Spectral Efficiency of Linear and Continuous Phase Modulations over Nonlinear Satellite Channels

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***Abstract*—We consider a frequency-division-multiplexed satellite system where nonlinear distortions may originate from the presence of high power nonlinear devices and can cause significant performance degradations. The spectral efficiency is used as a performance measure to compare, from an information-theoretic point of view, different transmission strategies and modulation formats. More precisely, we will consider transmission schemes employing continuous phase modulations, which are robust to nonlinearities, and schemes based on linear modulations and employing a detector taking into account the nonlinear effects or more traditional techniques, such those based on predistortion of the nonlinear device.**

I. Introduction

Many efforts in the literature of the last decades have been devoted to nonlinear channel compensation techniques for satellite communications, where nonlinear distortions may originate from the presence of high power nonlinear devices and can cause significant performance degradations. The approaches can be essentially classified in techniques applied at the transmitter or at the receiver side. Pre-compensation techniques at the transmitter try to mitigate the nonlinear effects through analog signal predistortion [1], [2] or data predistortion [3]. DVB-S2 [4] systems adopt advanced data predistortion methods to overcome the effect of transponder impairments and in [5] a dynamic predistortion algorithm is proposed which minimizes the total link degradation. If the characteristics of the channel are known at the receiver, the optimal maximum-a-posteriori (MAP) *symbol* detector for the nonlinear channel is perfectly defined and, as explained in [6], takes the form of the well-known BCJR algorithm [7] with the same branch metrics of the optimal MAP sequence detector [8]. Similarly to low-complexity detection schemes designed in the literature for linear intersymbol interference (ISI) channels [9], [10], alternative suboptimal detection algorithms for nonlinear channels are based on a Gaussian approximation of the (linear and nonlinear) ISI term (e.g., see [11], [12]). The algorithm in [12] relies on a more accurate signal model, based on a Volterra-series expansion [13], [14]. In [15], a low-complexity soft-input soft-output detection algorithm is derived by using the the framework based on factor graphs and the sum-product algorithm [16].

A valid alternative to nonlinear compensation techniques relies on the adoption of advanced modulation schemes which are robust to nonlinearities. In this paper, we will focus on continuous phase modulations (CPMs), which are appealing for satellite systems for their immunity against nonlinear distortions, stemming from the constant envelope, for their claimed power and spectral efficiency, and for their recursive nature which allows to employ them in serially concatenated schemes [17], [18]. CPMs are often employed in satellite communications and they have been recently included in the 2nd-generation Digital Video Broadcasting - Return Channel Satellite (DVB-RCS2) standard [19], with the aim of enabling the use of cheaper amplifier components in modems, and hence lower cost terminals.

In this work, all these approaches are evaluated from an information-theoretic point of view, assuming a realistic satellite channel. In particular, we evaluate the asymptotic performance of frequency-division-multiplexed (FDM) satellite systems when single-user detectors are employed at the receiver side by computing the achievable information rate (IR) and the spectral efficiency (SE). We use the information-theoretic approach to identify the most promising modulation schemes, and to compare different transmission strategies. In particular, we compare the performance of a detector taking into account the nonlinear effects with more traditional techniques, such those based on predistortion of the nonlinear device, and with a transmission scheme employing CPMs.

This paper is organized as follows. In Section II we introduce the system model. The framework that we use to compare different transmission strategies and modulation formats is described in Section III, whereas the results of our study are presented in Section IV. Finally, conclusions are drawn in Section V.

II. System Model

We consider a return-link satellite channel, where the satellite transponder bandwidth is supposed to be accessed with a FDM technique. In particular, we assume that $2U+1$ independent users simultaneously access the channel, adopt the same modulation format, and transmit at the same power. We assume that each user transmits N symbols and we denote by $x_n^{(u)}$ the information symbol transmitted by user u at discrete-time n and by $\mathbf{x}^{(u)}$ the vector collecting the N symbols of user u. We assume that the transmitted symbols, belonging to a given zero-mean M-th order complex constellation, are independent, uniformly distributed, and normalized to have unit power. The information symbols, possibly predistorted to form the vectors $\mathbf{y}^{(u)}$, are fed to the modulator.

The modulated signal passes through a high-power amplifier (HPA), which is a nonlinear memoryless device defined through its AM/AM and AM/PM characteristics [20], here assumed known at the receiver. They describe the amplitude and phase distortions caused on the signal at its input.

The received signal is also corrupted by additive white Gaussian noise whose low-pass equivalent $w(t)$ has power spectral density (PSD) $2N_0$. The low-pass equivalent of the received signal has thus expression

$$r(t) = \sum_{u=-U}^{U} s_A(t-\tau^{(u)}, \mathbf{y}^{(u)}) e^{j2\pi f^{(u)}(t-\tau^{(u)})} + w(t) \,, \quad (1)$$

where $s_A(t, \mathbf{y})$ is the signal at the output of the HPA, $\tau^{(u)}$ is the relative time offset of user u, and $f^{(u)}$ is the difference between the carrier frequency of user u and the frequency assumed as reference for the computation of the complex envelope. The overall system model is shown in Fig. 1. In this model, we assume that the on-board satellite amplifier works far from the saturation to avoid distortions on the composite signal—this is a common operating choice for this kind of systems.

In the following, we consider transmission schemes employing CPMs and schemes based on linear modulations.

A. Linear Modulations

When linear modulations are employed, the modulated signal for the generic user u reads

$$s(t, \mathbf{y}^{(u)}) = \sum_n y_n^{(u)} p(t-nT) \,,$$

where $p(t)$ is the shaping pulse and T is the symbol interval. We assume that $p(t)$ is a properly normalized root-raised-cosine- (RRC-) shaped pulse with roll-off factor α. In this work, we consider two scenarios based on linear modulations. In the first one, we assume that the transmitter does not know the nonlinear channel and the receiver employs an advanced detector which takes into account the nonlinear effects. In this case, no predistortion is applied and symbols $\{y_n^{(u)}\}$ are actually the information symbols. In the other scenario, techniques based on constellation predistortion and a simpler detector are assumed.

B. Continuous Phase Modulations

For schemes based on CPMs, we assume that symbols $\{x_n^{(u)}\}$ take on values in the M-ary alphabet $\{\pm1, \pm3, \cdots, \pm(M-1)\}$. Thanks to the robustness to nonlinearities of CPM signals, no predistortion technique is necessary in this case. The modulator in Fig. 1 is a CPM modulator and hence the signal $s(t, \mathbf{x}^{(u)})$ is the CPM information-bearing signal of user u,

$$s(t, \mathbf{x}^{(u)}) = \sqrt{\frac{1}{T}} \exp\Big\{ j2\pi h \sum_{n=0}^{N-1} x_n^{(u)} q(t-nT) \Big\},$$

where $q(t)$ the *phase-smoothing response*, and $h = r/p$ the modulation index (r and p are relatively prime integers). The derivative of $q(t)$ is the so-called *frequency pulse*, of length L symbol intervals. In the generic time interval $[nT, nT+T)$, the CPM signal of user u is completely defined by symbol $x_n^{(u)}$ and state $\sigma_n^{(u)} = (\omega_n^{(u)}, \phi_n^{(u)})$ [21], where

$$\omega_n^{(u)} = (x_{n-1}^{(u)}, x_{n-2}^{(u)}, \dots, x_{n-L+1}^{(u)})$$

is the correlative state and $\phi_n^{(u)}$ is the phase state which can be recursively defined as

$$\phi_n^{(u)} = [\phi_{n-1}^{(u)} + \pi h x_{n-L}^{(u)}]_{2\pi},$$

where $[\cdot]_{2\pi}$ denotes the "modulo 2π" operator, and takes on p values.

III. Information-Theoretic Analysis

In this section, we describe the framework that we use to evaluate, from an information-theoretic point of view, the ultimate performance limits of FDM satellite systems. More precisely, we compute the achievable IR when a single-user detection algorithm is adopted at the receiver. The described framework will allow us to find the optimal spacing between adjacent channels, and will be used to compare different transmission strategies and modulation formats.

As mentioned, we assume that all users transmit at the same power and employ the same modulation format. Moreover, we assume that the channels are equally spaced in frequency. Under these conditions, the frequency spacing is a measure of the signal bandwidth and the SE can thus be computed. In order to avoid boundary effects, we assume $U \to \infty$. However, for complexity reasons, we consider a single-user detector that assumes the presence of only one user and treats the other remaining users as additional noise. In the following, we consider single-user detection of the information symbols $\mathbf{x}^{(0)}$. Without loss of generality, the delay of the central channel can be assumed to be zero, that is $\tau^{(0)} = 0$.

We first consider the case where no channel state information (CSI) is available at the transmitter, that is no predistortion is applied on the symbol constellation. Since we consider single-user detection, the channel model assumed by the receiver is

$$r(t) = s_A(t, \mathbf{x}^{(0)}) e^{j2\pi f^{(0)} t} + n(t), \quad (2)$$

where $n(t)$ is a zero-mean circularly symmetric white Gaussian process with PSD $2(N_0 + N_I)$, N_I being a design parameter which will be optimized through computer simulations.

Notice that the goal here is to evaluate the ultimate performance limits achievable by a receiver designed for the auxiliary channel (2) when the actual channel is that in (1) with $U \to \infty$. This problem is an instance of mismatched detection [22], and can be solved by means of the simulation-based method described in [23]. The method in [23] requires the existence of an algorithm for exact MAP symbol detection over the auxiliary channel.

For communication systems based on CPM, algorithms for MAP symbol detection can be derived with a frontend based on the Rimoldi decomposition [21] and the BCJR algorithm [7].

In systems based on linear modulations, the nonlinear device will introduce ISI on the transmitted signal, since we are using

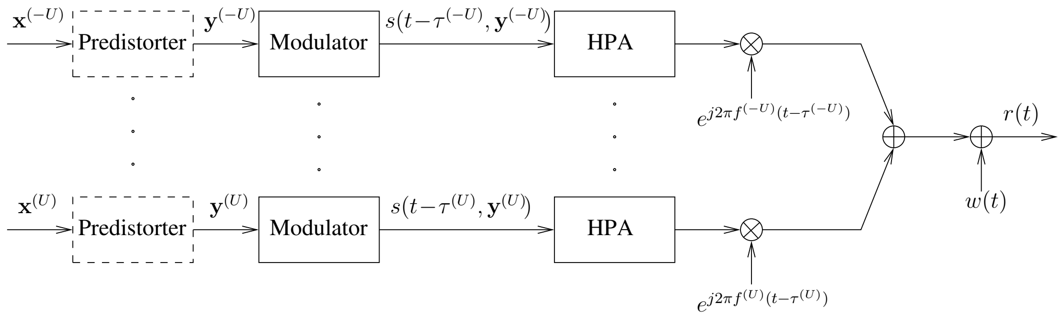

Fig. 1. System model.

a base pulse with support larger than one symbol interval. Assuming that the system is finite-memory, we can model the modulator and the HPA as a finite-state machine (FSM), whose input is the symbol sequence $\mathbf{x}^{(0)}$ and whose output $s_A(t, \mathbf{x}^{(0)})$ can be expressed as

$$s_A(t, \mathbf{x}^{(0)}) = \sum_{n=0}^{N-1} \bar{s}(t - nT, x_n^{(0)}, \sigma_n^{(0)}),$$

where signal $\bar{s}(t-nT, x_n^{(0)}, \sigma_n^{(0)})$ is assumed to have support in the interval $[n, (n+1)T)$. The state $\sigma_n^{(0)}$ of the FSM contains the previous L channel inputs, L being the memory length of the channel:

$$\sigma_n^{(0)} = (x_{n-1}^{(0)}, x_{n-2}^{(0)}, \ldots, x_{n-L}^{(0)}).$$

Therefore the optimal MAP symbol receiver consists of a bank of filters [14] matched to all possible $M^{(L+1)}$ waveforms $\bar{s}(t - nT, x, \sigma)$, followed by a BCJR detector. We point out that, in principle, the real channel memory can be much larger than that assumed by the detection algorithm, which adds a further degree of mismatch of the receiver—the choice of L is often dictated by implementation complexity reasons.

When CSI is available at the transmitter, the sequence of transmitted symbols $x_n^{(0)}$ can be properly modified in order to attenuate the effect of the following non-linearity and possibly reduce the resulting ISI. Here we consider the dynamic data predistortion technique described in [5], where the transmitted symbol at time n is a function of a sequence of input symbols $y_n^{(0)} = \ell(x_{n-L/2}^{(0)}, \ldots, x_{n+L/2}^{(0)})$. At the receiver a simple memoryless channel is assumed, corresponding to the following model for the auxiliary channel:

$$r(t) = \sum_{n=0}^{N-1} x_n'^{(0)} p(t - nT) e^{j2\pi f^{(0)} t} + n'(t). \qquad (3)$$

The corresponding optimal receiver is also memoryless and based on the sampled output of the matched filter. The mapping ℓ at the transmitter, implemented through a LUT, is obtained with a gradient algorithm that minimizes the signal-to-noise ratio of the auxiliary channel at the receiver. Also, the positions of the nominal constellation points at the receiver $\{x_n'^{(0)}\}$, called "centroids" in [5], are preliminarily computed.

We now evaluate the ultimate performance limits when single-user receivers for the auxiliary channels (2) and (3) are adopted. Denoting by $\mathbf{r}$ a set of sufficient statistics for the detection of $\mathbf{x}^{(0)}$, we first compute the IR as

$$I(\mathbf{x}^{(0)}; \mathbf{r}) = \lim_{N\to\infty} \frac{1}{N} E\left\{ \log \frac{p(\mathbf{r}|\mathbf{x}^{(0)})}{p(\mathbf{r})} \right\} \quad \left[\frac{\text{b}}{\text{ch. use}} \right]. \qquad (4)$$

In the case of linear modulations with advanced receiver and in the case of CPMs, the probability density functions $p(\mathbf{r}|\mathbf{x}^{(0)})$ and $p(\mathbf{r})$ can be computed by a forward recursion of the described MAP symbol detectors matched to the auxiliary channel (2) [23]. In (4), the expectation is with respect to the input and output sequences generated according to the model in (1). Assuming a system with an infinite number of users, the IR in (4) does not depend on the specific user. Moreover, we can define the system bandwidth as the separation between adjacent channels $F = |f^{(i)} - f^{(i-1)}|$ and use it in the definition of the achievable SE

$$\text{SE} = \frac{1}{FT} I(\mathbf{x}^{(0)}; \mathbf{r}) \quad [\text{bps/Hz}].$$

Notice that the approach described in this section, due to the use of a mismatched detector, leads to achievable lower bounds on the IR and the SE.

IV. Performance Evaluation

The assumed amplifier at the Earth station is a solid-state power amplifier (SSPA), whose AM/AM characteristic is described by the Rapp model [24], which specifies the amplitude A of the output signal as a function of the amplitude ρ of the input signal as

$$A(\rho) = \frac{\rho}{(1 + \rho^{2s})^{1/2s}}.$$

In this paper, we use $s = 2$. There is actually no generally accepted applicable model for the SSPA AM/PM characteristic, and the manufacturers only specify the maximum slope in degrees/dB and the input level where the phase crosses 0 degree. Here we assume a slope of 2 degrees/dB when the input level is bigger than -1.5 dB and no phase distortion below that level. The nonlinear transfer characteristics are shown in Fig. 2. The working point of the amplifier is generally given in terms of the output back-off (OBO), which is defined as the power ratio (in dB) between the unmodulated carrier at saturation and the modulated carrier after the HPA. Correspondingly, the intput back-off (IBO) is the input power

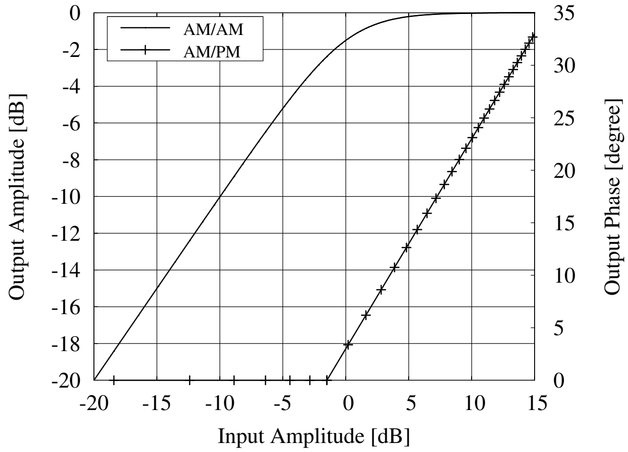

Fig. 2. SSPA characteristics: AM/AM and AM/PM.

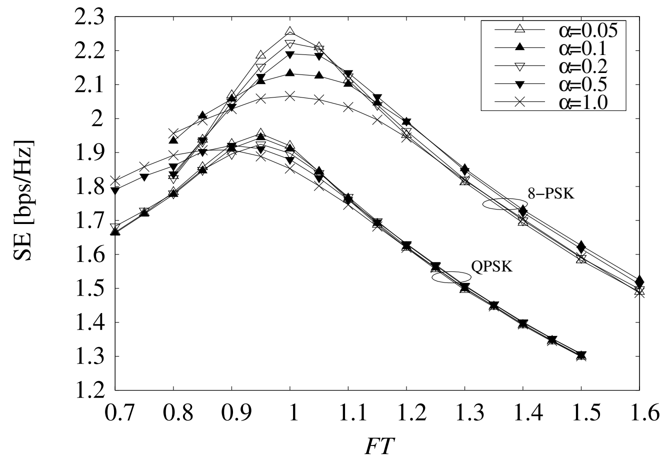

Fig. 3. Spectral efficiency of QPSK and 8-PSK modulations as a function of the normalized spacing, for different values of α and for PSNR=10 dB.

TABLE I
CONSIDERED MODULATION SCHEMES.

	M	h	pulse	FT		α	FT
B-CPM	2	1/3	2-REC	0.4	QPSK	0.05	0.95
Q-CPM	4	1/6	2-REC	0.5	8-PSK	0.05	1.0
8-CPM	8	1/7	2-REC	1.0	16-APSK	0.05	1.0

in dB relative to the value at saturation. In this work, for each modulation format and for each SE, the IBO is optimized through computer simulations.

A. Optimization of the Spectral Efficiency

For transmission schemes employing CPMs, we consider the best formats designed in [25], [26] in the case of a single-user detector, with rectangular frequency pulse of duration $2T$ (2-REC).

For the systems based on linear modulations, we consider phase shift keying (PSK) and amplitude/phase shift keying (APSK) modulations typically employed in satellite transmissions [4]. More precisely, we consider QPSK, 8-PSK and 4+12-APSK modulations, and we use the information-theoretic analysis to identify the most efficient schemes. We compute the IR for systems with no CSI at the transmitter and for systems with constellation predistorter, varying the frequency spacing F and optimizing the IBO, for *peak* signal-to-noise ratio (PSNR) values in the interval $[0, 20]$ dB. The PSNR is defined as the ratio between the transmitted energy per symbol when the amplifier is driven at saturation and the noise PSD. To limit the receiver complexity, the advanced detector for linear modulations assumes that the memory associated with the ISI is of three symbols ($L = 3$) for QPSK and 8-PSK formats, and of two symbols ($L = 2$) for APSK formats. On the other hand, to limit the complexity of the transmitter in the predistortion case we bounded the memory L of the predistorter to 4 for QPSK and to 2 for 8PSK and 16APSK. Fig. 3 shows the SE achievable by the use of the advanced detector, as a function of the spacing F for QPSK and 8-PSK modulation formats, and for PSNR=10 dB. The optimization is performed for RRC pulses with different roll-off factors α, ranging from 0.05 to 1. The results for 16-APSK are not shown for a lack of space. For each modulation, the analysis reveals that schemes with $\alpha = 0.05$ have the best performance in terms of SE. The same conclusion is drawn in the case of predistortion, and hence the best formats for systems with the advanced receiver are also the best ones for predistortion based systems. It is known that schemes employing RRC pulse with smaller roll-off are much more sensitive to nonlinear distortions due to the higher peak-to-average-power ratio, but interestingly our analysis shows that they allow to achieve higher SEs, which means that they provide a better trade-off between degradation due to nonlinear distortions and usage of the available spectrum.

In this analysis, we noticed that the dependence of the optimal spacing values on the PSNR is very limited, and for each modulation format we used the same value in the considered PSNR range. On the contrary, the IBO needs to be properly optimized. The optimized modulation schemes and the corresponding frequency spacings are listed in Table I. For APSK formats, the radii between the two constellation rings is set to 2.57, which is one of the values proposed in [27]. Better results could be obtained by re-optimizing this value for this specific scenario.

B. Comparison between Linear and Continuous Phase Modulations

We first consider systems with advanced receivers and compare linear and continuous phase modulations. In Fig. 4, we show the SE as a function of E_b/N_0+OBO, being E_b the energy per information bit. In this comparison, we consider the OBO to take into account the loss of the received power induced by the back-off of the amplifier on the linearly modulated signals. The figure shows that the considered schemes perform similarly for low values of SE and that quaternary and octal CPMs perform only slightly better than QPSK schemes. However, 8-PSK and 16-APSK formats allow achieving a higher SE.

C. Comparison between Predistortion Techniques and Advanced Detection

In Fig 5, we show the SE of linear modulations with and without preditortion at the transmitter. We recall that the

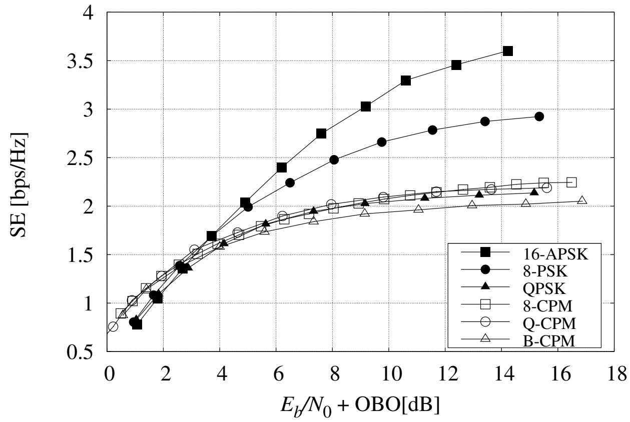

Fig. 4. Spectral efficiency as a function of E_b/N_0+OBO of linear and continuous phase modulations.

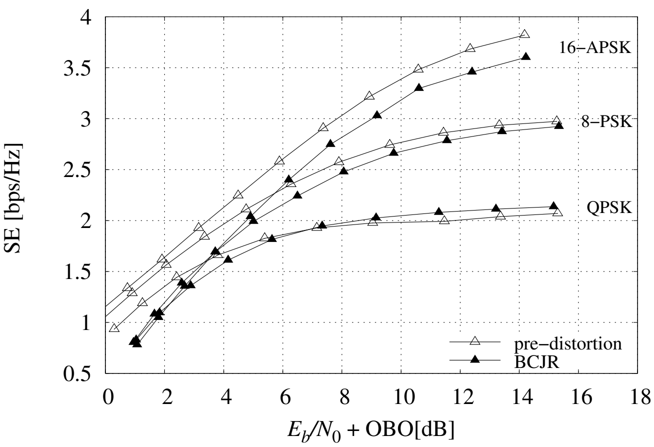

Fig. 5. Spectral efficiency as a function of E_b/N_0+OBO of linear modulations. Comparison between the advanced receiver and the symbol-by-symbol receiver with predistortion.

receiver is different in the two cases. The results show that in the considered scenario the adoption of symbol predistortion provides better results in terms of SE, despite the use of a less complex detection algorithm.

V. Conclusions

We considered frequency-division-multiplexed satellite systems where high power amplifiers may introduce nonlinear distortions. Through an information-theoretic analysis, we investigated the ultimate performance limits of these systems in terms of information rate and spectral efficiency. The proposed analysis allowed us to compare different modulation formats and transmission strategies, considering a realistic satellite system and without requiring extensive end-to-end performance simulations.

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