

Fig. 11.13 Microtune single-chip tuner integrated onto the receiver's main PCB³

The advantages of this concept are the following: Tuners are small, lightweight, are integrated in the receiver board and are cost-effective in production (no manual placement and soldering of the tuner module). On the other hand the disadvantages in comparison to the more classical tuners are the following: the power dissipation is higher, the performance is slightly lower and the cost of the components is higher – but this list is from a snapshot from the year 2002, and time will tell.

11.4.3.3 Network Interface Module (NIM) Technology

One interesting technology for reception of DVB signals in general – be it DVB-S, DVB-C, or DVB-T- is the implementation of a so-called Network Interface Module (NIM). A NIM integrates in one module all signal processing from RF input to MPEG-2 transport stream output, containing the whole tuner functionality, the IF processing and the DVB-T decoder chip. The size of such a module is roughly double that of a standard tuner module. A NIM shows a number of advantages for the design of a DVB-T receiver:

- It has all the technology and the related know-how "on board", in one module. The receiver designer is therefore not forced to bother with RF problems like crosstalk etc.
- The RF and IF processing and also the PCB layout are optimally adapted to the integrated DVB-T decoder chip.
- NIMs from various manufactures can be interchanged. The control software needs to be adapted, which usually is rather simply done by re-

³ With kind permission of Microtune (TEMIC-Tuners)

268 11 The Standard for Terrestrial Transmission and Its Decoding Technique

designing the appropriate I²C bus routines, and a rather small PCB layout adaptation is typically required.

 Since NIMs are available for the different DVB transmission systems, namely satellite, cable, and terrestrial, they allow for the design of exactly the same receiver for all DVB systems.

Figure 11.14 shows a typical Network Interface Module (NIM).



Fig. 11.14 Network Interface Module (NIM) from Philips⁴ (top and bottom view)

Integrating a NIM into a receiver design is a good choice for PC-extension cards and for set-top boxes. If the performance is good enough they may be used for IDTVs. But for other receiver classes, NIMs have two main disadvantages:

- The size of a NIM is still quite large, so it may not be the perfect choice for small receivers like in PDAs or USB card extensions.
- It is not possible to control the RF- and IF-processing parameters, especially of the AGC from outside the NIM. These parameters, therefore, cannot be adapted or optimised for difficult reception conditions like the mobile channel. The NIM is therefore less suited for automotive receivers and PDAs
- The designer has to take the module "as it is". Exchanging parts, e.g. the DVB-T decoder or the SAW filter, for a part from a different supplier is not possible.

⁴ With kind permission of Philips Components, Business Unit Tuners.

APPENDIX UNDERBOECK01

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949 a Q350.12 v.34 JanJuly 1988 w LCPER c 1 i 000047859104 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 11/2 949 a Q350.12 v.33 July-Nov.1987 w LCPER c 1 i 000047859098 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 11/2 949 a Q350.12 v.33 July-Nov.1987 w LCPER c 1 i 000047859098 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 11/2 949 a Q350.12 v.33 JanMay 1987 w LCPER c 1 i 000047859074 d 4/17/2003 e 4/17/2003 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 11/2 949 a Q350.12 v.31 1985 w LCPER c 1 i 000047859067 d 11/22/2002 CATO-PARK m UP-ANNEX q 1 r Y s Y t PERIODICAL u 949 a Q350.12 v.31 1985 w LCPER c 1 i 000047859067 d 11/22/2002 CATO-PARK m UP-ANNEX q 1 r Y s Y t PERIODICAL u	а	Q350.I2 v.34 SeptNov.1988 w LCPER c 1 i 00004	7859111 CATO-PARK m UP	P-ANNEX r Y s Y t PERIODICAL u 11/28/2001
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949 a Q350.12 v.33 JanMay 1987 w LCPER c 1 I 000047859081 I CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 11/2 949 a Q350.12 v.32 1986 w LCPER c 1 i 000047859074 d 4/17/2003 e 4/17/2003 I CATO-PARK m UP-ANNEX n 1 r Y s Y 949 a Q350.12 v.32 1986 w LCPER c 1 i 000047859074 d 4/17/2003 e 4/17/2003 I CATO-PARK m UP-ANNEX n 1 r Y s Y 949 a Q350.12 v.31 1985 w LCPER c 1 i 000047859067 d 11/22/2002 I CATO-PARK m UP-ANNEX q 1 r Y s Y t PERIODICAL u 11/28/2001	a	Q350.12 v.33 July-Nov.1987 w LCPER c 1 i 000047	859098 CATO-PARK m UP-/	ANNEX r Y s Y t PERIODICAL u 11/28/2001
949 al Q350.12 V.32 1986 WI LCPEK ci 11 000047859074 di 4/17/2003 ei 4/17/2003 ii CATO-PARK mi UP-ANNEX ni 1 ri Y si Y PERIODICAL ui 11/28/2001 949 al Q350.12 V.31 1985 WI LCPER ci 1 ii 000047859067 di 11/22/2002 ii CATO-PARK mi UP-ANNEX qi 1 ri Y si Y ti PERIODICAL	а	Q350.12 V.33 JanMay 1987 w LCPER c 1 i 000047	859081 I CATO-PARK m UP-	ANNEX FI Y SI Y LI PERIODICAL UI 11/28/2001
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949 a Q350.12 v.30 1984 w LCPER c 1 i 000047859050 d 4/24/2003 l CATO-PARK m UP-ANNEX q 1 r Y s Y t PERIODIC 11/28/2001	a 11,	Q350.12 v.30 1984 w LCPER c 1 i 000047859050 d 28/2001	4/24/2003 CATO-PARK m	UP-ANNEX q 1 r Y s Y t PERIODICAL u

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949	al 222012 אינא נואא און גראבא גען דון מממטאיאסטאס מן אידיזעטט פן אידיזעטט ון גאוט-אאגע מן טא-אאואבא מן דון גען
	PERIODICAL u 11/28/2001
949	a Q350.l2 v.27 1981 w LCPER c 1 i 000047859029 d 4/27/2004 l CATO-PARK m UP-ANNEX q 1 r Y s Y t PERIODICAL u 11/28/2001
949	a Q350.12 v.51 Sept-Oct 2005 w LCPER c 1 i 000054949973 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 1/27/2006
949	al 0350.12 v.51 July-Aug 2005 w LCPER cl 1 il 000054949997 L CATO-PARK m LUP-ANNEX r LY SLY L PERIODICAL uI 1/27/2006
949	al O35012 v 51 Anr-lune 2005 w 1 CPER c1 11 000054949980 U CATO-PARK m 1 UP-ANNEX c1 X s1 X t1 PERIODICAL u 1/27/2006
949	a [0350.12 v 50 Oct-Dec 2004 w] CPER c1 11 000054823471 CATO-PARK m] P-ANNEX r1 Y 51 Y 11 PERIODICAL 2/18/2005
949	a 1 0350 12 v 50 lulv-sent 2004 w 1 CPER c1 1 i 000054823587 II CATO-PARK m 1 UP-ANNEX r1 X 51 X 1 PERIODICAL u 1 2/18/2005
9/9	al (250.12 v.50 appel 2004 w) LCEER c1 11 000054819200 d1 2/11/2005 al 2/11/2005 LCATO-PARK m) LIP-ANNEX n1 11/2 X s1
545	al (2501) and 200 Al 200 Al 201 and 200 Al 201 and 201
010	2 O250 12 / 50 120 May 2004 WELCOPP of 110 000054919217 JECATO PADK on LID ANNEY of Viel Viel PEDIODICAL 01 0/21/2004
949	a COSTA 2 V.S. Jahman 2004 W LEFER 111 0000529(621) CATO FARKIN OF ANNEXT 1 ST 1 FEDDICICAL U 22/1/2004
040	al Q3002 V.49 OLOBEZ 2003 WILCEER of 11 0000520000370 LICATO PARK MIT DEPARTER 1 ST CLEER CONCIDENT AL 222004
949	al Q550.2 v.49 JUJ-54 pt 2003 w LCPER ct 11 0000525400701 LCATO-PARK mL UP-ANNEX 1 35 1 Ct PERIODICAL 01 37/2004
949	al Q350.12 v.49 Aprojune 2003 w LCPER CT 11 0000525468611 CATO-PARK m UP-ANNEX TTS TT PERIODICAL 01 9/9/2003
949	a Q350.12 v.49 Jan-Mar 2003 w LCPEK c 1 000052548659 CATO-PARK M UP-ANNEX r Y S Y T PERIODICAL U 9/9/2003
949	a Q350.12 v.48 Oct-Dec 2002 w LCPER C 1 11 00005248/965 I CATO-PARK M UP-ANNEX T Y S Y T PERIODICAL U 5/30/2003
949	a Q350.12 v.48 July-Sept 2002 w LCPER c 1 i 000051778002 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 3/4/2003
949	a Q350.12 v.48 Apr-June 2002 w LCPER c 1 i 0000517779991 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 3/4/2003
949	a Q350.l2 v.48 Jan-Mar 2002 w LCPER c 1 i 000052486746 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 5/30/2003
949	a Q350.l2 v.47 July-Nov.2001 w LCPER c 1 i 000048278676 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 3/7/2002
949	a Q350.l2 v.47 JanMay 2001 w LCPER c 1 i 000048282604 l CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 3/14/2002
949	a Q350.I2 v.46 July-Nov.2000 w LCPER c 1 i 000048231756 d 11/9/2000 l CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 4/13/2001
949	al Q350 12 v 46 Jan - May 2000 w 11 CPER c1 1 il 000047850545 11 CATO-PARK m1 UP-ANNEX r1 X 51 X 11 PERIODICAL uI 11/28/2001
949	al 035012 v 45 Mav-Nov 1999 w 1 LCPER c1 1 i 000047850538 d1 11/11/2003 [L CATO-PARK m] LIP-ANNEX g1 1 L Y 5 Y 1
545	
949	al 035012 v 45 Jan -Anr 1999 w 11 CPER c1 1 il 000047850521 II CATO-PARK m 110-ANNEX c1 V 51 V 51 PERIODICAL ul 11/28/2001
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040	al Q35012 v 44 jan May 1350 w ECERCE 11 000047550041 II CATO PARK THE ANNEX FLY ST VET DEDICAL wITH/20/2001
040	a (2550 2 × 45) UV 407 W 1 CEEK (1 11) 00004750491 (CATO PARK M 11 O PARKAET 1 S 11) PERIODICAL U 11/26/2001
949	al Q350.12 v.45 Jahr-May 1997 W LCPER of 11 0000478504941 CATO-PARK mH 0F-ANNEX FT ST FT PERIODICAL 0T 1726/2001
949	a 055.12 v.42 JULY-100.1596 W LCPER (111) 000047504771 CATO-PARK MI OF-ANNEXT 15 11 PERIODICAL (11726/2001
949	al Q550.12 V.42 JanMay 1996 W LCPER C 111 0000478504601 CATO-PARK M UP-ANNEX T Y S Y T PERIODICAL UT 11/28/2001
949	a Q350.12 V.41 July-NoV.1995 W LCPER C 11 0000478304331 CATO-PARK M UP-ANNEX F Y S Y T PERIODICAL U 11/28/2001
949	a Q350.12 v.41 JanMay 1995 W LCPER c 1 11 0000478504461 C ATO-PARK M UP-ANNEX T Y S Y T PERIODICAL U 11/28/2001
949	a] Q50.12 V.40 JUJY-NOV.1994 W] LCPER C[11] 000047850439 0] 6/17/2004 I] CATO-PARK M] UP-ANNEX Q[17] YS] YT]
- 10	
949	al Q350.12 v.40 JanMay 1994 W LCPER cf 11 0000478504221 CATO-PARK M UP-ANNEX rf Y st PERIODICAL UT 11/28/2001
949	a Q 350.12 V.39 JUIV-NOV.1993 W LCPEK C 1 1 00004/850415 d 6/10/20031 CATO-PARK M UP-ANNEX d 1 1 Y S Y T
949	a Q350.12 v.39 JanMay 1993 w LCPER c 1 00004/8592031 CATO-PARK m UP-ANNEX r Y S Y T PERIODICAL u 11/28/2001
949	a Q350.12 v.51 Nov-Dec 2005 w LCPER c 1 i 000054950009 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 1/27/2006
949	a Q350.12 v.51 Jan 2005 w LCPER c 1 i 000055036528 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 9/22/2005
949	a Q350.l2 v.51 Feb 2005 w LCPER c 1 i 000055036511 l CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 9/22/2005
949	a Q350.l2 v.51 Mar 2005 w LCPER c 1 i 000055036429 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 9/22/2005
949	a Q350.12 v.52 Jan 2006 w LCPER c 1 i 000055132749 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 7/25/2006
949	a Q350.l2 v.52 Feb 2006 w LCPER c 1 i 000055132657 l CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 7/25/2006
949	a Q350.I2 v.52 Mar 2006 w LCPER c 1 i 000055132756 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 7/25/2006
949	a Q350.I2 v.52 Apr 2006 w LCPER c 1 i 000055132763 l CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 7/25/2006
949	a Q350.I2 v.52 May 2006 w LCPER c 1 i 000055132046 I CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 7/25/2006
949	a Q350.I2 v.52 June 2006 w LCPER c 1 i 000055132732 I CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 7/25/2006
949	a Q350.12 v.52 July 2006 w LCPER c 1 i 000061745452 l CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 11/16/2007
949	a Q350.I2 v.52 Aug 2006 w LCPER c 1 i 000061745469 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 11/16/2007
949	a Q350.l2 v.52 Sept 2006 w LCPER c 1 i 000061745483 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 11/16/2007
949	a Q350.12 v.52 Oct 2006 w LCPER c 1 i 000061745490 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 11/16/2007
949	a Q350.I2 v.52 Nov 2006 w LCPER c 1 i 000061745445 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 11/16/2007
949	a Q350.12 v.52 Dec 2006 w LCPER c 1 i 000061745476 CATO-PARK m UP-ANNEX r Y s Y t PERIODICAL u 11/16/2007
949	a Q350.12 v.PGIT2-PGIT4 1953/54 w LCPER c 1 i 000025183719 I ACAD-BLDG m UP-ANNEX r Y s Y t PERIODICAL u
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349	מן עססטגב אווס-אוויד וססיוסס אן בכרבת כן דון טטטטסיזעסססע מן סיסיצטוע ון אכאט-אבעט מון טא-אממצע מן דרן זיגן צו גערבת כן דון טעטטסיזעססט מן געראט-אבעט מון טא-אממצע מן געראט-אבעט געראט אין געראט-אבעט מן געראט אין געראט אין ג
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949	aj Q350.12 v.115 v.116 1959/60 wj LCPER CJ 1 IJ 000052241635 IJ ACAD-BLDG mj UP-ANNEX FJ Y SJ Y LJ PERIODICAL U 2/18/2010
949	a Q350.12 v.IT7-v.IT8 1961/62 w LCPER c 1 000067429349 ACAD-BLDG m UP-ANNEX r Y s Y t PERIODICAL u 2/18/2010
949	a Q350.12 v.IT9-v.IT10 1963/64 w LCPER c 1 i 000067429257 l ACAD-BLDG m UP-ANNEX r Y s Y t PERIODICAL u 2/18/2010
949	a Q350.I2 v.IT11-v.IT12 1965/66 w LCPER c 1 i 000067429165 l ACAD-BLDG m UP-ANNEX r Y s Y t PERIODICAL u 2/18/2010
949	a Q350.l2 v.lT13 1967 w LCPER c 1 i 000067429158 l ACAD-BLDG m UP-ANNEX r Y s Y t PERIODICAL u 2/18/2010
949	al O350.I2 v.IT14 1968 wi LCPER ci 1 il 000067429141 ll ACAD-BLDG mi UP-ANNEX ri Y si Y ti PERIODICAL ul 2/18/2010

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949		a	Q350.I2 v.IT15 1969 w LCPER c 1 i 000067429240 I ACAD-BLDG m UP-ANNEX r Y s Y t PERIODICAL u 2/18/2010
949		a	Q350.12 v.IT16 1970 w LCPER c 1 i 000067429233 ACAD-BLDG m UP-ANNEX r Y s Y t PERIODICAL u 2/18/2010
949		a	Q350.12 v.IT17 1971 w LCPER c 1 i 000067429226 l ACAD-BLDG m UP-ANNEX r Y s Y t PERIODICAL u 2/18/2010
949		a	Q350.12 v.IT18 1972 w LCPER c 1 i 000067429219 ACAD-BLDG m UP-ANNEX r Y s Y t PERIODICAL u 2/18/2010
949		a	Q350.12 v.IT19 1973 w LCPER c 1 i 000067429271 ACAD-BLDG m UP-ANNEX r Y s Y t PERIODICAL u 2/18/2010
949		a	Q350.12 v.IT20 1974 w LCPER c 1 i 000067429264 ACAD-BLDG m UP-ANNEX r Y s Y t PERIODICAL u 2/18/2010
949		a	Q350.12 v.IT21 1975 w LCPER c 1 i 000067429288 l ACAD-BLDG m UP-ANNEX r Y s Y t PERIODICAL u 2/18/2010
949		a	Q350.12 v.IT22-v.IT23 1976/77 w LCPER c 1 i 000067429295 d 10/31/2013 ACAD-BLDG m UP-ANNEX q 1 r Y s Y t
		PEF	RIODICAL u 2/18/2010
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CONTENTS

	Pr	eface	xiz				
1	In	Introduction					
	1.1	Elements of a Digital Communication C	1				
	1.2	Communication Channels and Their Cl	1				
	1.3	Mathematical Models for Communication Classics	3				
	1.4	A Historical Perspective in the David	10				
	1.5	Overview of the Book	13				
	1.6	Bibliographical Notes and Beforenees	15				
		Biolographical Notes and References	16				
2	Pre	obability and Stochastic Processes	17				
	2.1	Probability	1/				
		2.1.1 Random Variables, Probability Distributions, and Probability Densities 2.1.2 Functions of Random Variables 2.1.3 Statistical Averages of Random Variables 2.1.4 Some Useful Buckshilli	17				
		Distributions 2.1.5 Upper Bounds on the Tail Probability					
		Sums of Random Variables and the Central Limit Theorem					
	2.2	Stochastic Processes	61				
		2.2.1 Statistical Averages 2.2.2 Power Density Spectrum 2.2.3					
		Response of a Linear Time-Invariant System to a Random Input					
		Signal 2.2.4 Sampling Theorem for Band-Limited Stochastic					
		Processes 2.2.5 Discrete-Time Stochastic Signals and Systems 2.2.6 Cyclostationary Processes					
	2.3	Bibliographical Notes and References	75				
		Problems	15				
			15				
;	Sou	rce Coding	80				
	3.1	Mathematical Models for Information Sources	00				
	3.2	A Logarithmic Measure of Information	80				
		3.2.1 Average Mutual Information and Entropy 3.2.2 Information Measures for Continuous Random Variables	82				
	3.3	Coding for Discrete Sources	00				
		3.3.1 Coding for Discrete Memoryless Sources 3 3 2 Discrete	90				
		Stationary Sources / 3.3.3 The Lempel–Ziv Algorithm					

3

	٠		
v	1		
л	1	Ŀ	

and a

		C	ontent
	3.4	Coding for Analog Sources—Optimum Quantization 3.4.1 Rate-Distortion Function 3.4.2 Scalar Quantization 3.4.3 Vector Quantization	103
	3.5	Coding Techniques for Analog Sources 3.5.1 Temporal Waveform Coding 3.5.2 Spectral Waveform Coding 3.5.3 Model-Based Source Coding	121
	3.6	Bibliographical Notes and References Problems	140 141
ŀ	Ch	aracterization of Communication Signals and Systems	148
	4.1	Representation of Band-Pass Signals and Systems 4.1.1 Representation of Band-Pass Signals 4.1.2 Representation of Linear Band-Pass Systems 4.1.3 Response of a Band-Pass System to a Band-Pass Signal 4.1.4 Representation of Band-Pass Stationary Stochastic Processes	148
	4.2	Signal Space Representations 4.2.1 Vector Space Concepts 4.2.2 Signal Space Concepts 4.2.3 Orthogonal Expansions of Signals	158
	4.3	Representation of Digitally Modulated Signals 4.3.1 Memoryless Modulation Methods 4.3.2 Linear Modulation with Memory 4.3.3 Non-linear Modulation Methods with Memory— CPFSK and CPM	168 1
	4.4	Spectral Characteristics of Digitally Modulated Signals 4.4.1 Power Spectra of Linearly Modulated Signals 4.4.2 Power Spectra of CPFSK and CPM Signals 4.4.3 Power Spectra of Modulated Signals with Memory	201
	4.5	Bibliographical Notes and References Problems	221 222
	Opt Cha	timum Receivers for the Additive White Gaussian Noise	221
	5.1	Optimum Receiver for Signals Corrupted by Additive White Gaussian Noise	231
		5.1.1 Correlation Demodulator 5.1.2 Matched-Filter Demodulator 5.1.3 The Optimum Detector 5.1.4 The Maximum-Likelihood Sequence Detector 5.1.5 A Symbol-by-Symbol MAP Detector for Signals with Memory	201
	5.2	Performance of the Optimum Receiver for Memoryless Modulation 5.2.1 Probability of Error for Binary Modulation 5.2.2 Probability of Error for M-ary Orthogonal Signals 5.2.3 Probability of Error for M-ary Biorthogonal Signals 5.2.4 Probability of Error for Simplex Signals 5.2.5 Probability of Error for M-ary Binary-Coded Signals 5.2.6 Probability of Error for M-ary PAM 5.2.7 Probability of Error for M-ary PSK 5.2.8 Differential PSK (DPSK)	254
		J. J	

	and Its Performance 5.2.9 Probability of Error for QAM 5.2.10	
	Comparison of Digital Modulation Methods	
5.3	Optimum Receiver for CPM Signals	283
	5.3.1 Optimum Demodulation and Detection of CPM 5.3.2	
	Performance of CPM Signals 5.3.3 Symbol-by-Symbol Detection of	
	CPM Signals 5.3.4 Suboptimum Demodulation and Detection of	
= 1	CPM Signals	
5.4	Optimum Receiver for Signals with Random Phase in AWGN Channel	300
	5.4.1 Optimum Receiver for Binary Signals 5.4.2 Optimum Receiver	
	Detection of M any Orthogonal Signals 5.4.5 Probability of Error for Envelope	
	Envelope Detection of Correlated Pinary Signals	
55	Performance Analysis for Wireling and Padia Communication S	
0.0	5.5.1 Regenerative Repeaters 5.5.2 Link Budget Analysis in Dedia	313
	Communication Systems	
5.6	Bibliographical Notes and References	210
	Problems	310
		515
Ca	rrier and Symbol Synchronziation	222
61	Signal Parameter Estimation	335
0.1	611 The Likelihood Function 612 Causian Baseness of Santa	333
	Synchronization in Signal Demodulation	
6.2	Carrier Phase Estimation	220
	6.2.1 Maximum-Likelihood Carrier Phase Estimation 1.6.2.2 The	338
	Phase-Locked Loon 6.2.3 Effect of Additive Noise on the Phase	
	Estimate 6.2.4 Decision-Directed Loops 6.2.5 Non-Decision	
	Directed Loops	
6.3	Symbol Timing Estimation	350
	6.3.1 Maximum-Likelihood Timing Estimation 6.3.2 Non-Decision-	557
	Directed Timing Estimation	
6.4	Joint Estimation of Carrier Phase and Symbol Timing	366
6.5	Performance Characteristics of ML Estimators	368
6.6	Bibliographical Notes and References	371
	Problems	372
CL		
Cha	annel Capacity and Coding	376
7.1	Channel Models and Channel Capacity	376
	7.1.1 Channel Models 7.1.2 Channel Capacity 7.1.3 Achieving	
	Channel Capacity with Orthogonal Signals 7.1.4 Channel Reliability	
= -	Functions	
7.2	Random Selection of Codes	392
	Participation Coding Based on M-ary Binary-Coded Signals 7.2.2	
	Comparison of P* with the C	
72	Comparison of K_0 with the Capacity of the AWGN Channel	
1.5	Communication System Design Based on the Cutoff Rate	402

			Contents
	7.4	4 Bibliographical Notes and References	408
		Problems	400
		I something the second se	
5	8 BI	ock and Convolutional Channel Codes	
	81	Linear Black Co. 1	416
	0.1	8 1 The Generator Matrix and d. D. i. Cl. 1 M.	416
		Specific Linear Block Codes 8.1.3 Cyclic Codes 8.1.4 Optimum Soft-Decision Decoding of Linear Block Codes 8.1.5 Hard-Decision Decoding of Linear Block Codes 8.1.6 Comparison of Performance Between Hard-Decision and Soft-Decision Decoding 8.1.7 Bounds on Minimum Distance of Linear Block Codes 6.1.7 Bounds on	ne
		Codes and Concatenated Block Codes 8.1.8 Nonbinary Block Data for Channels with Burst Errors 8.1.10 Serial and Parallel Concatenated Block Codes	
	8.2	Convolutional Codes 8.2.1 The Transfer Function of a Convolutional Code 8.2.2 Optimum Decoding of Convolutional Codes—The Viterbi Algorithm 8.2.3 Probability of Free for Code Decider	471
		Error for Hard-Decision Decoding 8.2.4 Probability of Error for Hard-Decision Decoding 8.2.5 Distance Properties of Binary Convolutional Codes 8.2.6 Punctured Convolutional Codes 8.2.7 Other Decoding Algorithms for Convolutional Codes 8.2.8	
	0.2	Practical Considerations in the Application of Convolutional Codes / 8.2.9 Nonbinary Dual-k Codes and Concatenated Codes / 8.2.10 Parallel and Serial Concatenated Convolutional Codes	
	8.3	Coded Modulation for Bandwidth-Constrained Channels-Trellis-Coo	ded
	8.4	Bibliographical Notes and Pafaranasa	522
		Problems	539
			541
9	Sig	al Design for Rond Limited Channel	
	0 1	Characterization of D 111 is a constant of the second	548
	9.2	Signal Design for Band Limited Channels	548
		9.2.1 Design of Band-Limited Signals for No Intersound of	554
		Interference—The Nyquist Criterion 9.2.2 Design of Rand-Limited	
		Signals with Controlled ISI—Partial-Response Signals 9.2.3 Data Detection for Controlled ISI 9.2.4 Signal Design for Channels with Distortion	
	9.3	Probability of Error in Detection of PAM	574
		9.3.1 Probability of Error for Detection of PAM with Zero ISI 9.3.2 Probability of Error for Detection of Partial-Response Signals	5/4
	9.4	Modulation Codes for Spectrum Shaping	578
	9.5	Bibliographical Notes and References	588
		1 TODICIIIS	588

xiv

10	Con	imunication Through Band-Limited Linear Filter Channels	598
	10.1	Optimum Receiver for Channels with ISI and AWGN 10.1.1 Optimum Maximum-Likelihood Receiver 10.1.2 A Discrete- Time Model for a Channel with ISI 10.1.3 The Viterbi Algorithm for the Discrete-Time White Noise Filter Model 10.1.4 Performance of MLSE for Channels with ISI	599
	10.2	Linear Equalization 10.2.1 Peak Distortion Criterion / 10.2.2 Mean-Square-Error (MSE)	616
		Criterion 10.2.3 Performance Characteristics of the MSE Equalizer 10.2.4 Fractionally Spaced Equalizers 10.2.5 Baseband and Passband Linear Equalizers	
	10.3	Decision-Feedback Equalization 10.3.1 Coefficient Optimization 10.3.2 Performance Characteristics of DFE 10.3.3 Predictive Decision-Feedback Equalizer 10.3.4 Equalization at the Transmitter—Tomlinson–Harashima Precoding	638
	10.4	Reduced Complexity ML Detectors	647
	10.5	Iterative Equalization and Decoding-Turbo Equalization	649
	10.6	Bibliographical Notes and References	651
		Problems	652
11	Adaj	ptive Equalization	660
	11.1	Adaptive Linear Equalizer	660
		11.1.1 The Zero-Forcing Algorithm 11.1.2 The LMS Algorithm 11.1.3 Convergence Properties of the LMS Algorithm 11.1.4 Excess	000
		MSE Due to Noisy Gradient Estimates 11.1.5 Accelerating the Initial Convergence Rate in the LMS Algorithm 11.1.6 Adaptive Fractionally Spaced Equalizer—The Tap Leakage Algorithm 11.1.7 An Adaptive Channel Estimator for ML Sequence Detection	
	11.2	Adaptive Decision-Feedback Equalizer	677
	11.3	Adaptive Equalization of Trellis-Coded Signals	678
	11.4	Recursive Least-Squares Algorithms for Adaptive Equalization 11.4.1 Recursive Least-Squares (Kalman) Algorithm 11.4.2 Linear Prediction and the Lattice Filter	682
	11.5	Self-Recovering (Blind) Equalization 11.5.1 Blind Equalization Based on the Maximum-Likelihood Criterion 11.5.2 Stochastic Gradient Algorithms 11.5.3 Blind Equalization Algorithms Based on Second- and Higher-Order Signal Statistic	693
	11.6	Statistics Bibliographical Notes and References	704
		Problems	705
2	Mult	ichannel and Multicarrier Systems	709
	12.1	Multichannel Digital Communications in AWGN Channels	700
		12.1.1 Binary Signals / 12.1.2 M-ary Orthogonal Signals	/09

XV

	12.2	Multicarrier Communications 12.2.1 Capacity of a Nonideal Linear Filter Channel 12.2.2 An	715
		FFT-Based Multicarrier System / 12.2.3 Minimizing Peak-to-Average Ratio in the Multicarrier Systems	
	12.3	Bibliographical Notes and References	723
		Problems	724
13	Spre	ad Spectrum Signals for Digital Communications	726
	13.1	Model of Spread Spectrum Digital Communication System	728
	13.2	Direct Sequence Spread Spectrum Signals 13.2.1 Error Rate Performance of the Decoder 13.2.2 Some Applications of DS Spread Spectrum Signals 13.2.3 Effect of Pulsed Interference on DS Spread Spectrum Systems 13.2.4 Excision of Narrowband Interference in DS Spread Spectrum Systems 13.2.5 Generation of PN Sequences	729
	13.3	Frequency-Hopped Spread Spectrum Signals 13.3.1 Performance of FH Spread Spectrum Signals in an AWGN Channel 13.3.2 Performance of FH Spread Spectrum Signals in Partial-Band Interference 13.3.3 A CDMA System Based on FH Spread Spectrum Signals	771
	13.4	Other Types of Spread Spectrum Signals	784
	13.5	Synchronization of Spread Spectrum Systems	786
	13.6	Bibliographical Notes and References	792
		Problems	794
14	Digi	tal Communications through Fading Multipath Channels	800
	14.1	Characterization of Fading Multipath Channels 14.1.1 Channel Correlation Functions and Power Spectra 14.1.2 Statistical Models for Fading Channels	801
	14.2	The Effect of Signal Characteristics on the Choice of a Channel Mode	1814
	14.3	Frequency-Nonselective, Slowly Fading Channel	816
	14.4	Diversity Techniques for Fading Multipath Channels 14.4.1 Binary Signals 14.4.2 Multiphase Signals 14.4.3 M-ary Orthogonal Signals	821
	14.5	Digital Signaling over a Frequency-Selective, Slowly Fading Channel 14.5.1 A Tapped-Delay-Line Channel Model 14.5.2 The RAKE Demodulator 14.5.3 Performance of RAKE Demodulator 14.5.4 Receiver Structures for Channels with Intersymbol Interference	840
	14.6	Coded Waveforms for Fading Channels 14.6.1 Probability of Error for Soft-Decision Decoding of Linear Binary Block Codes 14.6.2 Probability of Error for Hard-Decision Decoding of Linear Binary Block Codes 14.6.3 Upper Bounds on the Performance of Convolutional Codes for a Rayleigh Fading	852
		Channel 14.6.4 Use of Constant-Weight Codes and Concatenated Codes for a Fading Channel 14.6.5 System Design Based on the	

xvi

	Cu	toff F	Rate 14.6.6 Performance of Coded Phase-Coherent	
	Co	mmur	nication Systems—Bit-Interleaved Coded Modulation 14.6.7	7
14.7	Ire	ellis-C	oded Modulation	
14./	Dit		-Antenna Systems	878
14.0	BIC	hlom	aphical Notes and References	885
	PIC	oblem	s	887
15 Mul	tiuse	er Co	ommunications	896
15.1	Int	roduc	tion to Multiple Access Techniques	896
15.2	Caj	pacity	of Multiple Access Methods	899
15.3	Co	de-Di	vision Multiple Access	905
	15.	3.1 C	DMA Signal and Channel Models 15.3.2 The Optimum	
	Rea	ceiver	15.3.3 Suboptimum Detectors 15.3.4 Successive	
	Inte	erfere	nce Cancellation / 15.3.5 Performance Characteristics of	
	Dei	ector.	S	
15.4	Rai	ndom	Access Methods	922
	Svs	4.1 A. tems	LOHA Systems and Protocols / 15.4.2 Carrier Sense and Protocols	
15.5	Bib	liogra	phical Notes and References	031
	Pro	blems	description of the second second second second second	933
Annondix		The	Lovingon Dunkin Alassid	
Appendix	A	The	Levinson–Durbin Algorithm	939
Appendix	B	Erre	or Probability for Multichannel Binary Signals	943
Appendix	С	Erre	or Probabilities for Adaptive Reception of <i>M</i> -Phas	e
		Sigi	lais	949
		C.1	Mathematical Model for M-Phase Signaling Communicati	on
		C 2	System	949
		C.2	Characteristic Function and Probability Density Function	of
		C3	Error Probabilities for Slowly Dealed I. D. V. Cl.	952
		C.4	Error Probabilities for Time Inversiont and Diver E. J.	953
		0.4	Channels	0.57
				956
Appendix	D	Squ	are-Root Factorization	961
References	and	Bibl	liography	963
Lada				
Index				993

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Turbo-coded APSK modulations design for satellite broadband communications

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SUMMARY

This paper investigates the design of power and spectrally efficient coded modulations based on amplitude phase shift keying (APSK) modulation with application to satellite broadband communications. APSK represents an attractive modulation format for digital transmission over nonlinear satellite channels due to its power and spectral efficiency combined with its inherent robustness against nonlinear distortion. For these reasons APSK has been very recently introduced in the new standard for satellite Digital Video Broadcasting named DVB-S2. Assuming an ideal rectangular transmission pulse, for which no nonlinear inter-symbol interference is present and perfect pre-compensation of the nonlinearity, we optimize the APSK constellation. In addition to the minimum distance criterion, we introduce a new optimization based on the mutual information; this new method generates an optimum constellation for each spectral efficiency. To achieve power efficiency jointly with low bit error rate (BER) floor we adopt a powerful binary serially concatenated turbo-code coupled with optimal APSK modulations through bit-interleaved coded modulation. We derive tight approximations on the maximum-likelihood decoding error probability, and results are compared with computer simulations. The proposed coded modulation scheme is shown to provide a considerable performance advantage compared to current standards for satellite multimedia and broadcasting systems. Copyright © 2006 John Wiley & Sons, Ltd.

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KEY WORDS: turbo codes; amplitude-phase shift keying (APSK); modulation; bit-interleaved coded modulation (BICM); coded modulation; nonlinear channels; satellite communications

1. INTRODUCTION

A major strength of satellite communications systems lies on their ability to efficiently broadcast digital multi-media information over very large areas [1]. A notable example is the

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so-called direct-to-home (DTH) digital television broadcasting. Satellite systems also provide a unique way to complement the terrestrial telecommunication infrastructure in scarcely populated regions. The introduction of multi-beam satellite antennas with adaptive coding and modulation (ACM) schemes will allow an important efficiency increase for satellite systems operating at Ku or Ka-band [2]. Those technical enhancements require the exploitation of power- and spectrally efficient modulation schemes conceived to operate over the satellite nonlinear channel. In this paper, we will design high-efficiency 16- and 32-ary coded modulation schemes suited for nonlinear satellite channels. The analysis presented here is complemented in [3] with the effects related to satellite nonlinear distortion, band-limited transmission pulse, demodulator timing, amplitude and phase estimation errors.

To the authors' knowledge there are few references in the literature dealing with 16-ary constellation optimization over nonlinear channels, the typical environment for satellite channels. Previous work showed that 16-QAM does not compare favourably with either trellis-coded (TC) 16-PSK or uncoded 8-PSK in satellite nonlinear channels [4]. The concept of circular APSK modulation was already proposed 30 years ago by Thomas et al. [5], where several nonband-limited APSK sets were analysed by means of uncoded bit error rate bounds; the suitability of APSK for nonlinear channels was also made explicit, but concluded that for single carrier operation over nonlinear channel APSK performs worse than PSK schemes. In the current paper, we will revert the conclusion. It should be remarked that Reference [5] mentioned the possibility of modulator pre-compensation but did not provide performance results related to this technique. Foschini et al. [6] optimized QAM constellations using asymptotic uncoded probability of error under average power constraints, deriving optimal 16-ary constellation made of an almost equilateral lattice of triangles. This result is not applicable to satellite channels. In Reference [7] some comparison between squared QAM and circular APSK over linear channels was performed based on the computation of the error bound parameter, showing some minor potential advantage of APSK. Further work on mutual information for modulations with average and peak power constraints is reported in Reference [8], which proves the advantages of circular APSK constellations under those power constraints. Mutual information performance loss for APSK in peak power limited Gaussian complex channels is reported in Reference [9] and compared to classical QAM modulations; it is shown that under this assumption APSK considerably outperforms QAM in terms of mutual information, the gain particularly remarkable for 16- and 64-ary constellations.

Forward error correcting codes for our application must combine power efficiency and low BER floor with flexibility and simplicity to allow for high-speed implementation. The existence of practical, simple, and powerful such coding designs for binary modulations has been settled with the advent of turbo codes [10] and the recent re-discovery of low-density parity-check (LDPC) codes [11]. In parallel, the field of channel coding for nonbinary modulations has evolved significantly in the latest years. Starting with Ungerboeck's work on TC modulation (TCM) [12], the approach had been to consider channel code and modulation as a single entity, to be jointly designed and demodulated/decoded. Schemes have been published in the literature, where turbo codes are successfully merged with TCM [13]. Nevertheless, the elegance and simplicity of Ungerboeck's original approach gets somewhat lost in a series of *ad hoc* adaptations; in addition, the turbo-code should be jointly designed with a given modulation, a solution impractical for system supporting several constellations. A new pragmatic paradigm has crystallized under the name of bit-interleaved coded modulation (BICM) [13], where extremely good results are obtained with a standard nonoptimized, code. An additional

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advantage of BICM is its inherent flexibility, as a single mother code can be used for several modulations, an appealing feature for broadband satellite communication systems where a large set of spectral efficiencies is needed.

This paper is organized as follows. Section 2 gives the system model under the ideal case of a rectangular transmission pulse.[‡] Section 3 gives a formal description of APSK signal sets, describes the maximum mutual information and maximum minimum distance optimization criteria and discusses some of the properties of the optimized constellations. Section 4 deals with code design issues, describes the BICM approach, provides some analytical considerations based on approximate maximum-likelihood (ML) decoding error probability bounds, and provides some numerical results. The conclusions are finally drawn in Section 5.

2. SYSTEM MODEL

The baseband equivalent of the transmitted signal at time t, $s_T(t)$, is given by

$$s_T(t) = \sqrt{P} \sum_{k=0}^{L-1} x(k) p_T(t - kT_s)$$
(1)

where *P* is the signal power, x(k) is the *k*th transmitted symbol, drawn from a complex-valued APSK signal constellation \mathscr{X} , with $|\mathscr{X}| = M$, p_T is the transmission filter impulse response, and T_s is the symbol duration (in seconds), corresponding to one channel use. Without loss of generality, we consider transmission of frames with *L* symbols. The spectral efficiency *R* is defined as the number of information bits conveyed at every channel use, and in measured in bits per second per Hertz (bps/Hz).

The signal $s_T(t)$ passes through a high-power amplifier (HPA) operated close to the saturation point. In this region, the HPA shows nonlinear characteristics that induce phase and amplitude distortions to the transmitted signal. The amplifier is modelled by a memoryless nonlinearity, with an output signal $s_A(t)$ at time t given by

$$s_A(t) = F(|s_T(t)|) e^{j(\phi(s_T(t)) + \Phi(|s_T(t)|))}$$
(2)

where we have implicitly defined F(A) and $\Phi(A)$ as the AM/AM and AM/PM characteristics of the amplifier for a signal with instantaneous signal amplitude A. The signal amplitude is the instantaneous complex envelope, so that the baseband signal is decomposed as $s_T(t) = |s_T(t)|e^{j\phi(s_T(t))}$.

In this paper, we assume an (ideal) signal modulating a train of rectangular pulses. These pulses do not create inter-symbol interference when passed through an amplifier operated in the nonlinear region. Under these conditions, the channel reduces to an AWGN, where the modulation symbols are distorted following (2). Let x_A denote the distorted symbol corresponding to $x = |x|e^{j\phi(x)} \in \mathcal{X}$, that is, $x_A = F(|x|)e^{j(\phi(x)+\Phi(|x|))}$. After matched filtering and sampling at time kT_s , the discrete-time received signal at time k, y(k) is then given by

$$y(k) = \sqrt{E_s x_A(k) + n(k)}, \quad k = 0, \dots, L - 1$$
 (3)

[‡]This assumption has been dropped in the paper [14].

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with E_s the symbol energy, given by $E_s = PT_s$, $x_A(k)$ is the symbol at the kth time instant, as defined above, and $n(k) \sim \mathcal{N}_{\mathbb{C}}(0, N_0)$ is the corresponding noise sample.

This simplified model suffices to describe the nonlinearity up to the nonlinear ISI effect, and allows us to easily design constellation and codes. In the paper [14], the impact of nonlinear ISI has been considered, as well as other realistic demodulation effects such as timing and phase recovery.

3. APSK CONSTELLATION DESIGN

In this section, we define the generic multiple-ring APSK constellation family. We propose new criteria for the design of digital QAM constellations of 16 and 32 points, with special emphasis on the behaviour on nonlinear channels.

3.1. Constellation description

M-APSK constellations are composed of n_R concentric rings, each with uniformly spaced PSK points. The signal constellation points x are complex numbers, drawn from a set \mathscr{X} given by

$$\mathscr{X} = \begin{cases} r_1 e^{\mathbf{j}((2\pi/n_1)i+\theta_1)}, & i = 0, \dots, n_1 - 1 \quad (\text{ring 1}) \\ r_2 e^{\mathbf{j}((2\pi/n_2)i+\theta_2)}, & i = 0, \dots, n_2 - 1 \quad (\text{ring 2}) \\ \vdots \\ r_{n_R} e^{\mathbf{j}((2\pi/n_R)i+\theta_{n_R})}, & i = 0, \dots, n_{n_R} - 1 \quad (\text{ring } n_R) \end{cases}$$
(4)

where we have defined n_{ℓ} , r_{ℓ} and θ_{ℓ} as the number of points, the radius and the relative phase shift for the ℓ th ring. We will nickname such modulations as $n_1 + \cdots + n_{n_R}$ -APSK. Figure 1 depicts the 4 + 12- and 4 + 12 + 16-APSK modulations with quasi-Gray mapping. In particular, for next generation broadband systems [2, 15], the constellation sizes of interest are $|\mathscr{X}| = 16$ and 32, with $n_R = 2$ and 3 rings, respectively. In general, we consider that \mathscr{X} is normalized in energy, i.e. $E[|x|^2] = 1$, which implies that the radii r_{ℓ} are normalized such that $\sum_{\ell=1}^{n_R} n_{\ell} r_{\ell}^2 = 1$. Notice also that the radii r_{ℓ} are ordered, so that $r_1 < \cdots < r_{n_R}$.

Clearly, we can also define the phase shifts and the ring radii in relative terms rather than in absolute terms, as in (4); this removes one dimension in the optimization process, yielding a practical advantage. We let $\phi_{\ell} = \theta_{\ell} - \theta_1$ for $\ell = 1, ..., n_R$ be the phase shift of the ℓ th ring with respect to the inner ring. We also define $\rho_{\ell} = r_{\ell}/r_1$ for $\ell = 1, ..., n_R$ as the relative radii of the ℓ th ring with respect to r_1 . In particular, $\phi_1 = 0$ and $\rho_1 = 1$.

3.2. Constellation optimization in AWGN

We are interested in finding an APSK constellation, defined by the parameters $\mathbf{\rho} = (\rho_1, \dots, \rho_{n_R})$ and $\mathbf{\phi} = (\phi_1, \dots, \phi_{n_R})$, such that a given cost function $f(\mathcal{X})$ reaches a minimum. The simplest, and probably most natural, cost function is the minimum Euclidean distance between any two points in the constellation. Section 3.2.1 shows the results under this criterion. These results are extended in Section 3.2.2, where the cost function is replaced by the mutual information of the AWGN channel; it also shown that significant gains may be achieved for low and moderate values of signal-to-noise ratio (SNR) by fine-tuning the constellation.

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Figure 1. Parametric description and pseudo-Gray mapping of 16 and 32-APSK constellations with $n_1 = 4, n_2 = 12, \phi_2 = 0$ and $n_1 = 4, n_2 = 12, n_3 = 16, \phi_2 = 0, \phi_3 = \pi/16$. For the first two rings: mapping below corresponds to 4 + 12-APSK, mapping above to 4 + 12 + 16-APSK.

3.2.1. Minimum Euclidean distance maximization. The union bound on the uncoded symbol error probability [16] yields,

$$P_e \leqslant \frac{1}{M} \sum_{x \in \mathscr{X}} \sum_{\substack{x' \in \mathscr{X} \\ x' \neq x}} Q\left(\sqrt{\frac{E_s |x - x'|^2}{2N_0}}\right)$$
(5)

where $Q(x) = 1/\sqrt{2\pi} \int_x^{\infty} e^{-(t^2/2)} dt$ is the Gaussian tail function. At high SNR Equation (5) is dominated by the pairwise term at minimum squared Euclidean distance $\delta_{\min}^2 = \min_{x,x' \in \mathscr{X}} |x - x'|^2$. Due to the monotonicity of the Q function, it is clear that maximizing this distance optimizes the error performance estimated with the union bound at high SNR.

The minimum distance of the constellation depends on the number of rings n_R , the number of points in each ring n_1, \ldots, n_{n_R} , the radii r_1, \ldots, r_{n_R} , and the offset among the rings $\phi_1, \ldots, \phi_{n_R}$. The constellation geometry clearly indicates that the distances to consider are between points belonging to the same ring, or between points in adjacent rings. Simple calculations give the

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following formula,

266

$$\delta_{\operatorname{ring} i}^2 = 2r_i^2 \left[1 - \cos\left(\frac{2\pi}{n_i}\right) \right] \tag{6}$$

for the distance between points in *i*th ring, with radius r_i and n_i points. For the adjacent rings the calculation is only slightly more complicated, and gives the following:

$$\delta_{\text{rings } i,i+1}^2 = r_i^2 + r_{i+1}^2 - 2r_i r_{i+1} \cos \theta \tag{7}$$

where θ is the minimum relative offset between any pair of points of rings *i* and *i*+1, respectively. As the phase of point l_i in ring *i* is given by $\phi_i + 2\pi l_i/n_i$, we easily obtain:

$$\theta = \min_{l_i, l_{i+1}} \left| (\phi_i - \phi_{i+1}) + 2\pi \left(\frac{l_i}{n_i} - \frac{l_{i+1}}{n_{i+1}} \right) \right|$$
(8)

The minimum distance of the constellation is given by taking the minimum of all these inter-ring and intra-ring values:

$$\delta_{\min}^{2} = \min_{\substack{i=1,...,n_{R} \\ j=1,...,n_{R}-1}} \{\delta_{\min j}^{2}, \delta_{\min j,j+1}^{2}\}$$
(9)

For the sake of space limitations, we concentrate on 16-ary constellations. Thanks to symmetry considerations, is clear that the best offset between rings happens when $\phi_2 = \pi/n_2$. Figure 2 shows the minimum distance for several candidates: 4 + 12-, 6 + 10-, 5 + 11 and 1 + 5 + 10-APSK. It may be observed that the highest minimum distance is achieved for approximately $\rho_2 = 2.0$, except for 4 + 12-APSK, where $\rho_2 = 2.7$. The results for $\phi = 0$ are also plotted, and show that the corresponding minimum distance is smaller. We will see later in Section 3.2.2 how this effect translates into error rate performance.

3.2.2. Mutual information maximization. The mutual information (assuming equiprobable symbols) for a given signal set \mathscr{X} provides the maximum transmission rate (in bits/channel use) at which error-free transmission is possible with such signal set, and is given by (e.g. Reference [13]),

$$f(\mathscr{X}) = C = \log_2 M - \mathbb{E}_{x,n} \left\{ \log_2 \left[\sum_{x' \in \mathscr{X}} \exp\left(-\frac{1}{N_0} \left| \sqrt{E_s}(x - x') + n \right|^2 - \left| n \right|^2 \right) \right] \right\}$$
(10)

Interestingly, for a given SNR, or equivalently, for a given spectral efficiency R, an optimum constellation can be obtained, a procedure we apply in the following to 16- and 32-ary constellations.

In general, closed-form optimization of this expression is a daunting task, so we resort to numerical techniques. Expression (10) can be easily evaluated by using the Gauss–Hermite quadrature rules, making numerical evaluation very simple. Note, however, that it is possible to calculate a closed-form expression for the asymptotic case $E_s/N_0 \rightarrow +\infty$. First, note that the expectation in Equation (10) can be rewritten as

$$\lambda(\mathscr{X}) \stackrel{\scriptscriptstyle{\triangle}}{=} \mathbb{E}_{x,n} \left\{ \log_2 \left[\sum_{x' \in \mathscr{X}} \exp\left(-\frac{1}{N_0} \left(E_s |x - x'|^2 + 2 \operatorname{Re}\left(\sqrt{E_s} (x - x')n \right) \right) \right) \right] \right\}$$
(11)

Using the dominated convergence theorem [17], the influence of the noise term vanishes asymptotically, since the limit can be pushed inside the expectation. Furthermore, the only

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Figure 2. Minimum Euclidean distances for several 16-ary signal constellations. Solid lines correspond to $\phi = \pi/n_2$; dotted lines to $\phi = 0$.

remaining terms in the summation over $x' \in \mathscr{X}$ are x' = x and those closest in Euclidean distance $\delta_{\min}^2 = \min_{x' \in \mathscr{X}} |x - x'|^2$, of which there are $n_{\min}(x)$. Therefore the expectation becomes

$$\lambda(\mathscr{X}) \simeq \mathbb{E}_{x} \left\{ \log_{2} \left[1 + n_{\min}(x) \exp\left(-\frac{1}{N_{0}} E_{s} \delta_{\min}^{2}\right) \right] \right\}$$
(12)

Noting that the exponential takes very small values, the approximation $\log_2(1 + x) \simeq x \log_2 e$ for $|x| \ll 1$ holds, thus by simplifying further the expectation we obtain:

$$\lambda(\mathscr{X}) \simeq \mathbb{E}_{x} \left\{ n_{\min}(x) \exp\left(-\frac{E_{s}}{N_{0}} \delta_{\min}^{2}\right) \log_{2} e \right\} \simeq \alpha \exp\left(-\frac{E_{s}}{N_{0}} \delta_{\min}^{2}\right)$$
(13)

where α is a constant that does not depend on the constellation minimum distance δ_{\min} nor on SNR. Then the capacity at large SNR becomes:

$$f(\mathscr{X}) = \log_2 M - \alpha \exp\left(-\frac{E_s}{N_0}\delta_{\min}^2\right)$$
(14)

It appears then clear that the procedure corresponds to the maximization of the minimum Euclidean distance, as in Section 3.2.1.

Figure 3 shows the numerical evaluation of Equation (10) for a given range of values of ρ_2 and $\phi = \phi_2 - \phi_1$ for the 4 + 12-APSK constellation at $E_s/N_0 = 12$ dB. Surprisingly, there is no noticeable dependence on ϕ . Therefore, the two-dimensional optimization can be done by simply finding the ρ_2 that maximizes mutual information. This result was found to hold

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Figure 3. Capacity surface for the 16-APSK ($n_1 = 4$, $n_2 = 12$), with $E_s/N_0 = 12 \text{ dB}$.

true also for the other constellations and hence, in the following, mutual information optimization results do not account for ϕ . Figure 4 shows the union bound on the symbol error probability (5) for several 16-APSK modulations, and for the optimum value of ρ_2 at R = 3 bps/Hz (found with the mutual information analysis). Continuous lines indicate $\phi = 0$ while dotted lines refer to the maximum value of the relative phase shift, i.e. $\phi = \pi/n_2$, showing no dependence on ϕ at high SNR. This absence of dependency is justified by the fact that the optimum constellation separates the rings by a distance larger than the number of points in the ring itself, so that the relative phase ϕ has no significant impact in the distance spectrum of the constellation.

For 16-APSK it is also interesting to investigate the mutual information dependency on n_1 and n_2 . Figure 5(a) depicts the mutual information curves for several configurations of optimized 16-APSK constellations and compared with classical 16-QAM and 16-PSK signal sets. As we can observe, mutual information curves are very close to each other, showing a slight advantage of 6 + 10-APSK over the rest. In particular, note that there is a small gain, of about 0.1 dB, in using the optimized constellation for every R, rather than the calculated with the minimum distance (or high SNR). However, as discussed in Reference [14], 6 + 10-APSK and 1 + 5 + 10-APSK show other disadvantages compared to 4 + 12-APSK for phase recovery and nonlinear channel behaviour.

Similarly, Figure 5(b) reports capacity of optimized 4 + 12 + 16-APSK (with the corresponding optimal values of ρ_2 and ρ_3) compared to 32-QAM and 32-PSK. We observe slight capacity gain of 32-APSK over PSK and QAM constellations. Other 32-APSK constellations with different distribution of points in the three rings did not provide significantly better results.

Finally Table I provides the optimized 16- and 32-APSK parameters for various coding rates, giving an optimum constellation for each given spectral efficiency.

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Figure 4. Union bound on the uncoded symbol error probability for several APSK modulations. Note that the continuous line and the dashed line are indistinguishable because they are superimposed.

3.3. Constellation optimization for nonlinear channels

3.3.1. Peak-to-envelope considerations. For nonlinear transmission over an amplifier, 4 + 12-APSK is preferable to 6 + 10-APSK because the presence of more points in the outer ring allows to maximize the HPA DC power conversion efficiency. It is better to reduce the number of inner points, as they are transmitted at a lower power, which corresponds a lower DC efficiency. It is known that the HPA power conversion efficiency is monotonic with the input power drive up to its saturation point. Figure 6 shows the distribution of the transmitted signal envelope for 16-QAM, 4 + 12-APSK, 6 + 10-APSK, 5 + 11-APSK, and 16-PSK. In this case, the shaping filter is a square-root raised cosine (SRRC) with a roll-off factor $\alpha = 0.35$ as for the DVB-S2 standard [15]. As we observe, the 4 + 12-APSK envelope is more concentrated around the outer ring amplitude than 16-QAM and 6 + 10-PSK, being remarkably close to the 16-PSK case. This shows that the selected constellation represents a good trade-off between 16-QAM and 16-PSK, with error performance close to 16-QAM, and resilience to nonlinearity close to 16-PSK. Therefore, 4 + 12-APSK is preferable to the rest of 16-ary modulations considered. Similar advantages have been observed for 32-APSK compared to 32-QAM.

3.3.2. Static distortion compensation. The simplest approach for counteracting the HPA nonlinear characteristic for the APSK signal, as already introduced in Section 2, is to modify the complex-valued constellation points at the modulator side. Thanks to the multiple-ring nature of the APSK constellation, pre-compensation is easily done by a simple modification of the

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Figure 5. Capacity and ρ^{opt} for the optimized APSK signal constellations vs QAM and PSK: (a) 16-ary constellations (zoom around 3 bps/Hz); and (b) 32-ary constellations.

parameters ρ_{ℓ} , and ϕ_{ℓ} . The objective is to exploit the known AM/AM and AM/PM HPA characteristics in order to obtain a good replica of the desired signal constellation geometry after the HPA, as if it had not suffered any distortion. This can be simply obtained by artificially increasing the relative radii ρ_{ℓ} and modifying the relative phases ϕ_{ℓ} at the modulator side. This

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•	•	•	•	
Modulation order	Coding rate, r	Spectral eff. (bps/Hz)	$ ho_1^{ m opt}$	$\rho_2^{\rm opt}$
4+12-APSK	2/3	2.67	3.15	N/A
4 + 12-APSK	3/4	3.00	2.85	N/A
4 + 12-APSK	4/5	3.20	2.75	N/A
4 + 12-APSK	5/6	3.33	2.70	N/A
4 + 12-APSK	8/9	3.56	2.60	N/A
4 + 12-APSK	9/10	3.60	2.57	N/A
4 + 12 + 16-APSK	3/4	3.75	2.84	5.27
4 + 12 + 16-APSK	4/5	4.00	2.72	4.87
4 + 12 + 16-APSK	5/6	4.17	2.64	4.64
4 + 12 + 16-APSK	8/9	4.44	2.54	4.33
4 + 12 + 16-APSK	9/10	4.50	2.53	4.30

Table I. Optimized constellation parameters for 16- ary and 32-ary APSK.



Figure 6. Simulated histogram of the transmitted signal envelope power for 16-ary constellations.

approach neglects nonlinear ISI effects at the matched filter output which are not present under the current assumption of rectangular symbols; ISI issues has been discussed in Reference [14].

In the 16-ary APSK case, the new constellation points x' follow (4), with new radii r'_1 , r'_2 , such that $F(r'_1) = r_1$, and $F(r'_2) = r_2$. Concerning the phase, it is possible to pre-correct for the relative phase offset introduced by the HPA between inner and outer ring by simply changing the relative phase shift by $\phi'_2 = \phi_2 + \Delta\phi$, with $\Delta\phi = \phi(r'_2) - \phi(r'_1)$. These operations can be readily implemented in the digital modulator by simply modifying the reference constellation parameters ρ' , ϕ' , with no hardware complexity impact or out-of-band emission increase at the

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linear modulator output. On the other side, this allows to shift all the compensation effort into the modulator side allowing the use of an optimal demodulator/decoder for AWGN channels even when the amplifier is close to saturation. The pre-compensated signal expression at the modulator output is then

$$s_T^{\rm pre} = \sqrt{P} \sum_{k=0}^{L-1} x'(k) p_T(t - kT_s)$$
(15)

where now $x'(k) \in \mathscr{X}'$ are the pre-distorted symbols with r'_{ℓ} and ϕ'_{ℓ} for $\ell = 1, \ldots, n_R$.

4. FORWARD ERROR CORRECTION CODE DESIGN AND PERFORMANCE

In this section, we describe the coupling of turbo-codes and the APSK signal constellations through BICM and we discuss some of the properties of this approach.[§] As already mentioned in Section 1, such approach is a good candidate for flexible constellation format transmission. The main drivers for the selection of the FEC code have been flexibility, i.e. use a single mother code, independently of the modulation and code rates; complexity, i.e. have a code as compact and simple as possible; and good performance, i.e. approach Shannon's capacity bound as much as possible.

We consider throughout a coded modulation scheme for which the transmitted symbols $\mathbf{x} = (x_0, \ldots, x_{L-1})$ are obtained as follows: (1) The information bits sequence $\mathbf{a} = (a_0, \ldots, a_{K-1})$ is encoded with a binary code $\mathscr{C} \in \mathbb{F}_2^N$ of rate r = K/N; (2) the encoded sequence $\mathbf{c} = (c_0, \ldots, c_{N-1}) \in \mathscr{C}$ is bit-interleaved, with an index permutation $\boldsymbol{\pi} = (\pi_0, \ldots, \pi_{N-1})$; (3) the bit-interleaved sequence \mathbf{c}_{π} is mapped to a sequence of modulation symbols \mathbf{x} with a labelling rule $\mu : \mathbb{F}_2^M \to \mathscr{X}$, such that $\mu(a_1, \ldots, a_M) = x$. In addition to the description of the code, we also propose the use of some new heuristics to tune the final design of the BICM codes.

4.1. Code design

It was suggested in Reference [13] that the binary code \mathscr{C} can be optimized for a binary channel (such as BPSK or QPSK with AWGN). We substantiate this claim with some further insights on the effect of the code minimum distance in the error performance. The Bhattachharyya union bound (BUB) on the frame error probability P_e for a BICM modulation assuming that no iterations are performed at the demapper side is given by [13]:

$$P_e \leqslant \sum_d A(d) B(E_s/N_0)^d \tag{16}$$

where A(d) is the number of codewords at a Hamming distance d, d_{\min} is the minimum Hamming distance, with $B(E_s/N_0)$ denoting the Bhattachharyya factor, which is given by

$$B(E_s/N_0) = \frac{1}{M\log_2 M} \sum_{i=1}^{\log_2 M} \sum_{b=0}^{1} \sum_{x \in \mathcal{X}_{i=b}} \mathbb{E}_n \left\{ \sqrt{\frac{\sum_{z \in \mathcal{X}_{i=b}} \exp(-(1/N_0)|x-z+n|^2)}{\sum_{z \in \mathcal{X}_{i=b}} \exp(-(1/N_0)|x-z+n|^2)}} \right\}$$
(17)

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[§]The optimization method based on the mutual information proposed in Section 3.2.2 can be easily extended to the case of the BICM mutual information [13] with almost identical results assuming Gray mapping. However, we use the proposed method in order to keep the discussion general and not dependent on the selected coding scheme.



Figure 7. Lower bound d_0 vs normalized E_s/N_0 for target $P_e = 10^{-4}, 10^{-7}$, QPSK and 16- and 32-ary modulations and Gray labelling.

where $\mathscr{X}_{i=b} = \{x \in \mathscr{X} | \mu_i^{-1}(x) = b\}$, where $\mu_i^{-1}(x) = b$ denotes that the *i*th position of binary label x is equal to b. Equation (17) can be evaluated very efficiently using the Gauss–Hermite quadrature rules. For sufficiently large E_s/N_0 the BUB in Equation (5) is dominated by the term at minimum distance, i.e. the error floor

$$P_e \simeq A_{d_{\min}} B (E_s/N_0)^{d_{\min}} \tag{18}$$

From this equation, we can derive an easy lower bound on the d_0 on the minimum distance of \mathscr{C} for a given target error rate, modulation, and number of codewords at d_{\min} :

$$d_{\min} \ge \lceil d_0 \rceil$$
 where $d_0 = \frac{\log P_e - \log A_{d_{\min}}}{\log B}$ (19)

where [x] denotes the smallest integer greater or equal to x. Notice that the target error rate is fixed to be the error floor under ML decoding.[¶] The lowest error probability floor is achieved by a code \mathscr{C} with $A_{d_{\min}} = 1$. Figure 7 shows the lower bound d_0 with $A_{d_{\min}} = 1$, as a function of E_s/N_0 for target $P_e = 10^{-4}$, 10^{-7} , QPSK, 16-QAM, 16- and 32-APSK modulations and Gray labelling. In order to ease the comparison, a normalized SNR is used, defined as

$$\frac{E_s}{N_0}\Big|^{\text{norm}} = \frac{E_s}{N_0} \frac{1}{2^R - 1}$$
(20)

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[¶]Although this does not necessarily hold under iterative decoding, it does still provide a useful guideline into the performance.

where *R* is the spectral efficiency, and the normalization is thus to the channel capacity. The code rate has been taken r = 3/4 for all cases. Note that a capacity-achieving pair modulation-code would work at a normalized $E_s/N_0|^{\text{norm}} = 1$ or 0 dB.

A remarkable conclusion is that BICM with Gray mapping preserves the properties of \mathscr{C} regardless of the modulation used, since we observe that the requirements for nonbinary modulations are similar to those for binary modulations (in the error-floor region). In order to work at about 3 dB from capacity, that is, a normalized $E_s/N_0|^{\text{norm}} = 3 \text{ dB}$, the needed d_0 is about 5 and 10 for a frame error rate of 10^{-4} and 10^{-7} , respectively.

We consider that \mathscr{C} is a serial concatenatation of convolutional codes (SCCC) [18], with outer code \mathscr{C}^{O} of length L_{O} and rate r_{O} and inner code \mathscr{C}^{I} of length L_{I} and rate r_{I} . Obviously, $L_{I} = N$ and $r_{O}r_{I} = r$. The resulting spectral efficiency is $R = r \log_{2} M$. It provides two key advantages with respect to parallel turbo codes: lower error floor, possibly achieving the bit error rate requirements (BER $\leq 10^{-10}$) without any external code; and simpler constituent codes simpler than in turbo codes or in classical concatenated codes. In addition, with an SCCC the outer code is fully integrated into the decoding process, which includes iterations between decoding stages for the inner and outer codes. This avoids the need to use an additional external code, such as a Reed–Solomon (and its associated interleaver). In some sense, the outer code is already included in the SCCC code, thus saving one extra encoding/decoding step, and one memory level, therefore reducing the required complexity.

The best choice in terms of low error floor forces the lowest possible rate for the outer encoder, as this maximizes the interleaver gain, which increases exponentially with the outer code free distance [18]. We should then set the outer code rate equal to the total code rate, and the inner code rate to 1. Also, it turns out that the best choice for the inner encoder is the two-state differential encoder also known as 'accumulator'. It meets the requirements of simplicity, it is 'almost' systematic, in the sense that the dependency among the bits in its output sequence is very mild, and moreover, it is recursive as imposed by the design rules of SCCCs for the inner encoder. Last but not least, this choice leads to a very simple inner SISO, a highly desirable feature for a design working at high data rates.

In practice the maximum block length to be used shall be selected accounting for the maximum allowed end-to-end latency and decoder complexity. One recent finding [19] allows to split an arbitrary block interleaver in an arbitrary number of smaller nonoverlapping interleavers. This allows to greatly reduce the decoder complexity when parallel SISO units are used to achieve high-speed decoding as memory requirements does not increase with the degree of parallelism.

As an outer code, we have selected the standard binary 16-state convolutional code, rate 3/4 [20]. Its free distance is 4, large enough so that interleaving gain can be achieved, and the minimum distance of the concatenated code grows towards infinity with the blocklength [21]. Furthermore, and if required, the code may be punctured to higher rates [22], with no loss in the code distance. Further numerical results are presented in Section 4.3.

4.2. Demodulation

Decoding of BICM consists of a concatenation of two steps, namely maximum-*a posteriori* (MAP) soft-input soft-output (SISO) demapper (symbol-to-bit likelihood computer), and a MAP SISO decoder of \mathscr{C} . These two steps exchange extrinsic information messages $m_{\mu \to \mathscr{C}}$ (from

the demapper to the SISO decoder of \mathscr{C}), and $m_{\mathscr{C} \to \mu}$ (from the SISO decoder of \mathscr{C} to the demapper) through the iterations. Extrinsic information messages *m* (or metrics) can be in the form of likelihood probabilities, log-likelihood ratios or some combination or approximation of them. When either \mathscr{C}^{O} or \mathscr{C}^{I} , or both, are convolutional codes, MAP SISO decoding is efficiently computed by the BCJR algorithm [23]. For example, the extrinsic log-likelihood ratio corresponding to $m_{\mu \to \mathscr{C}}$ for the *i*th coded bit of the *k*th symbol and *l*th iteration is given by

$$\Lambda(c_{k,i}^{(l)}) = \log \frac{\sum_{x \in \mathscr{X}_{i=0}} p(y_k | x) \prod_{j \neq i} P_{\mathscr{C} \to \mu}^{(l-1)}(c_{k,j})}{\sum_{x \in \mathscr{X}_{i=1}} p(y_k | x) \prod_{j \neq i} P_{\mathscr{C} \to \mu}^{(l-1)}(c_{k,j})}$$
(21)

where $p(y_k|x) \propto \exp(-(1/N_0)|y_k - \sqrt{E_s}x|^2)$, $P_{\mathscr{C} \to \mu}^{(l)}(c)$ denotes the extrinsic probability message corresponding to $m_{\mathscr{C} \to \mu}$ on the coded bit c at the *l*th iteration.

There is a marginal information loss in considering no iterations at the demodulator side when Gray mapping is used for transmitting high rates [13], i.e. $P_{\mathscr{C} \to \mu}^{(l-1)}(c_{k,j}) = 0.5$ implies almost no loss in spectral efficiency using Gray mapping. When demapper iterations are allowed, Gray mapping is known not to gain through the iterations [24]. Moreover, when other mapping rules are used, scheduling the operations for such decoder (a SCC with BICM) can be a very complicated task and has been solved only for $N \to \infty$ (see e.g. Reference [25] for recent results on the subject). For all these aforementioned reasons, we will assume Gray mapping and that information flows from demodulator to decoder only, with no feed-back.

4.3. Performance analysis

Density evolution (or approximations such as EXIT charts [24]) of such turbo-coded BICM is a very complicated task due to the concatenation of three elements exchanging extrinsic information messages through the iterations. Such techniques lead in general to threedimensional surfaces which are difficult to deal with in practical decoding algorithms for finite length codes [25]. We will therefore resort to a mixture of computer simulations and bounds on maximum likelihood (ML) decoding error probability. Regarding convergence, simulations can accurately estimate the values of E_b/N_0 at which the decoding algorithm does not converge, as will be shown shortly.

We denote the binary-input channel between the modulator and demodulator as the equivalent binary-input BICM channel. It has been recently shown [26] that such channel can be very well approximated as AWGN^{||} with SNR $\gamma = -\log B(E_s/N_0)$. Therefore, standard bounds for binary-input channels can be successfully applied here. In particular, the standard union bound (UB) yields

$$P_e \lesssim \sum_d A_d Q(\sqrt{-2d \log B(E_s/N_0)})$$
(22)

At high SNR, (16) and (22) are dominated by the pairwise terms corresponding to the few codewords with low Hamming distance (error floor). When turbo-like codes are used, union bounding techniques are known not to provide good estimates of the error probability,

¹Notice that the Gaussian approximation (GA) is common practice in density evolution techniques [24].

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Figure 8. Bit-error probability bounds and simulations for BPSK and 16-APSK ($n_1 = 4$ and $n_2 = 12$) and 32-APSK ($n_1 = 4$, $n_2 = 12$ and $n_3 = 16$) with pseudo-Gray labelling: (a) RA code with r = 1/4 and K = 512 information bits per frame; and (b) SCCC with rate 3/4 16 states convolutional code as outer code and inner accumulator with N = 5000.

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and one typically resorts to improved bounds such as the tangential-sphere bound (TSB) [27],

$$P_{F} \lesssim \int_{-\infty}^{+\infty} \frac{dz_{1}}{\sqrt{2\pi\sigma^{2}}} e^{-z_{1}^{2}/2\sigma^{2}} \left\{ 1 - \overline{\Gamma}\left(\frac{N-1}{2}, \frac{r_{z_{1}}}{2\sigma^{2}}\right) + \sum_{d:\delta/2 < \alpha_{\delta}} A_{d} \overline{\Gamma}\left(\frac{N-2}{2}, \frac{r_{z_{1}}^{2} - \beta_{\delta}(z_{1})^{2}}{2\sigma^{2}}\right) \left[Q\left(\frac{\beta_{\delta}(z_{1})}{\sigma}\right) - Q\left(\frac{r_{z_{1}}}{\sigma}\right) \right] \right\}$$
(23)

where $\overline{\Gamma}(a, x) = (1/\Gamma(a)) \int_0^x t^{a-1} e^{-t} dt$ is the normalized incomplete gamma function and $\Gamma(x) = \int_0^{+\infty} t^{x-1} e^{-t} dt$ is the gamma function,

$$\sigma^2 = (-2\log B(E_s/N_0))^{-1} \tag{24}$$

$$r_{z_1} = r(1 - z_1/4R) \tag{25}$$

$$\beta_{\delta}(z_1) = (r_{z_1}/\sqrt{1 - \delta^2/4R^2})(\delta/2r)$$
(26)

$$\alpha_{\delta} = r\sqrt{1 - \delta^2/4R^2} \tag{27}$$

 $R^2 = N$, $\delta = 2\sqrt{d}$ and r, the cone radius, is the solution of

$$\sum_{d:\delta/2 < \alpha\delta} A_d \int_0^{\theta_k} \sin^{N-3} \phi \, \mathrm{d}\phi = \sqrt{\pi} \Gamma((N-2)/2) / \Gamma((N-1)/2)$$
(28)

with

$$\theta_k = \cos^{-1}((\delta/2r)(1/\sqrt{1-\delta^2/4R^2}))$$
(29)

Figure 8 shows the bit-error probability bounds using the Gaussian approximation, Equation (22) and simulations for BPSK, 4 + 12-APSK and 4 + 12 + 16-APSK with the pseudo-Gray labellings of Figure 1. In particular, in Figure 8(a) we use a repeat and accumulate (RA) code [28] with r = 1/4 and K = 512 information bits per frame. In this case, the weight enumerator can be computed in closed form [28]. We observe that as expected BICM preserves the properties of the underlying binary code C, since both waterfall and floor occur at almost the same probability values. We also observe that the approximations in Equations (22) and (23) are very accurate and yield much better error probability estimates than the standard Bhattacharyya bound. Same conclusions apply to Figure 8(b) where we use a SCCC with the optimal 16 states r = 3/4 convolutional code as outer code and inner accumulator, with interleaver size N = 5000. Storage limitations prevent us from showing the curves for larger N. However, we observe that already with N = 5000, we have very low error floors. In particular, in the DVB-S2 application, the interleaver size is set to 16 200 or 64 000, which implies that almost-error-free transmission is possible with such code.

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Figure 9. Turbo-coded APSK with N = 16200 vs 4 + 12-TCM.

Finally, Figure 9, shows the simulated BER performance for the same SCCC with the optimal 16 states r = 3/4 convolutional code as outer code and inner accumulator, with interleaver size $N = 16200^{**}$ with 4 + 12-APSK, 16-QAM, 4 + 12 + 16-APSK and 32-QAM. For the sake of comparison, we also plot the BER for a 4 + 12-APSK with a 16 states TCM, typical of satellite systems current standard [1]. As we observe, the SCCC codes yield a substantial performance improvement with respect to TCM. In the TCM case, one usually concatenates a Reed–Solomon code operating at an input BER of 10^{-4} , which usually diminishes the spectral efficiency and increases the receiver complexity. Notice as well that 32-APSK achieves a better performance than 32-QAM, giving a further justification to the use of modulations in the APSK family instead of the classical QAM.

5. CONCLUSIONS

Extensive analysis and simulations for turbo-coded APSK modulations, with particular emphasis on its applicability to satellite broadband communications have been presented in this paper. In particular, we have investigated APSK constellation optimization under mutual information and minimum Euclidean distance criteria, under the simplified assumption of

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^{**} The selected FEC block size ensures that the FEC floor is well below the required BER of 10⁻¹⁰ for satellite broadcasting systems.

rectangularly shaped transmission pulses. We have shown that the degrees of freedom in the design of an APSK modulation can be exploited thanks to the mutual information maximization, and this has been applied to the design of 16- and 32-ary constellations. This technique has been shown to extend the standard minimum Euclidean distance maximization, yielding a small but significant improvement.

The pragmatic approach of BICM allows for a good coupling between such optimized APSK modulations with powerful binary turbo-codes, due to its inherent flexibility for multiple-rate transmission. Some new heuristics have been used to further justify the design of a single mother code to be used for all rates. A theoretical explanation of the the fact that the error floor typical of turbo codes remains at a constant distance from capacity has been presented. We have presented some new ML decoding error probability bounds for BICM APSK, and we have compared them with simulations findings. Numerical results based on simulation of bit-error rate probability for high rate transmission with turbo-coded APSK have been presented, showing large advantage of the presented scheme over standard TCM.

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TURBO-CODED APSK MODULATIONS DESIGN

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