INTERNATIONAL JOURNAL OF SATELLITE COMMUNICATIONS AND NETWORKING
Int. J. Satell. Commun. Network. 2006; **24**:261–281
Published online 19 May 2006 in Wiley InterScience (www.interscience.wiley.com). DOI: 10.1002/sat.841

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.

Received 1 May 2005; Revised 1 February 2006; Accepted 29 March 2006

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

DOI: 10.1002/sat

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

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

where P is the signal power, $x(k)$ is the kth 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

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

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 $s_T(t) = |s_T(t)|\mathrm{e}^{\mathrm{j}\phi(s_T(t))}$.

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 $x = |x|\mathrm{e}^{\mathrm{j}\phi(x)} \in \mathscr{X}$, that is, $x_A = F(|x|)\mathrm{e}^{\mathrm{j}(\phi(x)+\Phi(|x|))}$. After matched filtering and sampling at time kT_s, the discrete-time received signal at time k, $y(k)$ is then given by

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

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

DOI: 10.1002/sat

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 $\mathcal{X}$ given by

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

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 ℓth 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 $|\mathcal{X}| = 16$ and 32, with $n_R = 2$ and 3 rings, respectively. In general, we consider that $\mathcal{X}$ 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 $\ell = 1, \ldots, n_R$ be the phase shift of the ℓth ring with respect to the inner ring. We also define $\rho_\ell = r_\ell / r_1$ for $\ell = 1, \cdots, n_R$ as the relative radii of the ℓth 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 $\boldsymbol{\rho} = (\rho_1, \ldots, \rho_{n_R})$ and $\boldsymbol{\phi} = (\phi_1, \ldots, \phi_{n_R})$, 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.

DOI: 10.1002/sat

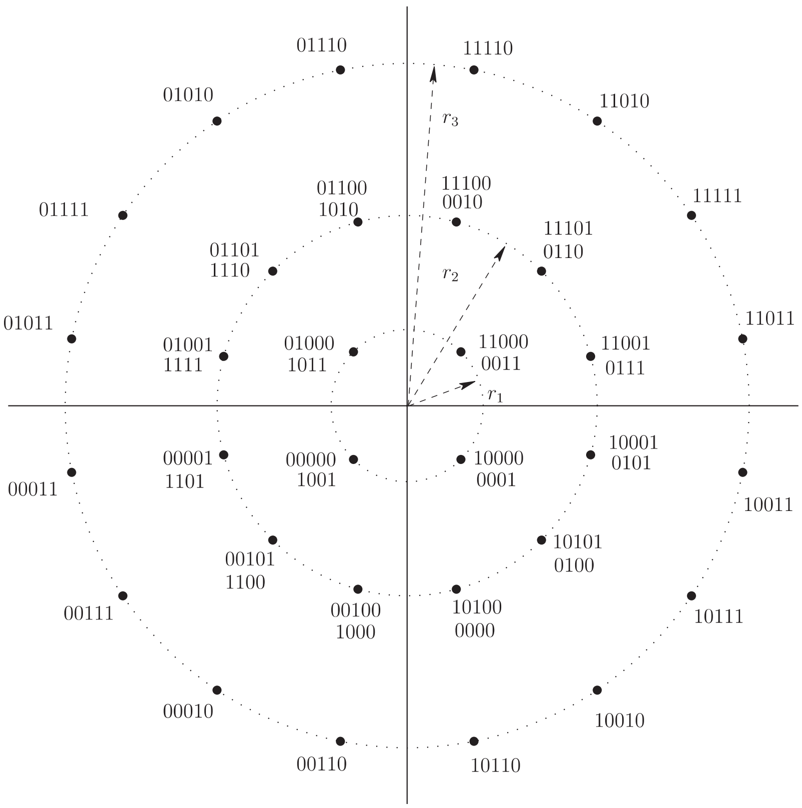

Figure 1. Parametric description and pseudo-Gray mapping of 16 and 32-APSK constellations with $n_1 = 4$, $n_2 = 12$, $\phi_2 = 0$ and $n_1 = 4$, $n_2 = 12$, $n_3 = 16$, $\phi_2 = 0$, $\phi_3 = \pi/16$. For the first two rings: mapping below corresponds to 4 + 12-APSK, mapping above to 4 + 12 + 16-APSK.

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

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

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 $\delta^2_{\min} = \min_{x,x' \in \mathcal{X}} |x - x'|^2$. 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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