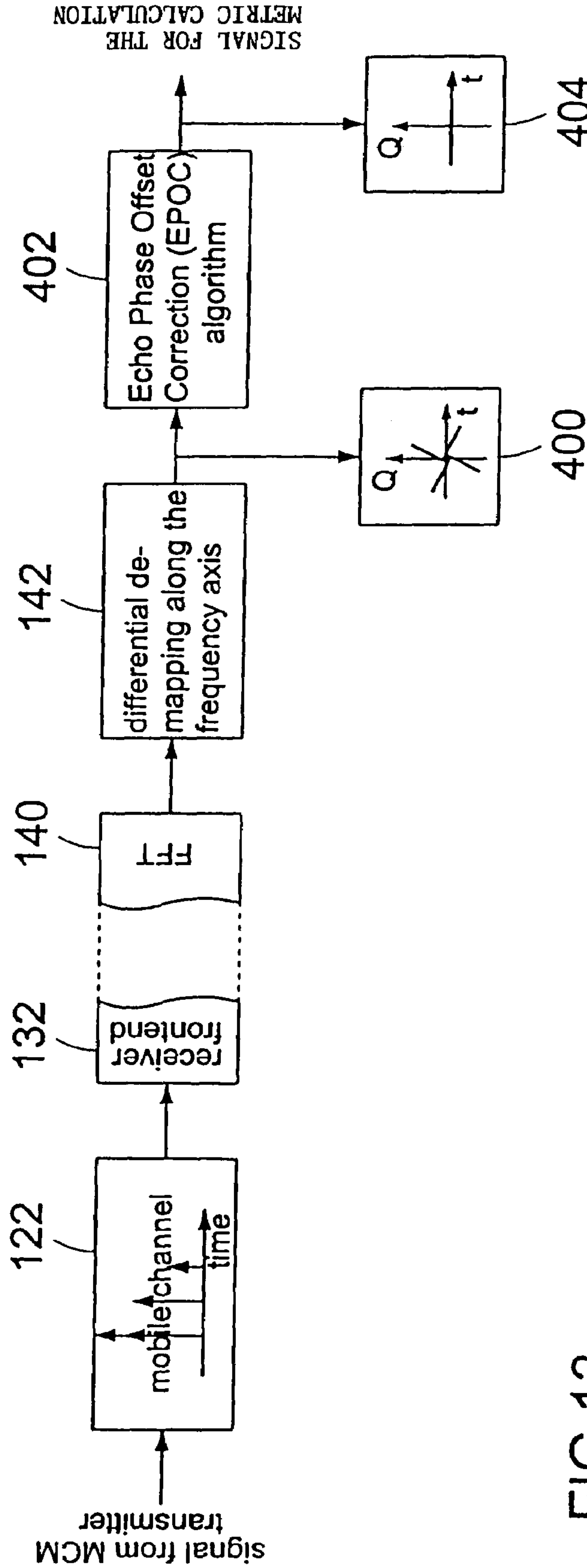

signal from MCM transmitter
122
mobile channel
time
132
receiver frontend
140
FFT
142
differential de-mapping along the frequency axis
400
Q
t
402
Echo Phase Offset Correction (EPOC) algorithm
404
Q
t
SIGNAL FOR THE METRIC CALCULATION

FIG.13

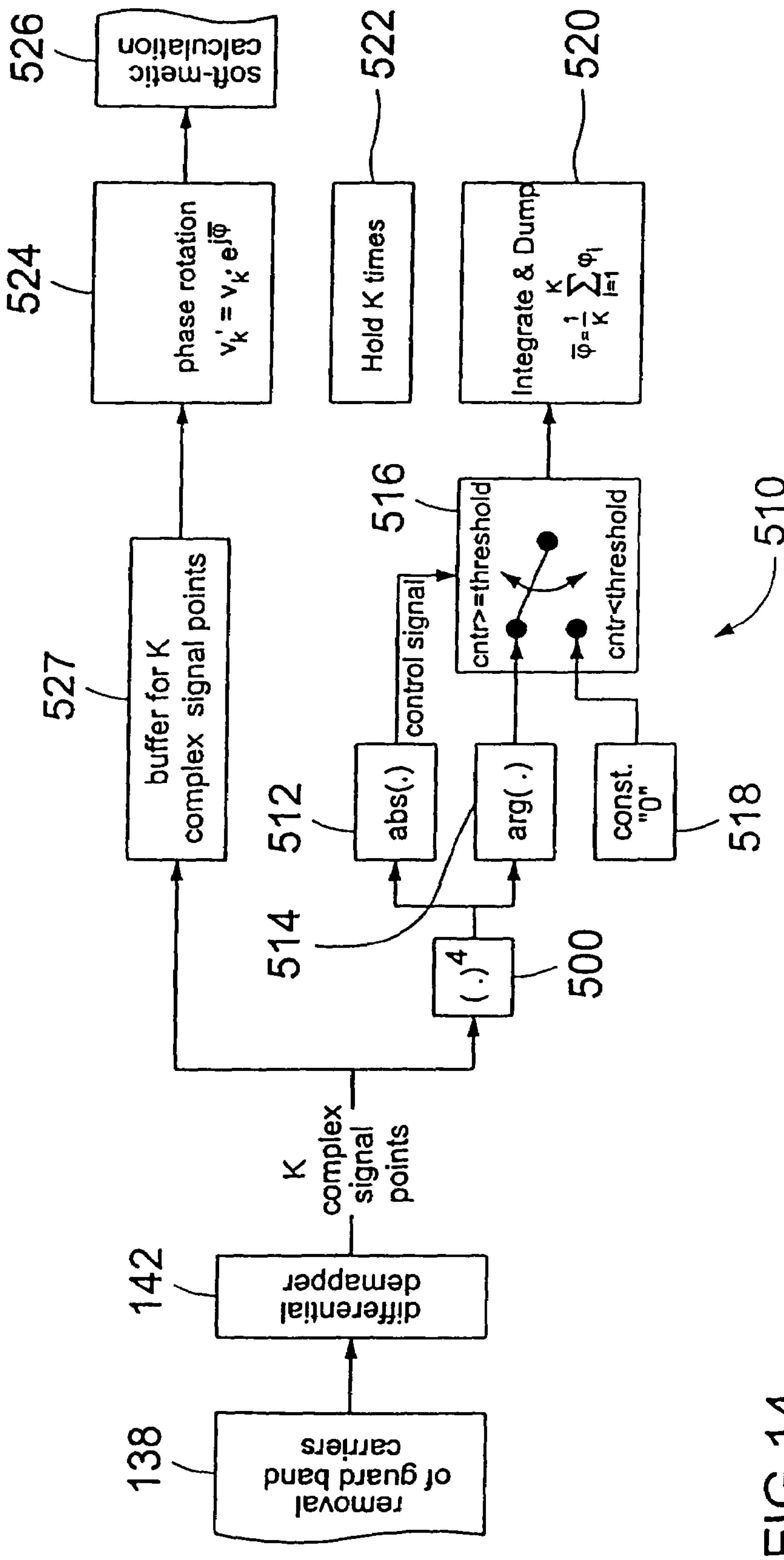

removal of guard band carriers
138
differential demapper
142
K complex signal points
buffer for K complex signal points
527
phase rotation
$v_k' = v_k' e^{j\overline{\varphi}}$
524
soft-metic calculation
526
$(.)^4$
500
514
abs(.)
512
arg(.)
const. "0"
518
control signal
cntr>=threshold
cntr<threshold
516
510
Integrate & Dump
$\overline{\varphi} = \frac{1}{K}\sum_{l=1}^{K}\varphi_l$
520
Hold K times
522

FIG.14

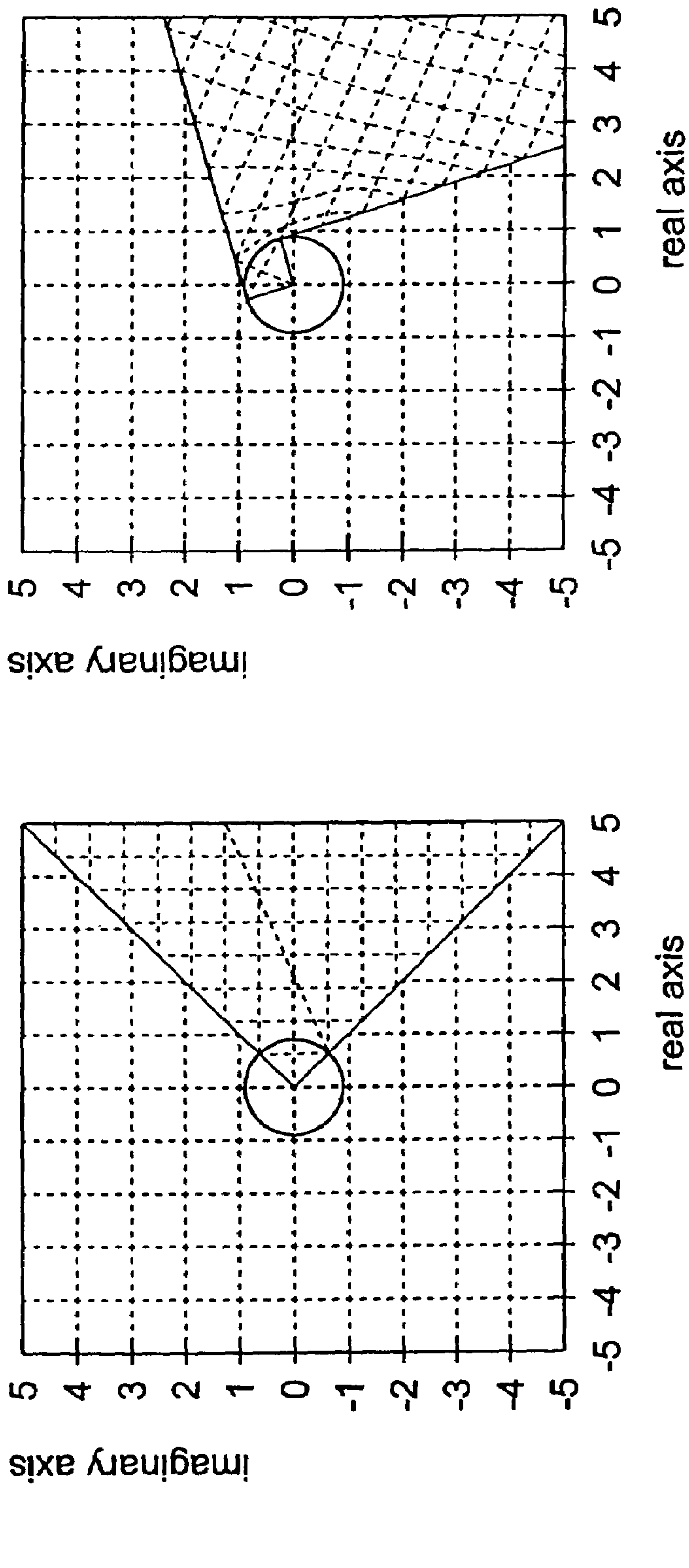


FIG.15

METHOD AND APPARATUS FOR FINE FREQUENCY SYNCHRONIZATION IN MULTI-CARRIER DEMODULATION SYSTEMS

This application is a 371 of PCT/EP98/02184 Apr. 14, 1998.

FIELD OF THE INVENTION

The present invention relates to methods and apparatus for performing a fine frequency synchronization in multi-carrier demodulation systems, and in particular to methods and apparatus for performing a fine frequency synchronization compensating for a carrier frequency deviation from an oscillator frequency in a multi-carrier demodulation system of the type capable of carrying out a differential phase decoding of multi-carrier modulated signals, wherein the signals comprise a plurality of symbols, each symbol being defined by phase differences between simultaneous carriers having different frequencies.

BACKGROUND OF THE INVENTION

In a multi carrier transmission system (MCM, OFDM), the effect of a carrier frequency offset is substantially more considerable than in a single carrier transmission system. MCM is more sensitive to phase noise and frequency offset which occurs as amplitude distortion and inter carrier interference (ICI). The inter carrier interference has the effect that the subcarriers are no longer orthogonal in relation to each other. Frequency offsets occur after power on or also later due to frequency deviation of the oscillators used for down-conversion into baseband. Typical accuracies for the frequency of a free running oscillator are about ±50 ppm of the carrier frequency. With a carrier frequency in the S-band of 2.34 Ghz, for example, there will be a maximum local oscillator (LO) frequency deviation of above 100 kHz (117.25 kHz). The above named effects result in high requirements on the algorithm used for frequency offset correction.

DESCRIPTION OF PRIOR ART

Most prior art algorithms for frequency synchronization divide frequency correction into two stages. In the first stage, a coarse synchronization is performed. In the second stage, a fine correction can be achieved. A frequently used algorithm for coarse synchronization of the carrier frequency uses a synchronization symbol which has a special spectral pattern in the frequency domain. Such a synchronization symbol is, for example, a CAZAC sequence (CAZAC=Constant Amplitude zero Autocorrelation). Through comparison, i.e. the correlation, of the power spectrum of the received signal with that of the transmitted signal, the frequency carrier offset can be coarsely estimated. These prior art algorithms all work in the frequency domain. Reference is made, for example, to Ferdinand Claßen, Heinrich Meyr, "Synchronization Algorithms for an OFDM System for Mobile Communication", ITG-Fachtagung 130, Codierung für Quelle, Kanal und Übertragung, pp. 105–113, Oct. 26–28, 1994; and Timothy M. Schmidl, Donald C. Cox, "Low-overhead, Low-Complexity [Burst] synchronization for OFDM", in Proceedings of the IEEE International conference on communication ICC 1996, pp. 1301–1306 (1996).

For the coarse synchronization of the carrier frequency, Paul H. Moose, "A Technique for orthogonal Frequency Division Multiplexing Frequency offset Correction", IEEE Transaction on communications, Vol. 42, No. 10, October 1994, suggest increasing the spacing between the subcarriers such that the subcarrier distance is greater than the maximum frequency difference between the received and transmitted carriers. The subcarrier distance is increased by reducing the number of sample values which are transformed by the Fast Fourier Transform. This corresponds to a reduction of the number of sampling values which are transformed by the Fast Fourier Transform.

WO 9205646 A relates to methods for the reception of orthogonal frequency division multiplexed signals comprising data which are preferably differentially coded in the direction of the time axis. Phase drift of the demodulated samples from one block to the next is used to indicate the degree of local oscillator frequency error. Phase drift is assessed by multiplying complex values by the complex conjugate of an earlier sample demodulated from the same OFDM carrier and using the resulting measure to steer the local oscillator frequency via a frequency locked loop.

SUMMARY OF THE INVENTION

It is an object of the present invention to provide methods and apparatus for performing a fine frequency synchronization which allow a fine frequency synchronization compensating for a carrier frequency deviation from an oscillator frequency in a MCM transmission system which makes use of MCM signals in which information is differential phase encoded between simultaneous sub-carriers having different frequencies.

In accordance with a first aspect, the present invention provides a method of performing a fine frequency synchronization compensating for a carrier frequency deviation from an oscillator frequency in a multi-carrier demodulation system of the type capable of carrying out a differential phase decoding of multi-carrier modulated signals, the signals comprising a plurality of symbols, each symbol being defined by phase differences between simultaneous carriers having different frequencies, the method comprising the steps of:

determining a phase difference between phases of the same carrier in different symbols;

determining a frequency offset by eliminating phase shift uncertainties related to the transmitted information from the phase difference making use of a M-PSK decision device; and

performing a feedback correction of the carrier frequency deviation based on the determined frequency offset.

In accordance with a second aspect, the present invention provides a method of performing a fine frequency synchronization compensating for a carrier frequency deviation from an oscillator frequency in a multi-carrier demodulation system of the type capable of carrying out a differential phase decoding of multi-carrier modulated signals, the signals comprising a plurality of symbols, each symbol being defined by phase differences between simultaneous carriers having different frequencies, the method comprising the steps of:

determining respective phases of the same carrier in different symbols;

eliminating phase shift uncertainties related to the transmitted information from the phases to determine respective phase deviations making use of a M-PSK decision device;

determining a frequency offset by determining a phase difference between the phase deviations; and

performing a feedback correction of said carrier frequency deviation based on the determined frequency offset.

In accordance with a third aspect, the present invention provides an apparatus for performing a fine frequency synchronization compensating for a carrier frequency deviation from an oscillator frequency, for a multi-carrier demodulation system of the type capable of carrying out a differential phase decoding of multi-carrier modulated signals, the signals comprising a plurality of symbols, each symbol being defined by phase differences between simultaneous carriers having different frequencies, the apparatus comprising:

means for determining a phase difference between phases of the same carrier in different symbols;

M-PSK decision device for determining a frequency offset by eliminating phase shift uncertainties related to the transmitted information from the phase difference; and

means for performing a feedback correction of the frequency deviation based on the determined frequency offset.

In accordance with a fourth aspect, the present invention provides an apparatus for performing a fine frequency synchronization compensating for a carrier frequency deviation from an oscillator frequency, for a multi-carrier demodulation system of the type capable of carrying out a differential phase decoding of multi-carrier modulated signals, said signals comprising a plurality of symbols, each symbol being defined by phase differences between simultaneous carriers having different frequencies, the apparatus comprising:

means for determining respective phases of the same carrier in different symbols;

M-PSK decision device for eliminating phase shift uncertainties related to the transmitted information from the phases to determine respective phase deviations;

means for determining a frequency offset by determining a phase difference between the phase deviations; and

means for performing a feedback correction of the frequency deviation based on the determined frequency offset.

The present invention relates to methods and apparatus for performing a fine frequency synchronization compensating for a carrier frequency deviation from an oscillator frequency. This fine frequency synchronization is preferably performed after completion of a coarse frequency synchronization, such that the frequency offsets after the coarse frequency synchronization are smaller than half the sub-carrier distance in the MCM signal. Since the frequency offsets which are to be corrected by the inventive fine frequency synchronization methods and apparatus, a correction of the frequency offsets by using a phase rotation with differential decoding and de-mapping in the time axis can be used. The frequency offsets are detected by determining the frequency differences between time contiguous sub-carrier symbols along the time axis. The frequency error is calculated by measuring the rotation of the I-Q Cartesian coordinates of each sub-carrier and, in preferred embodiments, averaging them over all n sub-carriers of a MCM symbol.

Firstly, the phase ambiguity or uncertainty is eliminated by using a M-PSK decision device and correlating the output of the decision device with the input signal for a respective sub-carrier symbol. Thus, the phase offset for a sub-carrier symbol is determined and can be used for restructuring the frequency error in form of a feed-backward structure. Alternatively, the phase offsets of the sub-carrier symbols of one MCM symbol can be averaged over all of the active carriers of a MCM symbol, wherein the averaged phase offset is used to restructure the frequency error.

In accordance with the present invention, the determination of the frequency offset is performed in the frequency domain. The feedback correction in accordance with the inventive fine frequency synchronization is performed in the time domain. To this end, a differential decoder in the time domain is provided in order to detect frequency offsets of sub-carriers on the basis of the phases of timely successive sub-carrier symbols of different MCM symbols.

BRIEF DESCRIPTION OF THE DRAWINGS

In the following, preferred embodiments of the present invention will be explained in detail on the basis of the drawings enclosed, in which:

FIG. **1** shows a schematic overview of a MCM transmission system to which the present application can be applied;

FIGS. **2**A and **2**B show schematic views representing a scheme for differential mapping in the time axis and a scheme for differential mapping in the frequency axis;

FIG. **3** shows a functional block diagram for performing a differential mapping in the frequency axis;

FIG. **4** shows a representation of time variation of all sub-carriers in MCM symbols;

FIG. **5** shows a QPSK-constellation for each sub-carrier with a frequency offset;

FIG. **6** shows a general block diagram illustrating the position of the inventive fine frequency synchronization device in a MCM receiver;

FIG. **7** shows a block diagram of the fine frequency error detector shown in FIG. **6**;

FIG. **8** shows a block diagram of a MCM receiver comprising a coarse frequency synchronization unit and a fine frequency synchronization unit;

FIG. **9** shows a block diagram of a unit for performing a coarse frequency synchronization;

FIG. **10** shows a schematic view of a reference symbol used for performing a coarse frequency synchronization;

FIG. **11** shows a schematic view of a typical MCM signal having a frame structure;

FIG. **12** shows scatter diagrams of the output of an (differential de-mapper of an MCM receiver for illustrating the effect of an echo phase offset correction;

FIG. **13** shows a schematic block diagram for illustrating the position and the functionality of an echo phase offset correction unit;

FIG. **14** shows a schematic block diagram of a preferred form of an echo phase offset correction device; and

FIG. **15** shows schematic views for illustrating a projection performed by another echo phase offset correction algorithm.

DETAILED DESCRIPTION OF THE EMBODIMENTS

Before discussing the present invention in detail, the mode of operation of a MCM transmission system is described referring to FIG. **1**.

Referring to FIG. **1**, at **100** a MCM transmitter is shown that substantially corresponds to a prior art MCM transmitter. A description of such a MCM transmitter can be found, for example, in William Y. Zou, Yiyan Wu, "COFDM: AN OVERVIEW", IEEE Transactions on Broadcasting, vol. 41, No. 1, March 1995.

A data source **102** provides a serial bitstream **104** to the MCM transmitter. The incoming serial bitstream **104** is applied to a bit-carrier mapper **106** which produces a sequence of spectra **108** from the incoming serial bitstream **104**. An inverse fast Fourier transform (IFFT) **110** is performed on the sequence of spectra **108** in order to produce a MCM time domain signal **112**. The MCM time domain signal forms the useful MCM symbol of the MCM time signal. To avoid intersymbol interference (ISI) caused by multipath distortion, a unit **114** is provided for inserting a guard interval of fixed length between adjacent MCM symbols in time. In accordance with a preferred embodiment of the present invention, the last part of the useful MCM symbol is used as the guard interval by placing same in front of the useful symbol. The resulting MCM symbol is shown at **115** in FIG. **1** and corresponds to a MCM symbol **160** depicted in FIG. **11**.

FIG. **11** shows the construction of a typical MCM signal having a frame structure. One frame of the MCM time signal is composed of a plurality of MCM symbols **160**. Each MCM symbol **160** is formed by an useful symbol **162** and a guard interval **164** associated therewith. AS shown in FIG. **11**, each frame comprises one reference symbol **166**. The present invention can advantageously be used with such a MCM signal, however, such a signal structure being not necessary for performing the present invention as long as the transmitted signal comprises a useful portion and at least one reference symbol.

In order to obtain the final frame structure shown in FIG. **11**, a unit **116** for adding a reference symbol for each predetermined number of MCM symbols is provided.

In accordance with the present invention, the reference symbol is an amplitude modulated bit sequence. Thus, an amplitude modulation of a bit sequence is performed such that the envelope of the amplitude modulated bit sequence defines a reference pattern of the reference symbol. This reference pattern defined by the envelope of the amplitude modulated bit sequence has to be detected when receiving the MCM signal at a MCM receiver. In a preferred embodiment of the present invention, a pseudo random bit sequence having good autocorrelation properties is used as the bit sequence that is amplitude modulated.

The choice of length and repetition rate of the reference symbol depends on the properties of the channel through which the MCM signal is transmitted, e.g. the coherence time of the channel. In addition, the repetition rate and the length of the reference symbol, in other words the number of useful symbols in each frame, depends on the receiver requirements concerning mean time for initial synchronization and mean time for resynchronization after synchronization loss due to a channel fade.

The resulting MCM signal having the structure shown at **118** in FIG. **1** is applied to the transmitter front end **120**. Roughly speaking, at the transmitter front end **120**, a digital/analog conversion and an up-converting of the MCM signal is performed. Thereafter, the MCM signal is transmitted through a channel **122**.

Following, the mode of operation of a MCM receiver **130** is shortly described referring to FIG. **1**. The MCM signal is received at the receiver front end **132**. In the receiver front end **132**, the MCM signal is down-converted and, furthermore, an analog/digital conversion of the down-converted signal is performed.

The down-converted MCM signal is provided to a symbol frame/carrier frequency synchronization unit **134**.

A first object of the symbol frame/carrier frequency synchronization unit **134** is to perform a frame synchronization on the basis of the amplitude-modulated reference symbol. This frame synchronization is performed on the basis of a correlation between the amplitude-demodulated reference symbol and a predetermined reference pattern stored in the MCM receiver.

A second object of the symbol frame/carrier frequency synchronization unit is to perform a coarse frequency synchronization of the MCM signal. To this end, the symbol frame/carrier frequency synchronization unit **134** serves as a coarse frequency synchronization unit for determining a coarse frequency offset of the carrier frequence caused, for example, by a difference of the frequencies between the local oscillator of the transmitter and the local oscillator of the receiver. The determined frequency is used in order to perform a coarse frequency correction. The mode of operation of the coarse frequency synchronization unit is described in detail referring to FIGS. **9** and **10** hereinafter.

As described above, the frame synchronization unit **134** determines the location of the reference symbol in the MCM symbol. Based on the determination of the frame synchronization unit **134**, a reference symbol extracting unit **136** extracts the framing information, i.e. the reference symbol, from the MCM symbol coming from the receiver front end **132**. After the extraction of the reference symbol, the MCM signal is applied to a guard interval removal unit **138**. The result of the signal processing performed hereherto in the MCM receiver are the useful MCM symbols.

The useful MCM symbols output from the guard interval removal unit **138** are provided to a fast Fourier transform unit **140** in order to provide a sequence of spectra from the useful symbols. Thereafter, the sequence of spectra is provided to a carrier-bit mapper **142** in which the serial bitstream is recovered. This serial bitstream is provided to a data sink **144**.

Next, referring to FIGS. **2**A and **2**B, two modes for differential mapping are described. In FIG. **2**A, a first method of differential mapping along the time axis is shown. AS can be seen from FIG. **2**A, a MCM symbol consists of K subcarriers. The sub-carriers comprise different frequencies and are, in a preferred embodiment, equally spaced in the frequency axis direction. When using differential mapping along the time axis, one or more bits are encoded into phase and/or amplitude shifts between two sub-carriers of the same center frequency in adjacent MCM symbols. The arrows depicted between the sub-carrier symbols correspond to information encoded in amplitude and/or phase shifts between two sub-carrier symbols.

A second method of differential mapping is shown in FIG. **2**B. The present invention is adapted for MCM transmission system using the mapping scheme shown in FIG. **2**B. This mapping scheme is based on a differential mapping inside one MCM symbol along the frequency axis. A number of MCM symbols **200** are shown in FIG. **2**B. Each MCM symbol **200** comprises a number of sub-carrier symbols **202**. The arrows **204** in FIG. **2**B illustrate information encoded between two sub-carrier symbols **202**. As can be seen from the arrows **204**, this mapping scheme is based on a differential mapping within one MCM symbol along the frequency axis direction.

In the embodiment shown in FIG. **2**B, the first sub-carrier (k=0) in an MCM symbol **200** is used as a reference sub-carrier **206** (shaded) such that information is encoded between the reference sub-carrier and the first active carrier **208**. The other information of a MCM symbol **200** is encoded between active carriers, respectively.

Thus, for every MCM symbol an absolute phase reference exists. In accordance with FIG. **2**B, this absolute phase reference is supplied by a reference symbol inserted into every MCM symbol (k=0). The reference symbol can either have a constant phase for all MCM symbols or a phase that varies from MCM symbol to MCM symbol. A varying phase can be obtained by replicating the phase from the last subcarrier of the MCM symbol preceding in time.

In FIG. **3** a preferred embodiment of a device for performing a differential mapping along the frequency axis is shown. Referring to FIG. **3**, assembly of MCM symbols in the frequency domain using differential mapping along the frequency axis according to the present invention is described.

FIG. **3** shows the assembly of one MCM symbol with the following parameters:

NFFT designates the number of complex coefficients of the discrete Fourier transform, number of subcarriers respectively.

K designates the number of active carriers. The reference carrier is not included in the count for K.

According to FIG. **3**, a quadrature phase shift keying (QPSK) is used for mapping the bitstream onto the complex symbols. However, other M-ary mapping schemes (MPSK) like 2-PSK, 8-PSK, 16-QAM, 16-APSK, 64-APSK etc. are possible.

Furthermore, for ease of filtering and minimization of aliasing effects some subcarriers are not used for encoding information in the device shown in FIG. **3**. These subcarriers, which are set to zero, constitute the so-called guard bands on the upper and lower edges of the MCM signal spectrum. At the input of the mapping device shown in FIG. **3**, complex signal pairs b**0**[k], b**1**[k] of an input bitstream are received. K complex signal pairs are assembled in order to form one MCM symbol. The signal pairs are encoded into the K differential phase shifts phi[k] needed for assembly of one MCM symbol. In this embodiment, mapping from Bits to the 0, 90, 180 and 270 degrees phase shifts is performed using Gray Mapping in a quadrature phase shift keying device **220**.

Gray mapping is used to prevent that differential detection phase errors smaller than 135 degrees cause double bit errors at the receiver.

Differential phase encoding of the K phases is performed in a differential phase encoder **222**. At this stage of processing, the K phases phi[k] generated by the QPSK Gray mapper are differentially encoded. In principal, a feedback loop **224** calculates a cumulative sum over all K phases. As starting point for the first computation (k=0) the phase of the reference carrier **226** is used. A switch **228** is provided in order to provide either the absolute phase of the reference subcarrier **226** or the phase information encoded onto the preceding (i.e. z^{-1}, where z^{-1} denotes the unit delay operator) subcarrier to a summing point **230**. At the output of the differential phase encoder **222**, the phase information theta [k] with which the respective subcarriers are to be encoded is provided. In preferred embodiments of the present invention, the subcarriers of a MCM symbol are equally spaced in the frequency axis direction.

The output of the differential phase encoder **222** is connected to a unit **232** for generating complex subcarrier symbols using the phase information theta[k]. To this end, the K differentially encoded phases are converted to complex symbols by multiplication with

$$\text{factor}*e^{j*[2*pi*(theta[k]+PHI)]} \qquad \text{(Eq.1)}$$

wherein factor designates a scale factor and PHI designates an additional angle. The scale factor and the additional angle PHI are optional. By choosing PHI=45° a rotated DQPSK signal constellation can be obtained.

Finally, assembly of a MCM symbol is effected in an assembling unit **234**. One MCM symbol comprising N_{FFT} subcarriers is assembled from N_{FFT}-K−1 guard band symbols which are "zero", one reference subcarrier symbol and K DQPSK subcarrier symbols. Thus, the assembled MCM symbol **200** is composed of K complex values containing the encoded information, two guard bands at both sides of the N_{FFT} complex values and a reference subcarrier symbol.

The MCM symbol has been assembled in the frequency domain. For transformation into the time domain an inverse discrete Fourier transform (IDFT) of the output of the assembling unit **234** is performed by a transformator **236**. In preferred embodiments of the present invention, the transformator **236** is adapted to perform a fast Fourier transform (FFT).

Further processing of the MCM signal in the transmitter as well as in the receiver is as described above referring to FIG. **1**.

At the receiver a de-mapping device **142** (FIG. **1**) is needed to reverse the operations of the mapping device described above referring to FIG. **3**. The implementation of the de-mapping device is straightforward and, therefore, need not be described herein in detail.

The differential mapping along the frequency axis direction is suitable for multi-carrier (OFCM) digital broadcasting over rapidly changing multi path channels. In accordance with this mapping scheme, there is no need for a channel stationarity exceeding one multi-carrier symbol. However, differential mapping into frequency axis direction may create a new problem. In multi path environments, path echoes succeeding or preceding the main path can lead to systematic phase offsets between sub-carriers in the same MCM symbol. Thus, it will be preferred to provide a correction unit in order to eliminate such phase offsets. Because the channel induced phase offsets between differential demodulated symbols are systematic errors, they can be corrected by an algorithm. In principle, such an algorithm must calculate the echo induced phase offset from the signal space constellation following the differential demodulation and subsequently correct this phase offset.

Examples for such echo phase correction algorithms are described at the end of this specification referring to FIGS. **12** to **15**.

Next, the fine frequency synchronization in accordance with the present invention will be described referring to FIGS. **4** to **8**. As mentioned above, the fine frequency synchronization in accordance with the present invention is performed after completion of the coarse frequency synchronization. Preferred embodiments of the coarse frequency synchronization which can be performed by the symbol frame/carrier frequency synchronization unit **134** are described hereinafter referring to FIGS. **9** and **10** after having described the fine frequency synchronization in accordance with the present invention.

With the fine frequency synchronization in accordance with the present invention frequency offsets which are smaller than half the sub-carrier distance can be corrected. Since the frequency offsets are low and equal for all subcarriers the problem of fine frequency synchronization is reduced to sub-carrier level. FIG. **4** is a schematical view of MCM symbols **200** in the time-frequency plane. Each MCM symbol **200** consists of 432 sub-carrier symbols C_1 to C_{432}. The MCM symbols are arranged along the time axis, the first MCM symbol **200** shown in FIG. **4** having associated therewith a time T_1, the next MCM symbol having associated therewith a time T_2 and so on. In accordance with a preferred embodiment of the present invention, the fine frequency synchronization is based on a phase rotation which is derived from the same sub-carrier of two MCM symbols which are adjacent in the time axis direction, for example C_1/T_1 and C_1/T_2.

In the following, the present invention is described referring to QPSK mapping (QPSK=Quadrature Phase Shift Keying). However, it is obvious that the present invention can be applied to any MPSK mapping, wherein M designates the number of phase states used for encoding, for example 2, 4, 8, 16 . . . .

FIG. **5** represents a complex coordinate system showing a QPSK constellation for each sub-carrier with frequency offset. The four possible phase positions of a first MCM symbol, MCM-symbol-**1** are shown at **300**. Changing from the sub-carrier (sub-carrier n) of this MCM symbol to the same sub-carrier of the next MCM symbol, MCM-symbol-**2**, the position in the QPSK constellation will be unchanged in case there is no frequency offset. If a frequency offset is present, which is smaller than half the distance between sub-carriers, as mentioned above, this frequency offset causes a phase rotation of the QPSK constellation of MCM-symbol-**2** compared with MCM-symbol-**1**. The new QPSK constellation, that is the four possible phase positions for the subject sub-carrier of MCM-symbol-**2** are shown at **302** in FIG. **5**. This phase rotation θ can be derived from the following equation:

$$C_n(kT_{MCM})=e^{j2\pi f_{offset}T_{MCM}}C_n((k-1)T_{MCM})$$

$$\theta=2\pi f_{offset}T_{MCM} \qquad \text{(Eq.2)}$$

C_n designates the QPSK constellation of a sub-carrier n in a MCM symbol. n is an index running from 1 to the number of active sub-carriers in the MCM symbol. Information regarding the frequency offset is contained in the term $e^{j2\pi f_{offset}T_{MCM}}$ of equation 2. This frequency offset is identical for all sub-carriers. Therefore, the phase rotation θ is identical for all sub-carriers as well. Thus, averaging overall sub-carrier of a MCM symbol can be performed.

FIG. **6** shows a block diagram of a MCM receiver in which the present invention is implemented. An analog/digital converter **310** is provided in order to perform an analog/digital conversion of a down-converted signal received at the receiver front end **132** (FIG. **1**). The output of the analog/digital converter **310** is applied to a low path filter and decimator unit **312**. The low path filter is an impulse forming filter which is identical to an impulse forming filter in the MCM transmitter. In the decimator, the signal is sampled at the MCM symbol frequency. As described above referring to FIG. **1**, guard intervals in the MCM signal are removed by a guard interval removal unit **132**. Guard intervals are inserted between two MCM symbols in the MCM transmitter in order to avoid intersymbol interference caused by channel memory.

The output of the guard interval removal unit **132** is applied to a MCM demodulator **314** which corresponds to the fast Fourier transformator **140** shown in FIG. **1**. Following the MCM demodulator **314** a differential decoding unit **316** and a de-mapping unit **318** are provided. In the differential decoding unit **316**, phase information is recovered using differential decoding. In the demapping unit **318**, demapping along the frequency axis direction is performed in order to reconstruct a binary signal from the complex signal input into the demapping unit **318**.

The output of the MCM demodulator **314** is also applied to fine frequency error detector **320**. The fine frequency error detector **320** produces a frequency error signal from the output of the MCM demodulator. In the depicted embodiment, the output of the fine frequency error detector **320** is applied to a numerical controlled oscillator **322** via a loop filter **324**. The loop filter **324** is a low pass filter for filtering superimposed interference portions of a higher frequency from the slowly varying error signal. The numerical controlled oscillator **322** produces a carrier signal on the basis of the filtered error signal. The carrier signal produced by the numerical controlled oscillator **322** is used for a frequency correction which is performed by making use of a complex multiplier **326**. The inputs to the complex multiplier **326** are the output of the low pass filter and decimator unit **312** and the output of the numerical controlled oscillator **322**.

A description of a preferred embodiment of the fine frequency error detector **320** is given hereinafter referring to FIG. **7**.

The fine frequency error detector **320** comprises a differential detector in the time axis **330**. The output of the MCM demodulator **314**, i.e. the FFT output (FFT=Fast Fourier Transform) is applied to the input of the differential detector **330** which performs a differential detection in the time axis in order to derive information on a frequency offset from the same sub-carrier of two subsequently arriving MCM symbols. In the embodiment shown in FIG. **7**, the number of active sub-carriers is **432**. Thus, the differential detector **330** performs a correlation between the first and the 433rd sample. The first sample is associated with MCM-symbol-1 (FIG. **5**), whereas the 433rd sample is associated with MCM-symbol-2 (FIG. **5**). However, both of these samples are associated with the same sub-carrier.

To this end, the input signal Y_k is applied to a z^{-1}-block **332** and thereafter to a unit **334** in order to form the complex conjugate of the output of the z^{-1}-block **332**. A complex multiplier **336** is provided in order to multiply the output of the unit **334** by the input signal Y_k. The output of the multiplier **336** is a signal Z_k.

The function of the differential detector **330** can be expressed as follows:

$$Z_k=Y_{k+K}\cdot Y_k^* \qquad \text{(Eq.3)}$$

$$Y=[Y_1, Y_2 \ldots, Y_k \ldots] \qquad \text{(Eq.4)}$$

$$Y=[C_1/T_1, C_2/T_1, \ldots, C_{432}/T_1, C_1/T_2 \ldots] \qquad \text{(Eq.5)}$$

Y_k designates the output of the MCM modulator **314**, i.e. the input to the differential detector **330**, at a time k. Z_k designates the output of the differential detector **330**. K designates the number of active carriers.

The output Z_k of the differential detector **330** contains an M-fold uncertainty corresponding to codeable phase shifts. In case of the QPSK mapping, this M-fold uncertainty is a 4-fold uncertainty, i.e., in the 0°, 90°, 180° and 270° phase shifts. This phase shift uncertainty is eliminated from the output Z_k by using an M-PSK decision device **340**. Such decision devices are known in the art and, therefore, are not described here in detail. The output of the decision device **340** $(\hat{a}_k)^*$ represents the complex conjugate of the codeable phase shift decided by the decision device **340**. This output of the decision device **340** is correlated with the output of the differential detector **330** by performing a complex multiplication using a multiplier **342**.

The output the multiplier **342** represents the phase offset for the respective sub-carriers. The phase offsets for the respective sub-carriers are averaged over one MCM symbol in an averaging unit **344** in accordance with a preferred embodiment of the present invention. The output of the averaging units **344** represent the output of the fine frequency error detector **320**.

The mathematical description for this procedure is as follows:

$$f_{offset} = \frac{1}{2\pi K T_{MCM}} \arg\left\{\sum_{n=1}^{K} Z_n \cdot (\hat{a}_n)^*\right\} \quad \text{(Eq. 6)}$$

In accordance with preferred embodiments of the present invention, the frequency control loop has a backward structure. In the embodiment shown in FIG. **6**, the feedback loop is connected between the output of the MCM demodulator **314** and the input of the guard interval removal unit **132**.

In FIG. **8**, a block diagram of a MCM receiver comprising a coarse frequency correction unit **350** and a fine frequency correction unit as described above is shown. As shown in FIG. **8**, a common complex multiplier **326** can be used in order to perform the coarse frequency correction and the fine frequency correction. As shown in FIG. **8**, the multiplier **326** can be provided preceding the low pass filter and decimator unit **312**. Depending on the position of the multiplier **326**, a hold unit has to be provided in the fine frequency synchronization feedback loop. In an alternative embodiment, it is possible to use two separate multipliers for the coarse frequency correction and for the fine frequency correction. In such a case, the multiplier for the coarse frequency correction will be arranged preceding the low path filter and decimator unit, whereas the multiplier for the fine frequency correction will be arranged following the low path filter and decimator unit.

Following, preferred embodiments for implementing a coarse frequency synchronization will be described referring to FIGS. **9** and **10**.

As it is shown in FIG. **9**, the output of the receiver front end **132** is connected to an analog/digital converter **310**. The down-converted MCM signal is sampled at the output of the analog/digital converter **310** and is applied to a frame/timing synchronization unit **360**. In a preferred embodiment, a fast running automatic gain control (AGC) (not shown) is provided preceding the frame/timing synchronization unit in order to eliminate fast channel fluctuations. The fast AGC is used in addition to the normally slow AGC in the signal path, in the case of transmission over a multipath channel with long channel impulse response and frequency selective fading. The fast AGC adjusts the average amplitude range of the signal to the known average amplitude of the reference symbol.

As described above, the frame/timing synchronization unit uses the amplitude-modulated sequence in the received signal in order to extract the framing information from the MCM signal and further to remove the guard intervals therefrom. After the frame/timing synchronization unit **360** it follows a coarse frequency synchronization unit **362** which estimates a coarse frequency offset based on the amplitude-modulated sequence of the reference symbol of the MCM signal. In the coarse frequency synchronization unit **362**, a frequency offset of the carrier frequency with respect to the oscillator frequency in the MCM receiver is determined in order to perform a frequency offset correction in a block **364**. This frequency offset correction in block **364** is performed by a complex multiplication.

The output of the frequency offset correction block **364** is applied to the MCM demodulator **366** formed by the Fast Fourier Transformator **140** and the carrier-bit mapper **142** shown in FIG. **1**.

In order to perform the coarse frequency synchronization described herein, an amplitude-demodulation has to be performed on a preprocessed MCM signal. The preprocessing may be, for example, the down-conversion and the analog/digital conversion of the MCM signal. The result of the amplitude-demodulation of the preprocessed MCM signal is an envelope representing the amplitude of the MCM signal.

For the amplitude demodulation a simple alpha_{max+} beta_{min-} method can be used. This method is described for example in Palacherla A.: DSP-µP Routine Computes Magnitude, EDN, Oct. 26, 1989; and Adams, W. T., and Bradley, J.: Magnitude Approximations for Microprocessor Implementation, IEEE Micro, vol. 3, No. 5, October 1983.

It is clear that amplitude determining methods different from the described alpha_{max+} beta_{min-} method can be used. For simplification, it is possible to reduce the amplitude calculation to a detection as to whether the current amplitude is above or below the average amplitude. The output signal then consists of a −1/+1 sequence which can be used to determine a coarse frequency offset by performing a correlation. This correlation can easily be performed using a simple integrated circuit (IC).

In addition, an oversampling of the signal received at the RF front end can be performed. For example, the received signal can be expressed with two times oversampling.

In accordance with a first embodiment, a carrier frequency offset of the MCM signal from an oscillator frequency in the MCM receiver is determined by correlating the envelope obtained by performing the amplitude-demodulation as described above with a predetermined reference pattern.

In case there is no frequency offset, the received reference symbol r(k) will be:

$$r(k)=S_{AM}(k)+n(k) \quad \text{(Eq.7)}$$

wherein n(k) designates "additive Gaussian noise" and S_{AM} denotes the AM sequence which has been sent. In order to simplify the calculation the additive Gaussian noise can be neglected. It follows:

$$r(k)\cong S_{AM}(k) \quad \text{(Eq.8)}$$

In case a constant frequency offset Δf is present, the received signal will be:

$$\tilde{r}(k)=S_{AM}(k)\cdot e^{j2\pi\Delta f k T_{MCM}} \quad \text{(Eq.9)}$$

Information regarding the frequency offset is derived from the correlation of the received signal $\tilde{r}(k)$ with the AM sequence s_{AM} which is known in the receiver:

$$\sum_{k=1}^{\frac{L}{2}} \tilde{r}(k)\cdot S_{AM}^*(k) = \sum_{k=1}^{\frac{L}{2}} |S_{AM}(k)|^2 e^{j2\pi\Delta f k T_{MCM}} \quad \text{(Eq. 10)}$$

Thus, the frequency offset is:

$$\Delta f = \frac{1}{2\pi T_{MCM}} \arg\left\{\sum_{k=1}^{\frac{L}{2}} r(k)\cdot S_{AM}^*(k)\right\} - \quad \text{(Eq. 11)}$$

-continued

$$\frac{1}{2\pi T_{MCM}}\arg\left(\sum_{k=1}^{\frac{L}{2}}|S_{AM}(k)|^2\right)$$

Since the argument of $|S_{AM}(k)|^2$ is zero the frequency offset is:

$$\Delta f=\frac{1}{2\pi T_{MCM}}\arg\left(\sum_{k=1}^{\frac{L}{2}}\tilde{r}(k)\cdot S_{AM}^*\right) \tag{Eq. 12}$$

In accordance with a second embodiment of the coarse frequency synchronization algorithm, a reference symbol comprising at least two identical sequences **370** as shown in FIG. **10** is used. FIG. **10** shows the reference symbol of a MCM signal having two identical sequences **370** of a length of L/2 each. L designates the number of values of the two sequences **370** of the reference symbol.

AS shown in FIG. **10**, within the amplitude-modulated sequence, there are at least two identical sections devoted to the coarse frequency synchronization. Two such sections, each containing L/2 samples, are shown at the end of the amplitude-modulated sequence in FIG. **10**. The amplitude-modulated sequence contains a large number of samples. For a non-ambiguous observation of the phase, only enough samples to contain a phase rotation of 2π should be used. This number is defined as L/2 in FIG. **10**.

Following, a mathematical derivation of the determination of a carrier frequency deviation is presented. In accordance with FIG. **10**, the following equation applies for the two identical sequences **370**:

$$s\left(0<k\le\frac{L}{2}\right)\equiv s\left(\frac{L}{2}<k\le L\right) \tag{Eq. 13}$$

If no frequency offset is present, the following equation 14 will be met by the received signal:

$$r\left(k+\frac{L}{2}\right)\equiv r(k)\;0<k\le\frac{L}{2} \tag{Eq. 14}$$

r(k) designates the values of the identical sequences. k is an index from one to L/2 for the respective samples.

If there is a frequency offset of, for example, Δf, the received signal is:

$$\tilde{r}(k)=r(k)\cdot e^{j2\pi\Delta fkT_{MCM}} \tag{Eq.15}$$

$$\tilde{r}\left(k+\frac{L}{2}\right)=r(k)\cdot e^{j2\pi\Delta f\left(k+\frac{L}{2}\right)T_{MCM}} \tag{Eq. 16}$$

r(k) designates sample values of the received portion which are based on the identical sequences. Information regarding the frequency offset is derived from the correlation of the received signal $\tilde{r}(k+L/2)$ with the received signal $\tilde{r}(k)$. This correlation is given by the following equation:

$$\sum_{k=1}^{\frac{L}{2}}\tilde{r}^*\left(k+\frac{L}{2}\right)\tilde{r}(k)=\sum_{k=1}^{\frac{L}{2}}|r(k)|^2e^{-j2\pi\Delta f\frac{L}{2}T_{MCM}} \tag{Eq. 17}$$

$\tilde{r}^*$ designates the complex conjugate of the sample values of the portion mentioned above.

Thus, the frequency offset is

$$\Delta f=\frac{1}{2\pi\frac{L}{2}T_{MCM}}\arg\left(\sum_{k=1}^{\frac{L}{2}}\tilde{r}\left(k+\frac{L}{2}\right)\cdot\tilde{r}^*(k)\right)-\frac{1}{2\pi\frac{L}{2}T_{MCM}}\arg\left(\sum_{k=1}^{\frac{L}{2}}|\tilde{r}(k)|^2\right) \tag{Eq. 18}$$

Since the argument of $|\tilde{r}(k)|^2$ equals zero, the frequency offset becomes

$$\Delta f=\frac{1}{2\pi\frac{L}{2}T_{MCM}}\arg\left(\sum_{k=1}^{\frac{L}{2}}\tilde{r}\left(k+\frac{L}{2}\right)\cdot\tilde{r}^*(k)\right) \tag{Eq. 19}$$

Thus, it is clear that in both embodiments, described above, the frequency position of the maximum of the resulting output of the correlation determines the estimated value of the offset carrier. Furthermore, as it is also shown in FIG. **9**, the correction is performed in a feed forward structure.

In case of a channel with strong reflections, for example due to a high building density, the correlations described above might be insufficient for obtaining a suitable coarse frequency synchronization. Therefore, in accordance with a third embodiment of the present invention, corresponding values of the two portions (i.e., which are correlated in accordance with a second embodiment) can be weighted with corresponding values of stored predetermined reference patterns corresponding to said two identical sequences of the reference symbol. This weighting can maximize the probability of correctly determining the frequency offset. The mathematical description of this weighting is as follows:

$$\Delta f=\frac{1}{2\pi\frac{L}{2}T_{MCM}}\arg\left(\sum_{k=1}^{\frac{L}{2}}\left[\tilde{r}\left(k+\frac{L}{2}\right)\cdot\tilde{r}^*(k)\right]\cdot\left[S_{AM}(k)S_{AM}^*\left(k+\frac{L}{2}\right)\right]\right) \tag{Eq. 20}$$

S_{AM} designates the amplitude-modulated sequence which is known in the receiver, and S_{AM}^* designates the complex conjugate thereof.

If the above correlations are calculated in the frequency domain, the amount of

$$\sum_{k=1}^{\frac{L}{2}} \left[\tilde{r}\left(k+\frac{L}{2}\right)\cdot \tilde{r}^*(k)\right]\cdot\left[S_{AM}(k)S_{AM}^*\left(k+\frac{L}{2}\right)\right] \quad \text{(Eq. 21)}$$

is used rather than the argument. This amount is maximized as a function of a frequency correction. The position of the maximum determines the estimation of the frequency deviation. As mentioned above, the correction is performed in a feed forward structure.

Preferred embodiments for performing an echo phase offset correction when using a differential mapping in the frequency axis will be described hereinafter referring to FIGS. **12** to **15**.

Systematic phase shifts stemming from echoes in multipath environments may occur between subcarriers in the same MCM symbol. These phase offsets can cause bit errors when demodulating the MCM symbol at the receiver. Thus, it is preferred to make use of an algorithm to correct the systematic phase shifts stemming from echoes in multipath environments.

In FIG. **12**, scatter diagrams at the output of a differential demapper of an MCM receiver are shown. As can be seen from the left part of FIG. **12**, systematic phase shifts between subcarriers in the same MCM symbol cause a rotation of the demodulated phase shifts with respect to the axis of the complex coordinate system. In the right part of FIG. **12**, the demodulated phase shifts after having performed an echo phase offset correction are depicted. Now, the positions of the signal points are substantially on the axis of the complex coordinate system. These positions correspond to the modulated phase shifts of 0°, 90°, 180° and 270°, respectively.

An echo phase offset correction algorithm (EPOC algorithm) must calculate the echo induced phase offset from the signal space constellation following the differential demodulation and subsequently correct this phase offset.

For illustration purposes, one may think of the simplest algorithm possible which eliminates the symbol phase before computing the mean of all phases of the subcarriers. To illustrate the effect of such an EPOC algorithm, reference is made to the two scatter diagrams of subcarrier symbols contained in one MCM symbol in FIG. **12**. These scatter diagrams have been obtained as result of an MCM simulation. For the simulation, a channel has been used which might typically show up in single frequency networks. The echoes of this channel stretched to the limits of the MCM guard interval. The guard interval was chosen to be 25% of the MCM symbol duration in this case.

FIG. **13** represents a block diagram for illustrating the position and the functionality of an echo phase offset correction device in an MCM receiver. The signal of a MCM transmitter is transmitted through the channel **122** (FIGS. **1** and **13**) and received at the receiver frontend **132** of the MCM receiver. The signal processing between the receiver frontend and the fast Fourier transformator **140** has been omitted in FIG. **13**. The output of the fast Fourier transformator is applied to the de-mapper, which performs a differential de-mapping along the frequency axis. The output of the de-mapper are the respective phase shifts for the subcarriers. The phase offsets of these phase shifts, which are caused by echoes in multipath environments, are illustrated by block **400** in FIG. **13**, which shows an example of a scatter diagram of the subcarrier symbols without an echo phase offset correction.

The output of the de-mapper **142** is applied to the input of an echo phase offset correction device **402**. The echo phase offset correction device **402** uses an EPOC algorithm in order to eliminate echo phase offsets in the output of the demapper **142**. The result is shown in block **404** of FIG. **13**, i.e. only the encoded phase shifts, 0°, 90°, 180° or 270° are present at the output of the correction device **402**. The output of the correction device **402** forms the signal for the metric calculation which is performed in order to recover the bitstream representing the transmitted information.

A first embodiment of an EPOC algorithm and a device for performing the same is now described referring to FIG. **14**.

The first embodiment of an EPOC algorithm starts from the assumption that every received differentially decoded complex symbol is rotated by an angle due to echoes in the multipath channel. For the subcarriers equal spacing in frequency is assumed since this represents a preferred embodiment. If the subcarriers were not equally spaced in frequency, a correction factor would have to be introduced into the EPOC algorithm.

FIG. **14** shows the correction device **402** (FIG. **13**) for performing the first embodiment of an EPOC algorithm.

From the output of the de-mapper **142** which contains an echo phase offset as shown for example in the left part of FIG. **12**, the phase shifts related to transmitted information must first be discarded. To this end, the output of the de-mapper **142** is applied to a discarding unit **500**. In case of a DQPSK mapping, the discarding unit can perform a "$(.)^4$" operation. The unit **500** projects all received symbols into the first quadrant. Therefore, the phase shifts related to transmitted information is eliminated from the phase shifts representing the subcarrier symbols. The same effect could be reached with a modulo-4 operation.

Having eliminated the information related symbol phases in unit **500**, the first approach to obtain an estimation would be to simply compute the mean value over all symbol phases of one MCM symbol. However, it is preferred to perform a threshold decision before determining the mean value over all symbol phases of one MCM symbol. Due to Rayleigh fading some of the received symbols may contribute unreliable information to the determination of the echo phase offset. Therefore, depending on the absolute value of a symbol, a threshold decision is performed in order to determine whether the symbol should contribute to the estimate of the phase offset or not.

Thus, in the embodiment shown in FIG. **14**, a threshold decision unit **510** is included. Following the unit **500** the absolute value and the argument of a differentially decoded symbol is computed in respective computing units **512** and **514**. Depending on the absolute value of a respective symbol, a control signal is derived. This control signal is compared with a threshold value in a decision circuit **516**. If the absolute value, i.e. the control signal thereof, is smaller than a certain threshold, the decision circuit **516** replaces the angle value going into the averaging operation by a value equal to zero. To this end, a switch is provided in order to disconnect the output of the argument computing unit **514** from the input of the further processing stage and connects the input of the further processing stage with a unit **518** providing a constant output of "zero".

An averaging unit **520** is provided in order to calculate a mean value based on the phase offsets ϕ_i determined for the individual subcarrier symbols of a MCM symbol as follows:

$$\overline{\varphi} = 1/K\sum_{i=1}^{K}\varphi_i \quad \text{(Eq. 22)}$$

In the averaging unit **520**, summation over K summands is performed. The output of the averaging unit **520** is provided to a hold unit **522** which holds the output of the averaging unit **520** K times. The output of the hold unit **522** is connected with a phase rotation unit **524** which performs the correction of the phase offsets of the K complex signal points on the basis of the mean value $\overline{\phi}$.

The phase rotation unit **524** performs the correction of the phase offsets by making use of the following equation:

$$v_k' = v_k \cdot e^{-j\overline{\phi}} \quad \text{(Eq.23)}$$

In this equation, v_k' designates the K phase corrected differentially decoded symbols for input into the soft-metric calculation, whereas v_k designates the input symbols. As long as a channel which is quasi stationary during the duration of one MCM symbols can be assumed, using the mean value over all subcarriers of one MCM symbol will provide correct results.

A buffer unit **527** may be provided in order to buffer the complex signal points until the mean value of the phase offsets for one MCM symbol is determined. The output of the phase rotation unit **524** is applied to the further processing stage **526** for performing the soft-metric calculation.

With respect to the results of the above echo phase offset correction, reference is made again to FIG. **12**. The two plots stem from a simulation which included the first embodiment of an echo phase offset correction algorithm described above. At the instant of the scatter diagram snapshot shown in the left part of FIG. **12**, the channel obviously distorted the constellation in such a way, that a simple angle rotation is a valid assumption. As shown in the right part of FIG. **12**, the signal constellation can be rotated back to the axis by applying the determined mean value for the rotation of the differentially detected symbols.

A second embodiment of an echo phase offset correction algorithm is described hereinafter. This second embodiment can be preferably used in connection with multipath channels that have up to two strong path echoes. The algorithm of the second embodiment is more complex than the algorithm of the first embodiment.

What follows is a mathematical derivation of the second embodiment of a method for echo phase offset correction. The following assumptions can be made in order to ease the explanation of the second embodiment of an EPOC algorithm.

In this embodiment, the guard interval of the MCM signal is assumed to be at least as long as the impulse response h[q], q=0, 1, . . . , Qh−1 of the multipath channel.

At the transmitter every MCM symbol is assembled using frequency axis mapping explained above. The symbol of the reference subcarrier equals 1, i.e. 0 degree phase shift. The optional phase shift PHI equals zero, i.e. the DQPSK signal constellation is not rotated.

Using an equation this can be expressed as

$$a_k = a_{k-1} a_k^{inc} \quad \text{(Eq.24)}$$

with

k	index k = 1, 2, . . . , K of the active subcarrier;
$a_k^{inc} = e^{j\frac{\pi}{2}m}$	complex phase increment symbol; m = 0, 1, 2, 3 is the QPSK symbol number which is derived from Gray encoding pairs of 2 Bits;
a0 = 1	symbol of the reference subcarrier.

At the DFT output of the receiver the decision variables

$$e_k = a_k H_k \quad \text{(Eq.25)}$$

are obtained with

$$H_k = \sum_{i=0}^{Q_h-1} h[i] \cdot e^{-j\frac{2\pi}{K}ki} \quad \text{(Eq. 26)}$$

being the DFT of the channel impulse response h[q] at position k.

With $|ak|^2=1$ the differential demodulation yields

$$v_k = e_k \cdot e_{k-1}^* = a_k^{inc} H_k H_{k-1}^* \quad \text{(Eq.27)}$$

For the receiver an additional phase term ϕ_k is introduced, which shall be used to correct the systematic phase offset caused by the channel. Therefore, the final decision variable at the receiver is

$$v_k' = v_k \cdot e^{j\phi k} = a_k^{inc} \cdot e^{j\phi k} \cdot H_k \cdot H_{k-1}^* \quad \text{(Eq.28)}$$

As can be seen from the Equation 28, the useful information a_k^{inc} is weighted with the product $e^{j\phi_k} \cdot H_k \cdot H_{k-1}^*$ (rotation and effective transfer function of the channel). This product must be real-valued for an error free detection. Considering this, it is best to choose the rotation angle to equal the negative argument of $H_k \cdot H_{k-1}^*$. To derive the desired algorithm for 2-path channels, the nature of $H_k \cdot H_{k-1}^*$ is investigated in the next section.

It is assumed that the 2-path channel exhibits two echoes with energy content unequal zero, i.e. at least two dominant echoes. This assumption yields the impulse response

$$h[q] = c_1\delta_0[q] + c_2\delta_0[q-q_0] \quad \text{(Eq.29)}$$

with

c_1, c_2:	complex coefficients representing the path echoes;
q_0:	delay of the second path echo with respect to the first path echo;
δ_0:	Dirac pulse; $\delta_0[\kappa]$ = 1 for k = 0; $\delta_0[\kappa]$ = 0 else

The channel transfer function is obtained by applying a DFT to Equation 29:

$$H_k = H\left(e^{j\frac{2\pi}{K}k}\right) = c_1 + c_2 \cdot e^{-j\frac{2\pi}{K}kq_0} \quad \text{(Eq. 30)}$$

With Equation 30 the effective transfer function for differential demodulation along the frequency axis is:

$$H_k \cdot H_{k-1}^* = \left(c_1 + c_2 e^{-j\frac{2\pi}{K}kq_0}\right)\cdot\left(c_1^* + c_2^* e^{+j\frac{2\pi}{K}(k-1)q_0}\right) = c_a + c_b \cos\left(\frac{\pi}{K}q_0(2k-1)\right) \quad \text{(Eq. 31)}$$

Assuming a noise free 2-path channel, it can be observed from Equation 31 that the symbols on the receiver side are located on a straight line in case the symbol 1+j0 has been send (see above assumption). This straight line can be characterized by a point

$$c_a = |c_1|^2 + |c_2|^2 \cdot e^{-j\frac{2\pi}{K}q_0} \quad \text{(Eq. 32)}$$

and the vector

$$c_b = 2c_1c_2^* \cdot e^{-j\frac{\pi}{K}q_0} \quad \text{(Eq. 33)}$$

which determines its direction.

With the above assumptions, the following geometric derivation can be performed. A more suitable notation for the geometric derivation of the second embodiment of an EPOC algorithm is obtained if the real part of the complex plane is designated as x=Re{z}, the imaginary part as y=Im{z}, respectively, i.e. z=x+jy. With this new notation, the straight line, on which the received symbols will lie in case of a noise-free two-path channel, is

$$f(x)=a+b\cdot x \quad \text{(Eq.34)}$$

with

$$a = \mathrm{Im}\{c_a\} - \frac{\mathrm{Re}\{c_a\}}{\mathrm{Re}\{c_b\}}\cdot \mathrm{Im}\{c_b\} \quad \text{(Eq. 35)}$$

and

$$b = -\frac{\mathrm{Im}\{c_a\} - \frac{\mathrm{Re}\{c_a\}}{\mathrm{Re}\{c_b\}}\cdot \mathrm{Im}\{c_b\}}{\mathrm{Re}\{c_a\} - \frac{\mathrm{Im}\{c_a\}}{\mathrm{Im}\{c_b\}}\cdot \mathrm{Re}\{c_b\}} \quad \text{(Eq. 36)}$$

Additional noise will spread the symbols around the straight line given by Equations 34 to 36. In this case Equation 36 is the regression curve for the cluster of symbols.

For the geometric derivation of the second embodiment of an EPOC algorithm, the angle ϕ_k from Equation 28 is chosen to be a function of the square distance of the considered symbol from the origin:

$$\phi_k = f_K(|z|^2) \quad \text{(Eq.37)}$$

Equation 37 shows that the complete signal space is distorted (torsion), however, with the distances from the origin being preserved.

For the derivation of the algorithm of the second embodiment, $f_K(\cdot)$ has to be determined such that all decision variables v_k' (assuming no noise) will come to lie on the real axis:

$$Im\{(x+jf(x))\cdot e^{jf_K(|z|^2)}\}=0 \quad \text{(Eq.38)}$$

Further transformations of Equation 38 lead to a quadratic equation which has to be solved to obtain the solution for ϕ_k.

In case of a two-path channel, the echo phase offset correction for a given decision variable v_k is

$$v_k' = v_k \cdot e^{j\phi_k} \quad \text{(Eq.39)}$$

with

$$\varphi_k = \begin{cases} -a\tan\left(\dfrac{a + b\sqrt{|v_k|^2(1+b^2) - a^2}}{-ab + \sqrt{|v_k|^2(1+b^2) - a^2}}\right) & \text{for } |v_k|^2 \geq \dfrac{a^2}{1+b^2} \\ a\tan\left(\dfrac{1}{b}\right) & \text{for } |v_k|^2 < \dfrac{a^2}{1+b^2} \end{cases} \quad \text{(Eq. 40)}$$

From the two possible solutions of the quadratic equation mentioned above, Equation 40 is the one solution that cannot cause an additional phase shift of 180 degrees.

The two plots in FIG. **15** show the projection of the EPOC algorithm of the second embodiment for one quadrant of the complex plane. Depicted here is the quadratic grid in the sector $|\arg(z)| \leqq \pi/4$ and the straight line y=f(x)=a+b·x with a =−1.0 and b=0.5 (dotted line). In case of a noise-free channel, all received symbols will lie on this straight line if 1+j0 was sent. The circle shown in the plots determines the boarder line for the two cases of Equation 40. In the left part, FIG. **15** shows the situation before the projection, in the right part, FIG. **15** shows the situation after applying the projection algorithm. By looking on the left part, one can see, that the straight line now lies on the real axis with 2+j0 being the fix point of the projection. Therefore, it can be concluded that the echo phase offset correction algorithm according to the second embodiment fulfills the design goal.

Before the second embodiment of an EPOC algorithm can be applied, the approximation line through the received symbols has to be determined, i.e. the parameters a and b must be estimated. For this purpose, it is assumed that the received symbols lie in sector $|\arg(z)| \leqq \pi/4$, if 1+j0 was sent. If symbols other than 1+j0 have been sent, a modulo operation can be applied to project all symbols into the desired sector. Proceeding like this prevents the necessity of deciding on the symbols in an early stage and enables averaging over all signal points of one MCM symbol (instead of averaging over only ¼ of all signal points).

For the following computation rule for the EPOC algorithm of the second embodiment, x_i is used to denote the real part of the i-th signal point and y_i for its imaginary part, respectively (i=1, 2, . . . , K). Altogether, K values are available for the determination. By choosing the method of least squares, the straight line which has to be determined can be obtained by minimizing

$$(a, b) = \arg\min_{(\tilde{a},\tilde{b})} \sum_{i=1}^{K} \left(y_i - (\tilde{a} + \tilde{b}\cdot x_i)\right)^2 \quad \text{(Eq. 41)}$$

The solution for Equation 41 can be found in the laid open literature. It is

$$b = \frac{\sum_{i=1}^{K}(x_i - \bar{x})\cdot y_i}{\sum_{i=1}^{K}(x_i - \bar{x})^2}, \quad a = \bar{y} - \bar{x}\cdot b \qquad \text{(Eq. 42)}$$

with mean values

$$\bar{x} = \frac{1}{N}\sum_{i=1}^{K} x_i, \quad \bar{y} = \frac{1}{N}\sum_{i=1}^{K} y_i \qquad \text{(Eq. 43)}$$

If necessary, an estimation method with higher robustness can be applied. However, the trade-off will be a much higher computational complexity.

To avoid problems with the range in which the projection is applicable, the determination of the straight line should be separated into two parts. First, the cluster's centers of gravity are moved onto the axes, following, the signal space is distorted. Assuming that a and b are the original parameters of the straight line and a is the rotation angle, $f_K(\cdot)$ has to be applied with the transformed parameters

$$b' = \frac{b\cdot\cos(\alpha) - \sin(\alpha)}{\cos(\alpha) + b\cdot\sin(\alpha)}, \quad a' = a\cdot(\cos(\alpha) - b'\cdot\sin(\alpha)) \qquad \text{(Eq. 44)}$$

Besides the two EPOC algorithms explained in the above section, different algorithms can be designed that will, however, most likely exhibit a higher degree of computational complexity.

The invention claimed is:

1. A method of performing a fine frequency synchronization compensating for a carrier frequency deviation from an oscillator frequency in a multi-carrier demodulation system capable of carrying out a differential phase decoding of multi-carrier modulated signals, said signals comprising a plurality of symbols, each symbol being differentially coded in the direction of the frequency axis, said method comprising the steps of:

a) determining a phase difference between phases of the same carrier in different symbols;

b) determining a frequency offset by eliminating phase shift uncertainties related to the transmitted information from said phase difference making use of a M-PSK decision device; and

c) performing a feedback correction of said carrier frequency deviation based on said determined frequency offset, wherein

said steps a) and b) are performed for a plurality of carriers in said symbols,

an averaged frequency offset is determined by averaging said determined frequency offsets of said plurality of carriers, and

said feedback correction of said frequency deviation is performed based on said averaged frequency offset.

2. The method according to claim **1**, wherein said step a) comprises the step of determining a phase difference between phases of the same carrier in symbols which are adjacent in the time axis direction.

3. The method according to claim **1**, wherein said step b) comprises the step of eliminating phase shift uncertainties corresponding to M-ary phase shifts.

4. An apparatus for performing a fine frequency synchronization compensating for a carrier frequency deviation from an oscillator frequency, for a multi-carrier demodulation system capable of carrying out a differential phase decoding of multi-carrier modulated signals, said signals comprising a plurality of symbols, each symbol being defined by

phase differences between simultaneous carriers having different frequencies, said apparatus comprising:

means for determining respective phases of the same carrier in different symbols;

M-PSK decision device for eliminating phase shift uncertainties related to the transmitted information from said phases to determine respective phase deviations;

means for determining a frequency offset by determining a phase difference between said phase deviations; and

means for performing a feedback correction of said frequency deviation based on said determined frequency offset;

wherein said means for determining respective phases comprises means for determining respective phases of the same carrier in symbols which are adjacent in the time axis direction.

5. The apparatus according to claim **4**, further comprising:

means for determining an averaged frequency offset by averaging determined frequency offsets of a plurality of carriers, wherein

said means for performing a feedback correction performs said feedback correction of said frequency deviation based on said averaged frequency offset.

6. The apparatus according to claim **4**, wherein said means for performing a feedback correction of said frequency deviation comprises a numerical controlled oscillator and a complex multiplier.

7. The apparatus according to claim **6**, wherein said means for performing a feedback correction of said frequency deviation further comprises a low path filter preceding said numerical controlled oscillator.

* * * * *

Exhibit E

[Execution Copy]

Firm Fixed Price Contract

Contract #001

Technical Consulting on XM Radio

BETWEEN

XM Satellite Radio Inc.
1250 23rd Street, N.W.
Suite 57
Washington, D.C. 20037

AND

Fraunhofer Gesellschaft zur Förderung
Der angewandten Forschung e.V.,

ADDRESS

Leonrodstr. 54
D-80636 Munchen, Germany

July 16, 1999

791716.10

Firm Fixed Price Contract
NO. 001

This Contract is made and entered into this 16th day of July, 1999 by and between **XM SATELLITE RADIO INC.**, a satellite digital audio radio services (DARS) company, with offices at 1250 23rd Street, N.W., Washington, DC 20037, U.S.A. (hereinafter referred to as the "Customer" which expression shall include its successors and permitted assigns) and the **FRAUNHOFER-GESELLSCHAFT ZUR FÖRDERUNG DER ANGEWANDTEN FORSCHUNG E.V.,** a corporation organized and existing under the laws of Germany, (hereinafter referred to as "FhG" which expression shall include its successors and permitted assigns).

WITNESSETH THAT: The Customer and FhG (collectively referred to hereunder as "Parties" and individually as "Party") hereto mutually agree as follows:

1. CONTRACT TYPE AND SCOPE OF WORK

FhG shall furnish the necessary personnel, equipment, material, services and facilities to provide the deliverables specified in Exhibit A ("Statement of Work") and in accordance with the Payment Plan in Exhibit B.

2. PERIOD OF PERFORMANCE

The period of performance for this Contract is July 16, 1999 through December 30, 2000.

3. PRICE

3.1 For the performance of the requirements of this Firm-Fixed Price Contract, FhG shall receive payment by Customer in DM, which shall be paid in accordance with Article 4 of this Contract entitled "Payment".

3.2 All travel costs incurred by FhG in any month in connection with its performance under this Contract shall be communicated to Customer in the following month in a report using the form attached hereto as Exhibit E, including any other information that Customer may reasonably request from time to time. Customer shall reimburse FhG the amounts set forth in such reports, on a fixed sum per trip basis, as set out in Exhibit B, subject to Customer's reasonable approval, in accordance with the terms of Article 4 hereof.

4. PAYMENT

4.1 Payment shall be made by customer in accordance with completion of milestones by FhG set forth in Exhibit B. All payments shall be in DM until such time as the Euro is the only legal currency in the Federal Republic of Germany, at which time the payment obligations hereunder, particularly the monetary value stipulated in this Contract, shall be regarded as being stipulated in Euro. In any case the conversion of D-Mark in Euro will be effected on the basis of the then-current published fixed conversion rate, which is currently 1.95583 DM per Euro.

4.2 Any amounts due to FhG shall be paid within thirty (30) days after FhG provides to Customer (i) a facsimile invoice stating the milestone (with reference to the Payment Plan) for which payment is being requested and (ii) an e-mail to Stell Patsiokas and Paul Marko notifying them that such invoice has been faxed. Fixed monthly payments according to the Payment Plan (Exhibit B) shall be paid by Customer within thirty (30) days after FhG provides an invoice for such amount to Customer. Customer shall have ten (10) business days following receipt of such invoice to notify FhG that such milestone has not been met in accordance with the requirements of this Contract. The Parties acknowledge that the "substantial" standard for performance set forth in Article 18.1 hereof does not apply to the acceptance of performance milestones. If FhG does not receive such notice within such ten (10) business day period, then such milestone shall be deemed to have been met. Payment shall be made by wire transfer to FhG, Reference No. 44-808636, Deutsche Bank München, Account No. 7421 933, BLZ 700 700 01, or said payments shall be express couriered to FhG at the address shown in Article 26.1 of the Contract.

4.3 Any amounts for payments under Exhibit B that have been met or deemed to have been met that are not received by FhG within the above stated time periods shall bear interest from the date of invoice, at the rate of one percent (1%) per month or the maximum rate allowed by law, whichever is lower. If any invoiced amount is in dispute, the Party disputing such amount shall notify the other Party thereof in writing within ten (10) business days after such amount was due. No interest shall be payable unless, until, and to the extent the dispute is settled in favor of FhG. The Parties agree that work under this Contract shall continue during the resolution of such dispute.

4.4 Subject to Article 4.3, and provided FhG is not in default hereunder, the Customer's failure to make payments in accordance with the terms of this Contract shall be deemed to be a default. FhG will provide the Customer with a written notice of the default and if the default is not cured by the Customer within 15 days of receipt of FhG's written notice, FhG may suspend all work hereunder until such default is cured. If such default is not cured within 30 days of receipt of FhG's written notice, FhG may terminate work under this Contract without further obligation to the Customer. In the event of such termination, the Customer's liability shall be determined by the provisions relating to the Customer's liability in Article 19 of the Contract hereof entitled Termination for Convenience.

4.5 In the event FhG suspends work due to late payment by the Customer, as described in Article 4.4 above, and if following receipt of payment by the Customer FhG subsequently proceeds with the work, the period of performance set forth in this Contract shall be extended by an amount of time equal to the period of time necessary for FhG to reconstitute its efforts plus an amount of time equal to the number of days that the work was suspended. Furthermore, any additional cost impact resulting from such suspension shall be determined and agreed by the Parties as an allowable cost under this Contract. Any extension of time or additional cost agreed upon by the Parties shall be included in an agreement to amend the Contract. The Parties agree that work under this Contract shall continue during the negotiation of such additional costs (or any disputes related thereto).

5. TECHNICAL AND CONTRACTUAL REPRESENTATIVES

The following authorized representatives are hereby designated for this Contract:

XM Satellite Radio Inc.	FhG
Technical and Program: Stell Patsiokas	Technical and Program: Stefan Meltzer
Contractual: Joseph Titlebaum	Contractual: Dr. Birgit Homma

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6. TAXES AND DUTIES

All national, regional or local taxes, duties, and similar liabilities, shall be paid by FhG.

7. CONSULTATIONS

At the request of Customer, FhG shall make available all relevant key personnel pertaining to this Contract and the Statement of Work for meetings either in the United States or Europe in connection with design reviews, troubleshooting, other requests for assistance, or progress reports.

8. FORCE MAJEURE

8.1 FhG shall not be liable for any loss damage, detention, or delay resulting from causes beyond its reasonable control, including, but not limited to: acts of God, acts of governments in their sovereign capacity (except to the extent that such acts of governments arise out of an act or failure to act by FhG), fires, floods, epidemics, quarantine restrictions, strikes labor disputes, freight embargoes, unusually severe weather, insurrection or riot, damage in transportation, and inability due to causes beyond its reasonable control to obtain necessary labor, materials, or facilities. In the event of a temporary delay specified in this Contract is required, an equitable adjustment shall be made in the Contract price and the performance dates hereof shall be extended at least by an amount of time equal to the number of days that work was suspended, not to exceed a reasonable length of time. In the event the parties agree that it is reasonably likely that such an excusable delay will exceed forty-five (45) days (an "Extended Force Majeur") resulting from any such causes, the Customer shall be entitled to terminate this Contract in accordance with Article 17 hereof entitled Termination for Extended Force Majeure. The Parties acknowledge that the occurrence of any force majeur event described herein shall not excuse any default by FhG existing prior to the occurrence of such event.

8.2 FhG shall take all reasonable steps to mitigate the impact of any force majeure event.

9. INDEMNITY AND LIMITATION OF LIABILITY

9.1 The liability of FhG with respect to any service, sale, or anything done in connection therewith, such as the performance or breach thereof, or from the manufacture, sale, delivery, resale, installation or use of any goods or services covered by or furnished under this Contract whether arising out of statute, contract, negligence, strict liability in tort, or under any warranty, or otherwise, and whether or not occasioned by FhG's negligence, shall not exceed 25% of the price of each work package (*i.e.*, system engineering, channel decoder development and test equipment) affected by the action or inaction resulting in such liability and resulting from the willful misconduct or gross negligence of FhG.

9.2 Notwithstanding any other provision of this Contract, neither Party shall under any circumstances be liable for special, incidental, indirect or consequential damages, such as, but not limited to, loss or damage of other property or equipment, loss of profits or revenue cost of capital, cost of purchased or replaced goods, or claims of customers or contractors of the other Party for, but not limited to delays, penalties or service interruptions.

9.3 FhG shall indemnify, defend and hold harmless Customer against any costs or expenses, with the exception of those costs identified in Article 9.2 above, including reasonable attorneys' fees, incurred by Customer as a result of legal actions brought against Customer by third parties caused by default of FhG in the performance of the work under the Contract, subject to the limitation of liability set forth as Clause 9.1 above.

9.4 Notwithstanding the foregoing, the Customer and FhG agree to a no-fault, no-subrogation inter-party waiver of liability under which each Party shall be responsible for any damage it sustains as a result of damage to its own property and employees, including death, while involved in the conduct of the activities which are the subject of this Contract, which damage is not caused by the other Party. It is the intent of the Parties that this inter-party waiver of liability be construed broadly to achieve the intended objectives.

10. PROPRIETARY INFORMATION AND INTELLECTUAL PROPERTY

10.1 Each Party shall identify to the other, and the receiving Party shall hold in confidence, any proprietary or confidential information marked as proprietary or confidential or obtained in connection with FhG's work under this Contract, any deliverables hereunder, or any proprietary or confidential information so marked furnished by one Party to the other. Subject to Customer's rights described in Article 10.2 hereof, each Party shall use the same efforts to avoid disclosure, publication or dissemination of such proprietary or confidential information as they use with respect to their own proprietary or confidential information, but in no event less than best efforts. Said information shall remain the proprietary information of the Party disclosing it and shall not be disclosed to others without the disclosing Party's prior written consent either during or after the term of the Contract. All technical data based upon proprietary or confidential information furnished by Customer that is essential to the design, function or operation of any deliverable under this Contract (including, without limitation, any patents or patent applications anywhere in the world owned by or licensed to Customer) shall be considered as the Customer's proprietary or confidential data and shall not be disclosed to others without the Customer's prior written consent either during or after the term of the Contract. Proprietary or confidential information or data shall not include information or data which becomes generally known in the industry, or is known to either Party prior to its disclosure by the other Party as demonstrated by written records, or is authorized in writing by the disclosing Party for release, or which is subject to judicial or governmental compelled disclosure.

10.2 FhG hereby grants to Customer:

(a) An exclusive, paid-up, royalty-free, transferable, perpetual, and irrevocable world-wide license in and to the synthesizable VHDL code or other source or executable code for channel decoder developed by FhG under this Contract and all patent applications, patents, trade secrets, mask works, copyrights and other intellectual property for work or inventions performed or developed in connection with such Code for Customer (the "Code"). Such license shall include the right of Customer to use, copy, maintain, modify, create derivative works of, transfer, sublicense or otherwise convey the Code as necessary in connection with Customer's business in Customer's sole discretion and without the consent of FhG.

(b) A non-exclusive, non-transferable, paid-up, royalty-free perpetual and irrevocable worldwide license to use the test equipment developed hereunder in connection with Customer's business.

(c) The Parties acknowledge that nothing in this Article 10.2 is intended to grant to Customer the rights to modify, transfer or otherwise convey any rights in the FhG Background IP (defined in Article 10.5).

10.3 FhG agrees that its ownership rights in the Code shall be limited by the following:

(a) FhG shall not sell, transfer, distribute, loan, disclose, reproduce or otherwise convey the Code (in its entirety as delivered to Customer hereunder) to or for any other person or entity. FhG acknowledges that the Code comprises a material element in Customer's digital radio system and that Customer would suffer material financial and market share losses in the event the Code were to be disclosed. Accordingly, FhG agrees to maintain strict confidentiality of the Code.

(b) Subject to the limitation described in Article 10.3(c) below, FhG may modify the Code for use in future projects for which FhG is commissioned by future FhG clients; provided, however, that such modifications shall be substantial enough to make the Code unrecognizable to such clients and such that such modifications will not be able to be reverse-engineered to obtain the Code.

(c) Under no circumstances shall FhG modify the Code for any Customer Competitor. For purposes of this Contract, "Customer Competitor" shall mean (i) any entity involved in the business of satellite or terrestrial digital broadcasting in North or South America, (ii) any individual employed by, serving as an officer or director for, or owning any equity interest in, such an entity, (iii) any entity controlling, controlled by, or under common control with such an entity, or also (iv) any entity owning any equity interest in such an entity. "Customer Competitor" shall also include the following companies, as well as any other company that Customer notifies FhG in writing shall be considered a Customer Competitor: CD Radio and any company involved with IBOC technology or having an equity interest in a license for the wireless communication system (WCS) frequency spectrum. "Customer Competitor" shall not include WorldSpace Satellite Company or any principal affiliate thereof as it operates in South America.

(d) For FhG and FhG customer purposes other than XM Radio Program related projects, FhG will retain all rights to reuse the following blocks and subblocks included in the channel decoder design:

- MCM Demodulation
- External Memory Controller
- Viterbi Convolutial Decoder including Depuncturing Unit
- FEC Management Controller
- Bit Error Rate (BER) Measurement Controller
- I2C Interface Controller.

10.4 Customer hereby grants to FhG a royalty-free, non-exclusive, non-transferable license for the development, manufacturing and sale of the test equipment specified in Exhibit A; provided, however, that FhG shall be entitled to sublicense such rights to Fraunhofer IZT. FhG shall inform its customers that such license is

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for engineering development uses and not for consumer receiver production tests.

10.5 The Parties acknowledge that the intellectual property listed in Exhibit D hereto, as may be modified from time to time during the term of this Contract, belonging to FhG, with associated patent applications, patents, trade secrets, and copyright references identified (the "FhG Background IP") are all of the technologies owned by FhG required to complete the deliverables under this Contract. FhG hereby grants a royalty-free, non-exclusive, non-transferable perpetual and irrevocable worldwide right and license to Customer to such FhG Background IP. Customer shall provide FhG on a non-exclusive, non-transferable, royalty-free basis with all licenses to intellectual property owned by Customer necessary for the performance of the Contract.

10.6 FhG shall notify Customer promptly upon its filing of any patents resulting from or in connection with the deliverables under this Contract.

11. PATENT INDEMNITY

11.1 FhG shall reasonably ensure that the FhG Background IP and any intellectual property arising out of this Agreement embodied in each deliverable does not infringe on third Party patents or other proprietary right, and FhG represents that, to the best of its knowledge, no such infringement claims are pending or threatened against FhG. FhG shall promptly notify the Customer in writing of potential patent infringement claims related to the FhG Background IP and any intellectual property arising out of this Agreement and FhG shall diligently defend or settle such claims at its own expense.

11.2 In the event that use of the services and deliverables to be provided by FhG hereunder is enjoined or that the parties mutually determine that an infringement claim is likely, FhG shall either obtain the necessary license for the Customer or modify the subject deliverable to avoid infringement. The Customer would have no further remedy beyond these two actions. The costs of modifications or additional licenses shall be borne by Customer up to $500,000. FhG shall in no respect bear the costs of additional licenses.

11.3 FhG agrees to indemnify and hold harmless the Customer and its officers and defend at its own expense any claims, actions, or proceeding based on an allegation that FhG uses materials or items for the performance of services or provision of goods under this Contract which directly infringes any trademark, copyright, or trade secret, provided that FhG is given prompt written notice of such claims by the Customer. Notwithstanding the foregoing, FhG shall have no liability or responsibility for any infringement resulting from FhG following the directions of or specifications provided by the Customer.

11.4 This Article 11 is subject to the limitations of Article 9.1.

12. COPYRIGHT

Copyright in all reports and other documents which are produced by FhG and delivered under this Contract shall vest in and be the sole property of the Customer. The foregoing shall not be deemed to preclude FhG from making and retaining copies of such materials for record keeping purposes. All such reports and other documents shall clearly state that they were authored by FhG.

13. WARRANTY

FhG warrants that the services provided under this Contract and the Statement of Work shall be carried out with all reasonable skill, care and diligence. Any claim for breach of the foregoing warranty shall be deemed to have been waived unless asserted in writing within one (1) year after completion of performance of the services to be provided under this Contract. FhG agrees to correct defects discovered by Customer within 30 days of notification of such defect(s).

FhG warrants that the products and other deliverables provided under this Contract shall be free from defects in materials and workmanship under normal use and service, and such products and other deliverables will be new, of good quality, and of recent manufacture (unless otherwise permitted by Customer) for one (1) year from the date of acceptance of such product or deliverable by Customer.

14. DISPUTES

FhG and the Customer shall make every effort to reach an amicable settlement of any dispute or disagreement arising under this Contract including if deemed appropriate and requested by either Party submission to the principal officers of FhG and the Customer. If no agreement can be reached within 7 days, the matter shall be settle definitively by three (3) arbitrators, using the Rules of the International Chamber of Commerce (ICC), who shall sit in London, England applying the laws of Switzerland. Each Party shall designate one (1) arbitrator and both Parties shall designate the third arbitrator. All arbitration shall be in the English language. The arbitration award shall be final and binding upon the Parties and judgment may be entered thereon, upon the application of either Party, by any court having jurisdiction. Each Party shall bear the cost of preparing and presenting its case, and the cost of the arbitration (including fees and expenses of the arbitrators) shall be shared equally by the Parties unless the award otherwise provides.

Pending a decision by the arbitrators, each Party shall, unless directed by the other Party in writing, fulfill all of its obligations under this Contract, including the obligation to take all steps necessary during the pendency of the arbitration to ensure the services and deliverables will be delivered within the time stipulated, or within such extended time as may be allowed under this Contract, provided Customer shall continue to make payments therefore in accordance with this Contract.

15. GOVERNMENT APPROVALS

15.1 FhG shall be responsible for obtaining any governmental authorizations, consents, and approvals in Germany necessary for the performance of FhG's obligations herein. In the event that lawful performance of this Contract or any part of this Contract by either Party is delayed or rendered impossible by, or as a consequence of any law, regulation, or any Government having jurisdiction over it, such Party shall not be considered to be in default by reason of such delay or failure to perform, and the Parties shall consult in good faith to develop an equitable resolution of the issue with due regard for the respective interests of the Parties.

15.2 All provisions in this Article shall remain binding on the Parties after the termination of the Contract.

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16. LANGUAGE AND COMMUNICATIONS

16.1 All data, documents, descriptions, reports, certificates, studies, technical data provided by FhG shall be written in English.

16.2 This Contract and all documentation and communications required hereunder, shall be in the English language.

17. TERMINATION FOR EXTENDED FORCE MAJEURE

In the event of an Extended Force Majeur resulting from a force majeure event as stated in Article 8, the Customer by written notice, may terminate this Contract in whole or in part at any time upon giving thirty (30) days notice to FhG. The Customer shall pay FhG for the services rendered and ODC's incurred up to the date of termination in accordance with the provisions of Article 4 of this Contract. FHG shall take all reasonable steps to mitigate costs incurred after receiving notice of any such force majeure event.

18. TERMINATION FOR DEFAULT

18.1 The Customer may, by written Notice of Default to FhG, terminate the whole or any part of this Contract in any one of the following circumstances; (1) if FhG fails substantially to make delivery of any deliverable as defined in Exhibit A or to perform the services within the time specified herein; or (2) if FhG fails to make progress as to materially endanger performance of this Contract in accordance with its terms; or (3) the amounts described in Article 11.2 hereof exceed the limit stated therein; or (4) if proceedings are commenced or threatened, the result of which will be to place FhG into liquidation, receivership or administration or FhG otherwise becomes insolvent or admits its inability to meet its debts as they fall due or if it enters into any form of composition or arrangement with its creditors and in either of (1) (2) or (3) above FhG does not effect a satisfactory plan to cure such failure within a period of thirty (30) days (or such longer period as the Customer may authorize in writing) after receipt of notice from the Customer specifying such failure.

18.2 In the event of a default under Article 18.1 above, the Customer may: (i) rescind the whole or portion of the Contract so terminated whereupon FhG shall promptly reimburse the Customer all amounts previously paid to FhG by the Customer for the work so terminated (less the amounts for supplies or services delivered and accepted or performed and accepted prior to the date of termination or desired by the Customer notwithstanding such termination). .FhG shall have no further liability.

19. TERMINATION FOR CONVENIENCE

The Customer may, by thirty (30) calendar days written notice, terminate the whole or any part of this Contract for any reason. All work performed and expenses incurred by FhG up to and including said thirty (30) day period shall be paid by Customer according to the provisions of this agreement. In the event of termination for convenience, FhG shall receive termination fees in an amount equal to FhG's reasonable wind-down costs plus ten percent (10%) of such wind-down costs.

20. KEY PERSONNEL

20.1 FhG agrees that the individuals identified in the schedule attached hereto as Exhibit C, entitled Key Personnel, is necessary for the successful completion of its services under the Contract. Upon the completion of any work package, the key personnel associated with such work package shall no longer be considered key personnel (except for purposes of Article 7 hereof), even if their names remain on Exhibit C, provided they are not to be assigned to subsequent work packages hereunder.

20.2 Such key personnel shall not be removed from the performance of the Contract unless replaced with personnel of substantially equal qualifications and ability. The Customer shall have the right to review the qualifications of any proposed replacements and, if for good and sufficient reasons, the Customer deems such personnel to be unsuitable, the Customer may require FhG to offer alternative candidates where such are available.

20.3 Notwithstanding the Customer's role in approving key personnel and their replacements, nothing in this Article shall relieve FhG of any of its obligations under this Contract or of its responsibility for any acts or omissions of its personnel.

21. REPORTS

21.1 FhG shall submit reports to the Customer consistent with the requirements contained in Exhibit A, Statement of Work. These reports will reflect the status of the activities by FhG and other information related to the project.

21.2 FhG will promptly inform the Customer about extraordinary circumstances arising during the performance of the services and about all matters under this Contract requiring the consent of the Customer.

21.3 FhG shall furnish to the Customer such information related to the services as the Customer may reasonably request from time to time. This information will be furnished on a non-interfering basis with the progress of the project.

22. AMENDMENTS TO THE CONTRACT

Any amendment or modification of this Contract, except in writing and signed by the authorized representatives of the Parties (in the case of the Customer, the Chief Executive Officer), shall be void and of no effect.

23. DOCUMENTS FORMING PART OF CONTRACT

The following documents shall be deemed to form and be read and construed as parts of this Contract, and shall hereinafter be called the Contract documents:

1) These Contractual Terms and Conditions
2) Statement of work (Exhibit A)
3) Exhibit B (Payment Plan)
4) Exhibit C (Key Personnel)
5) Exhibit D (FhG Background IP)
6) Exhibit E (Monthly Air Travel Report Form)

In the event of conflict or inconsistencies between this Contract and the Exhibits attached hereto, this Contract shall take precedence over such Exhibits and Appendices.

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24. INFORMATION AND ACCESS

24.1 FhG shall grant Customer reasonable access to all information performed under the contract and all work in progress.

24.2 The Customer shall ensure that FhG shall have the access that the Customer is given at all reasonable times to the facilities of all relevant spacecraft program contractors, launch vehicle program contractor and subcontractors, and that FhG shall have full access at all reasonable times to relevant data available from such contractors and subcontractors.

25. PERMITS AND AUTHORIZATIONS

FhG shall be responsible for all permits and authorizations required in Germany to perform the efforts defined in this contract. The Customer shall obtain all necessary US licenses.

26. NOTICE

26.1 Any notice, request, demand, approval, consent, or other communication ("Communication") permitted or required to be given by this Contract shall be effective only if in writing and delivered (I) personally, or (ii) by registered or certified mail, postage prepaid, return receipt requested, or (iii) by prepaid domestic courier, receipt acknowledged, or (iv) by facsimile or other electronic communications or similar conveyance, transmission confirmed, and addressed as follows:

If to FhG:

FhG Fraunhofer Institut für Integrierte Schaltungen
Am Weichselgarten 3,
D-91058 Erlangen, Germany
Attn: Stefan Meltzer
Contracts Manager
Telephone: +9131-776-6340
Facsimile: +9131-776-6399

If to Customer:

XM Satellite Radio Inc.
XM Innovation Center
600 West Hillsboro Blvd., Suite 210
Deerfield Beach, FL 33441
Attn: Dr. Stell Patsiokas
Senior Vice President, Technology
Telephone: (954) 419-9693
Facsimile: (954) 419-1694

26.2 If delivered personally, or by facsimile or their electronic conveyance, the deemed date of delivery shall be the date on which the Communication is dispatched. If delivered by mail or by courier, the deemed date of delivery shall be the date on which the Communication is received. All Communications shall bear the date on which the are dispatched or deposited in the mail. Either Party may change the address at which it will receive Communications upon the giving of notice to the other Party as provided above.

27. GENERAL

27.1 Except for assignment of delegation to a wholly-owned subsidiary of FhG or the Customer, neither Party shall assign or delegate this Contract or any of its rights, duties or obligations thereunder to any other person without the prior written consent of the other Party, which consent shall not be unreasonably withheld, provided that the assignor shall execute a guarantee in a form acceptable to the other Party which guarantees the due performance of the assignees and the observation of its duties and obligations under the Contract. Any attempt by either Party to assign or delegate any of its rights, duties or obligations under this Contract without such consent shall be void and of no effect.

27.2 FhG may subcontract portions of the work. Customer's written approval is required for any such subcontracts accounting for more than 300,000 -DM; or in the case of software, only for those subcontracts accounting for 150,000 -- DM or more. FhG shall provide sufficient detail on subcontractors experience, expertise and compliance with this Contract so that the Customer can adequately assess such subcontractor.

27.3 If either Party, at its option, agrees to a waiver of any of the terms and conditions recited herein, such waiver shall not for any purpose be construed as a waiver of any succeeding breach of the same or any other terms and conditions; not shall such a waiver be deemed as a course of conduct.

27.4 If any provision or clause, or portion thereof, of this Contract, or application thereof to any person or circumstances is held invalid or unconscionable, such invalidity or unconscionability shall not affect other provisions, or portions thereof, or applications of this Contract which can be given effect without the invalid or unconscionable provision, or portion thereof, or application, and to this end the provisions of these terms and conditions are declared to be severable.

27.5 Except as required to obtain necessary licenses or Governmental approvals, each Party shall give the other thirty (30) days advanced written notice to comment upon the content and timing of news releases, articles, brochures, advertisements, prepared speeches and other information releases, concerning this Contract or the work performed or to be performed hereunder.

27.6 Unless otherwise provided herein, any time limits to which this Contract binds FhG or the Customer shall be counted in calendar days from the day following that of the event marking the start of the time limit, and shall end of the last day of the period laid down. When the last day of a time limit is a Saturday or Sunday, or a recognized public holiday in the country in which the particular contractual performance is required, such time limit shall be extended to the first working day following.

27.7 This Contract shall be governed by the laws of Switzerland.

28. ON CALL SERVICES

The Customer may request FhG to provide other or additional related services beyond those described in Exhibit A, Statement of work or additional work upon the expiration of the Period of Performance as defined in Article 2. FhG will provide a quotation of labor, expenses and associated schedule in response to any such request on a time and materials basis. Upon written notification by the Customer of approval of the quotation, and formal incorporation into this Contract, and subject to Article 15 hereof, FhG shall provide the requested services.

29. ENTIRE AGREEMENT

This Contract constitutes the entire agreement between the Parties in connection with the subject matter hereof, and there are no other agreements or understandings, written or oral, except as provided herein.

IN WITNESS WHEREOF, the representatives of the parties hereto have executed this Contract, the present text of which shall be the only authentic version.

FhG

By: ________________
Name: Dr. Birgit Homma
Title: Head of R&D Contracts
Date: 10.09.1999

By: ________________
Name: Dr. Dirk Meints Polter
Title: Senior Vice President
Date: 13.09.1999

XM Satellite Radio Inc.

By: ________________
Name: John Wormington
Title: Sr. VP Engineering & Operations
Date: 10/13/99

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JS 44 (Rev. 12/12)

CIVIL COVER SHEET

The JS 44 civil cover sheet and the information contained herein neither replace nor supplement the filing and service of pleadings or other papers as required by law, except as provided by local rules of court. This form, approved by the Judicial Conference of the United States in September 1974, is required for the use of the Clerk of Court for the purpose of initiating the civil docket sheet. *(SEE INSTRUCTIONS ON NEXT PAGE OF THIS FORM.)*

I. (a) PLAINTIFFS

Fraunhofer-Gesellschaft Zur Förderung der angewandten Forschung e.V.

(b) County of Residence of First Listed Plaintiff

(EXCEPT IN U.S. PLAINTIFF CASES)

(c) Attorneys *(Firm Name, Address, and Telephone Number)*

Brian E. Farnan, Farnan LLP
919 N. Market Street, 12th Floor
Wilmington, DE 19801

DEFENDANTS

Sirius XM Radio Inc.

County of Residence of First Listed Defendant

(IN U.S. PLAINTIFF CASES ONLY)

NOTE: IN LAND CONDEMNATION CASES, USE THE LOCATION OF THE TRACT OF LAND INVOLVED.

Attorneys *(If Known)*

II. BASIS OF JURISDICTION *(Place an "X" in One Box Only)*

- ❒ 1 U.S. Government Plaintiff
- ☒ 3 Federal Question *(U.S. Government Not a Party)*
- ❒ 2 U.S. Government Defendant
- ❒ 4 Diversity *(Indicate Citizenship of Parties in Item III)*

III. CITIZENSHIP OF PRINCIPAL PARTIES *(Place an "X" in One Box for Plaintiff and One Box for Defendant)*

(For Diversity Cases Only)

	PTF	DEF		PTF	DEF
Citizen of This State	❒ 1	❒ 1	Incorporated *or* Principal Place of Business In This State	❒ 4	❒ 4
Citizen of Another State	❒ 2	❒ 2	Incorporated *and* Principal Place of Business In Another State	❒ 5	❒ 5
Citizen or Subject of a Foreign Country	❒ 3	❒ 3	Foreign Nation	❒ 6	❒ 6

IV. NATURE OF SUIT *(Place an "X" in One Box Only)*

CONTRACT
- ❒ 110 Insurance
- ❒ 120 Marine
- ❒ 130 Miller Act
- ❒ 140 Negotiable Instrument
- ❒ 150 Recovery of Overpayment & Enforcement of Judgment
- ❒ 151 Medicare Act
- ❒ 152 Recovery of Defaulted Student Loans (Excludes Veterans)
- ❒ 153 Recovery of Overpayment of Veteran's Benefits
- ❒ 160 Stockholders' Suits
- ❒ 190 Other Contract
- ❒ 195 Contract Product Liability
- ❒ 196 Franchise

REAL PROPERTY
- ❒ 210 Land Condemnation
- ❒ 220 Foreclosure
- ❒ 230 Rent Lease & Ejectment
- ❒ 240 Torts to Land
- ❒ 245 Tort Product Liability
- ❒ 290 All Other Real Property

TORTS

PERSONAL INJURY
- ❒ 310 Airplane
- ❒ 315 Airplane Product Liability
- ❒ 320 Assault, Libel & Slander
- ❒ 330 Federal Employers' Liability
- ❒ 340 Marine
- ❒ 345 Marine Product Liability
- ❒ 350 Motor Vehicle
- ❒ 355 Motor Vehicle Product Liability
- ❒ 360 Other Personal Injury
- ❒ 362 Personal Injury - Medical Malpractice

PERSONAL INJURY
- ❒ 365 Personal Injury - Product Liability
- ❒ 367 Health Care/ Pharmaceutical Personal Injury Product Liability
- ❒ 368 Asbestos Personal Injury Product Liability

PERSONAL PROPERTY
- ❒ 370 Other Fraud
- ❒ 371 Truth in Lending
- ❒ 380 Other Personal Property Damage
- ❒ 385 Property Damage Product Liability

CIVIL RIGHTS
- ❒ 440 Other Civil Rights
- ❒ 441 Voting
- ❒ 442 Employment
- ❒ 443 Housing/ Accommodations
- ❒ 445 Amer. w/Disabilities - Employment
- ❒ 446 Amer. w/Disabilities - Other
- ❒ 448 Education

PRISONER PETITIONS

Habeas Corpus:
- ❒ 463 Alien Detainee
- ❒ 510 Motions to Vacate Sentence
- ❒ 530 General
- ❒ 535 Death Penalty

Other:
- ❒ 540 Mandamus & Other
- ❒ 550 Civil Rights
- ❒ 555 Prison Condition
- ❒ 560 Civil Detainee - Conditions of Confinement

FORFEITURE/PENALTY
- ❒ 625 Drug Related Seizure of Property 21 USC 881
- ❒ 690 Other

LABOR
- ❒ 710 Fair Labor Standards Act
- ❒ 720 Labor/Management Relations
- ❒ 740 Railway Labor Act
- ❒ 751 Family and Medical Leave Act
- ❒ 790 Other Labor Litigation
- ❒ 791 Employee Retirement Income Security Act

IMMIGRATION
- ❒ 462 Naturalization Application
- ❒ 465 Other Immigration Actions

BANKRUPTCY
- ❒ 422 Appeal 28 USC 158
- ❒ 423 Withdrawal 28 USC 157

PROPERTY RIGHTS
- ❒ 820 Copyrights
- ☒ 830 Patent
- ❒ 840 Trademark

SOCIAL SECURITY
- ❒ 861 HIA (1395ff)
- ❒ 862 Black Lung (923)
- ❒ 863 DIWC/DIWW (405(g))
- ❒ 864 SSID Title XVI
- ❒ 865 RSI (405(g))

FEDERAL TAX SUITS
- ❒ 870 Taxes (U.S. Plaintiff or Defendant)
- ❒ 871 IRS—Third Party 26 USC 7609

OTHER STATUTES
- ❒ 375 False Claims Act
- ❒ 400 State Reapportionment
- ❒ 410 Antitrust
- ❒ 430 Banks and Banking
- ❒ 450 Commerce
- ❒ 460 Deportation
- ❒ 470 Racketeer Influenced and Corrupt Organizations
- ❒ 480 Consumer Credit
- ❒ 490 Cable/Sat TV
- ❒ 850 Securities/Commodities/ Exchange
- ❒ 890 Other Statutory Actions
- ❒ 891 Agricultural Acts
- ❒ 893 Environmental Matters
- ❒ 895 Freedom of Information Act
- ❒ 896 Arbitration
- ❒ 899 Administrative Procedure Act/Review or Appeal of Agency Decision
- ❒ 950 Constitutionality of State Statutes

V. ORIGIN *(Place an "X" in One Box Only)*

- ☒ 1 Original Proceeding
- ❒ 2 Removed from State Court
- ❒ 3 Remanded from Appellate Court
- ❒ 4 Reinstated or Reopened
- ❒ 5 Transferred from Another District *(specify)*
- ❒ 6 Multidistrict Litigation

VI. CAUSE OF ACTION

Cite the U.S. Civil Statute under which you are filing ***(Do not cite jurisdictional statutes unless diversity)***:
35 USC § 271

Brief description of cause:
Infringement of U.S. Patents

VII. REQUESTED IN COMPLAINT:

❒ CHECK IF THIS IS A **CLASS ACTION** UNDER RULE 23, F.R.Cv.P.

DEMAND $

CHECK YES only if demanded in complaint:
JURY DEMAND: ☒ Yes ❒ No

VIII. RELATED CASE(S) IF ANY

(See instructions): JUDGE DOCKET NUMBER

DATE
02/22/2017

SIGNATURE OF ATTORNEY OF RECORD
/s/ Brian E. Farnan

FOR OFFICE USE ONLY

RECEIPT # AMOUNT APPLYING IFP JUDGE MAG. JUDGE