Adaptive Coded Continuous-Phase Modulations for Frequency-Division Multiuser Systems

Alberto Perotti, Piotr Remlein and Sergio Benedetto

***Abstract*—In this paper we study adaptive coded modulations for wireless channels. Nonlinearity due to the radio-frequency power amplifier is considered and continuous-phase modulations (CPM) are adopted in order to make the nonlinearity effects negligible. The structure and properties of CPM are reviewed in a semi-tutorial fashion and coded modulation schemes are proposed. Moreover, frequency-division multiplexing (FDM) multiuser systems employing coded CPM are considered. Modulation-coding schemes achieving increasing spectral efficiencies in a multiuser scenario with tight intercarrier frequency spacing are designed and their performance in terms of spectral efficiency and error rate is assessed.**

***Index Terms*—Continuous-phase modulation, coded, multiuser, adaptive, cancellation, iterative, nonlinear.**

I. Introduction

AMONG the most relevant challenges in the design of wireless transceivers are those concerning the physical layer. It consists of an analog front-end (AFE), including a baseband and radio-frequency (RF) analog processing subsystem and a digital baseband processor. At the transmitter, the AFE performs digital to analog conversion and frequency conversion from baseband RF. The converted signal is then amplified and sent to the antenna. At the receiver side, the AFE consists of the analog subsystem that amplifies the received signal, converts it to baseband and then to digital samples.

The impairments caused by the AFE on the transmitted signal include the distortion caused by the nonlinear characteristic of the transmitter's RF high power amplifier (HPA), which is known to cause a possibly significant degradation. HPAs are often driven close to or at saturation in order to achieve the highest possible transmitted power. In such case, however, their nonlinear input-output characteristic results in a transmitted signal affected by several kinds of degradations, such as *intermodulation distortion*, *spectral widening*, *amplitude* and *phase distortion*.

These impairments can be easily explained considering the memoryless nonlinear model [1]–[3] often used to characterize the HPA. According to this model, the complex envelope of the output signal is

$$x(t) = A[|r(t)|]\exp\{j\Phi[|r(t)|]\}r(t) \tag{1}$$

where $r(t)$ (resp. $x(t)$) is the complex envelope of the HPA input (resp. output) signal, $A[|r(t)|]$ is the input-output amplitude characteristic (the AM-AM conversion characteristic) , and $\Phi[|r(t)|]$ is the input-output phase characteristic (the AM-PM conversion characteristic).

When viewed in the frequency domain, the nonlinear transformation experienced by the input signal corresponds to a possibly multiple convolution of $R(f) = \mathcal{F}[r(t)]$ with itself[1]. Hence, the power spectral density (PSD) of $x(t)$ is spread over a larger frequency range, thus resulting in a wider spectrum.

When viewed in time domain, the HPA's nonlinear memoryless transformation applied to the sum of two monochromatic signals with distinct frequencies, f_1 and f_2, in the signal bandwidth causes an unwanted amplitude modulation which results in infinite monochromatic components – the intermodulation products – at the PA output, each characterized by a frequency $f_{ij} = if_1 + jf_2$, with $i, j \in \mathbb{Z}$. Some of these components fall into the signal bandwidth, hence causing a kind of self-interference called *intermodulation*.

A possible approach to overcome these impairments is predistortion (see [4] and references therein). This approach requires an accurate knowledge of the PA characteristic and the implementation of a suitable predistortion circuit, which adds complexity to the system. The key observation that both the AM-AM conversion and AM-PM conversion in (1) depend only on the *amplitude* $|r(t)|$ of the input signal leads to the following consideration: an input signal whose complex envelope exhibits *constant-amplitude* would eliminate the impairments due to AM-AM conversion and result in a fixed phase rotation of the complex envelope, which could be easily recovered by the receiver's carrier phase recovery circuit without requiring further compensation.

This issue is perhaps the main motivation that led to the adoption of Continuous-Phase Modulations (CPM) [5], a rich class of bandwidth-efficient modulation techniques featuring the *constant envelope* property. In fact, as shown in Fig. 1, while the complex envelope of a CPM signal has a constant amplitude (see Fig. 1), the complex envelope of a typical QPSK signal exhibits deep amplitude variations.

In the rest of this paper, we will focus on the study of constant-envelope CPM signals and coded CPM systems in a multiuser uplink scenario. The case of many consumer-grade terminals equipped with a low-cost transmitter AFE is interesting, e.g., in the context of uplink satellite systems like the Digital Video Broadcasting (DVB) standard's Return Channel to Satellite (RCS) [6].

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Alberto Perotti and Sergio Benedetto are with Politecnico di Torino, Dipartimento di Elettronica, Corso Duca degli Abruzzi, 24 - I-10129 - Torino (Italy). Piotr Remlein is with Poznan University of Technology, Faculty of Electronics and Telecommunications, Piotrowo 3a; 60-965 Poznan (Poland).

[1]Here, $\mathcal{F}$ denotes the Fourier transform

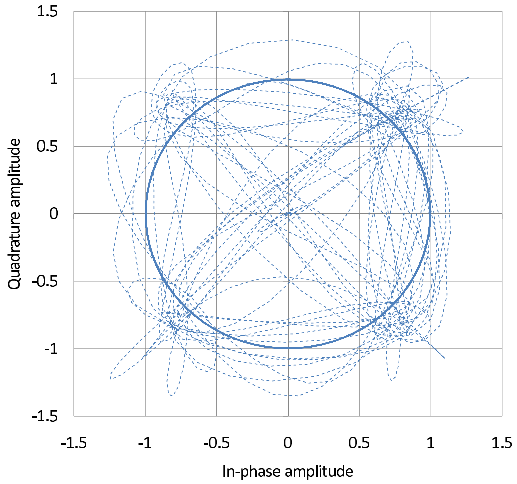


Fig. 1. Trajectories of QPSK (dashed line) and CPM (solild line) complex envelopes with unit power.

CPM has long been considered a complex modulation with poor spectral efficiency. In fact, due to the constant-envelope constraint, the spectral efficiency of simple CPM schemes like the *minimum-shift keying* (MSK) [5], is low. Recently, however, new results presented in [7]–[9] show that CPM modulations can achieve high spectral efficiencies, at the cost of a possibly higher receiver complexity. In fact, CPM modulators are devices with memory and, as we will see in detail in the following sections, their maximum-likelihood (ML) receiver is exponentially complex in the modulator's memory size which is, in turn, inversely related to the bandwidth occupied by the modulated signal. As a consequence, only CPM schemes with low complexity, and therefore large bandwidth, have been considered in the past for practical implementations. In the GSM standard, for example, the MSK scheme, although with a modified (Gaussian) frequency pulse, has been adopted.

Multiuser CPM systems in which users share the same spectrum [10], [11] or feature a tight intercarrier frequency spacing [12] have been studied recently. In [10], multiuser CPM systems with zero frequency and phase offset (*downlink* assumption) over Rayleigh fading channels have been studied. It has been shown that, using an antenna array at the receiver and an optimum multiuser detector, such systems perform very close to systems free of interference. In [11], an asynchronous multiuser CPM system wherein all users share the same carrier frequency has been considered. Optimum and reduced-complexity multiuser detectors based on the Viterbi algorithm have been proposed. In [12], the authors considered a multiuser SCCPM system in which users are allowed to have individual energy levels, carrier frequencies and phases. The authors show that, letting the code word size grow to infinity, the bit error probability vanishes when the channel SNR is sufficiently large.

In this paper, we will show that it is possible to design adaptive coded CPM systems achieving a high spectral efficiency. We will show that even the simplest CPM schemes, like the MSK, achieve a relatively high SE when used in an FDM context with low intercarrier frequency spacing. To obtain an improved SE, interference cancellation is performed at the receiver. Interference cancellation results essential in multiuser coded FDM-CPM adaptive systems in which all users are allowed to choose the best modulation and coding (ModCod) format according to the channel conditions. In fact, the PSD of a CPM signal has infinite support and therefore the amount of interchannel interference (ICI) experienced by each user link depends on the ModCod used in the adjacent links. In such a scenario, if ICI cancellation is not performed, the choice of the optimal ModCod scheme for each link, a task performed at the receiver by means of a suitable adaptivity algorithm, would result in a very complex optimization problem. Using ICI cancellation, each user link becomes virtually immune to the interference caused by its frequency-adjacent links. As a result, ModCod selection can be performed on a link-by-link basis, thus greatly simplifying the adaptivity algorithm.

In our analysis, we will make the following assumptions: an FDM-CPM system with a fixed and tight intercarrier frequency spacing is considered. Moreover, we allow each user to have individual phases and delays, resulting in an *uplink* assumption. Ideal power control is assumed, therefore the signal power at the receiver is equal for all users. We show that it is possible to obtain significant improvements in SE by using a simple iterative ICI cancellation technique at the receiver. The resulting receiver complexity is slightly increased by the ICI canceler, although it only grows linearly with the number of users.

The paper is organized as follows: in Sec. II, we review the properties of CPM signals and the structure of CPM modulators and receivers. In Sec. III, we introduce in detail the adopted system model. Sec. IV describes the multiuser receiver including the proposed iterative ICI cancellation algorithm for uncoded CPM. In Sec. V, the channel coding scheme is described and the iterative ICI cancellation algorithm is embedded into the iterative decoding process. Sec. VI describes the design of the adaptive coding and modulation schemes. Sec. VII presents the obtained results.

II. Structure of CPM Signals and Modulators

A continuous-phase, constant envelope modulated signal can be written as

$$x(t) = \sqrt{\frac{2E_s}{T}} e^{j\psi(t)} \tag{2}$$

where E_s is the symbol energy, T is the symbol interval and

$$\psi(t) = 2\pi h \sum_{n=-\infty}^{\infty} a_n q(t - nT). \tag{3}$$

The phase process $\psi(t)$ depends on the input information symbols $a_n \in \{\pm 1, \pm 3, \ldots, \pm(M-1)\}$, where $M = 2^m$ is the size of the input alphabet and $h = Q/P$ is the *modulation index* (Q and P are relatively prime integers). The *phase pulse* $q(t)$ is a continuous function whose support extends over L consecutive symbol intervals[2] and exhibits the following

[2]When $L = 1$, the resulting CPM scheme is *full-response*, while for $L > 1$ we have a *partial response* scheme.

properties

$$q(t) = \begin{cases} 0 & t \le 0 \\ \frac{1}{2} & t \ge LT \end{cases} .$$

The phase pulse is usually defined as the integral of a *frequency pulse* $s(t)$

$$q(t) = \int_{-\infty}^{t} s(\tau) d\tau.$$

Different frequency pulse shapes have been proposed [5]. The most commonly used are the rectangular (REC) and the raised-cosine (RC) frequency pulses. A CPM scheme is then defined by specifying its parameters M, h, L and $s(t)$.

When a signal like (2) enters a nonlinear device, whose characteristic is modeled as in (1), it undergoes amplitude and phase distortion. However, since its amplitude

$$|r| = \sqrt{\frac{2E_s}{T}}$$

is constant, the amplitude distortion results in a gain factor $A[\sqrt{2E_s/T}]$ and the phase distortion results in a constant phase shift $\Phi[\sqrt{2E_s/T}]$, which is compensated for by the carrier phase recovery circuit at the receiver.

A nice feature of CPM signals is their bandwidth efficiency: although the PSD of CPM signals has infinite support, it exhibits a large main lobe centered around the carrier frequency with rapidly decreasing tails. The width and height of the main lobe are related to parameter L, also called the *correlation length*. In fact, a large L induces a correlation between adjacent symbols in the modulated process. Since in this case the autocorrelation has a larger support, the signal power is more concentrated around the carrier frequency, resulting in a narrower bandwidth. The bandwidth efficiency results increased but, as we will see later, high values for L result in a higher receiver complexity.

The properties of CPM have been the subject of several studies in the past. Some key advances have been performed after introducing new CPM signal representations, like the Laurent's decomposition (LD) [13], which is one of the most cited and exploited in the scientific literature. The LD writes a *binary* CPM as the sum of amplitude-modulated pulses each having distinct durations and shapes. As a result, a continuous-phase modulator can be implemented as a bank of pulse-amplitude modulators, each using a different pulse, whose outputs are added to form the continuous-phase modulated signal. The success of the LD during the early days of CPM is perhaps due to a characteristic that was frequently exploited for the design of reduced-complexity modulators and receivers. In fact, although the number of pulses resulting from the LD may be very large, an accurate approximation can be obtained using only a small subset of them, i.e., those with the largest energy. This result has been exploited in [14], where a low-complexity coherent receiver consisting of a bank of matched filters followed by a Viterbi processor has been proposed. In [15], it has been exploited to design low-complexity receivers for SCCPM systems. The LD has been extended to non binary CPM in [16]. In [17] it has been used to derive capacity bounds based on multiple-access channel theory.

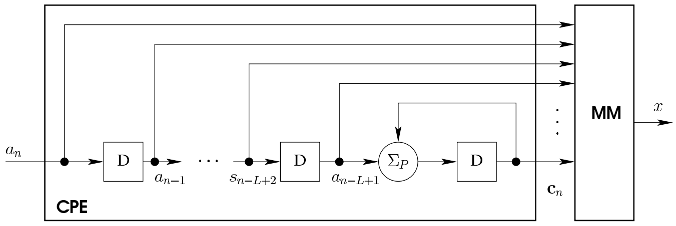


Fig. 2. Block diagram of the CPM encoder and modulator.

Another popular representation has been proposed in [18] and it is known as the Rimoldi decomposition (RD). According to this result, the CPM modulator can be decomposed into the cascade of a time-invariant convolutional encoder operating on rings of integers (the continuous-phase encoder, CPE) and a memoryless modulator (MM) (see Fig. 2). A relevant feature of this representation is the separation of the modulation process into two subprocesses, the first of which accounts for memory across adjacent symbols. In fact, the CPE is a device with memory, while the MM performs a *memoryless* mapping between symbols at the CPE output and complex waveforms whose support extends over one symbol interval. The CPE is a convolutional encoder operating on integer rings $\mathbb{Z}_M$ and $\mathbb{Z}_P$. Its trellis representation has $N_\sigma = PM^{L-1}$ states and $N_\epsilon = PM^L$ edges. At its output, vectors $\mathbf{c}_n \in (\mathbb{Z}_M)^L \times \mathbb{Z}_P$ are generated. Finally, the MM maps these vectors to the complex waveforms of the MM set Σ_{MM}.

The *maximum-likelihood* (ML) receiver for the modulator shown in Fig. 2 consists of a bank of filters matched to the MM signal space (the space spanned by the signals in Σ_{MM}) followed by a trellis processor matched to the CPE trellis. The matched filters provide to the trellis processor a sufficient statistics for the decoding of CPM signal, which is usually implemented using the Viterbi algorithm, for maximum-likelihood (ML) detection or the BCJR algorithm [19] for maximum a-posteriori (MAP) detection. In coded systems, the BCJR algorithm is usually executed within a SISO [20] block that computes the *extrinsic information* needed to implement an iterative turbo-like [21] receiver.

A comparison of the Laurent and Rimoldi results shows that both representations require a large set of waveforms, whose size grows exponentially with L but, while the Laurent waveforms have different energies, the Rimoldi waveforms have equal energy E_S. This characteristic made it difficult to use the RD for the design of low-complexity receivers for long time, until the Authors of [22] proposed an orthogonalization technique based on *principal components* (PC) and applied it to the MM waveforms. In brief, given a signal space with finite dimension like the one spanned by the MM waveforms, PC orthogonalization provides the orthonormal basis whose truncation to fewer dimensions results in minimum error energy. Applying the PC technique to the MM signal space significantly reduces the number of matched filters while the resulting degradation is negligible.

Fig. 3 shows the structure of a CPM receiver with filter bank designed according to the PC orthogonalization procedure. The filter outputs are sampled at instants $t = nT$ to

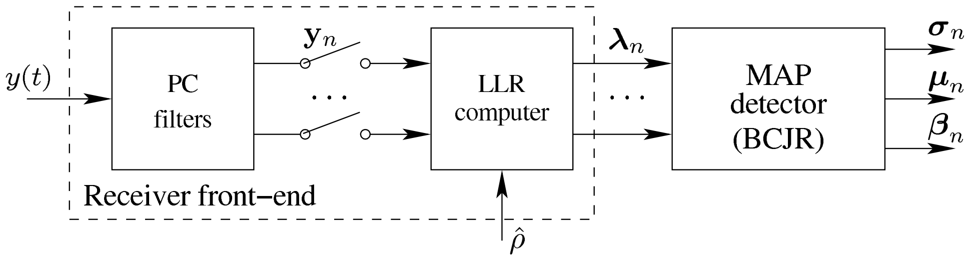


Fig. 3. Block diagram of the CPM receiver.

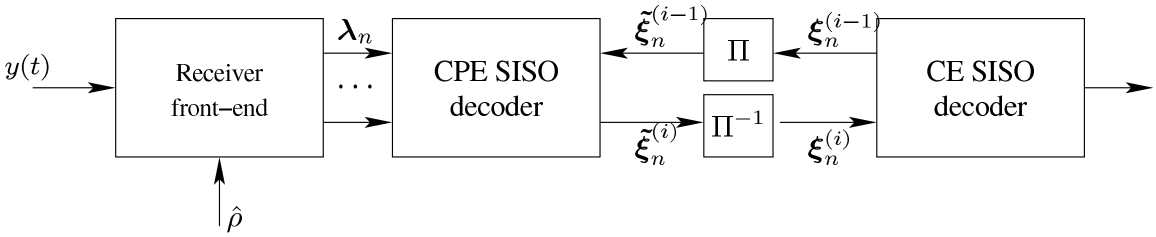


Fig. 4. Block diagram of the SCCPM iterative decoder.

obtain a sufficient statistics $\mathbf{y}_n \in \mathbb{C}^{N_{PC}}$ needed for ML or MAP decoding. Here, N_{PC} is the number of complex filters resulting from the PC orthogonalization. The log-likelihood ratios of the MM waveforms $s_i \in \Sigma_{MM}$ are then computed as

$$(\boldsymbol{\lambda}_n)_i = \log \frac{P(s_i|\mathbf{y}_n)}{P(s_0|\mathbf{y}_n)}$$

where $(\boldsymbol{\lambda}_n)_i$ indicates the ith element of vector $\boldsymbol{\lambda}_n$. In order to compute the LLRs, the receiver relies on the knowledge of an estimate $\hat{\rho}$ of the symbol energy to noise ratio $E_S/N_0 = \rho$. Clearly, when $\hat{\rho} = \rho$ the receiver provides optimal outputs, while a possible estimation error causes the receiver to operate in the so-called *mismatched* condition [23] which results in performance degradation.

The MAP detector computes the signal-wise, symbol-wise and bitwise *a-posteriori* LLRs

$$(\boldsymbol{\sigma}_n)_s = \log \frac{P(s_s|\mathbf{y})}{P(s_0|\mathbf{y})}, \quad s = 0, \ldots, |\Sigma_{\mathrm{MM}}| - 1 \tag{4}$$

$$(\boldsymbol{\mu}_n)_m = \log \frac{P(a_m|\mathbf{y})}{P(a_0|\mathbf{y})}, \quad m = 0, \ldots, M - 1 \tag{5}$$

$$(\boldsymbol{\beta}_n)_d = \log \frac{P(b_d = 1|\mathbf{y})}{P(b_d = 0|\mathbf{y})}, \quad d = 0, \ldots, \log_2(M) - 1 \tag{6}$$

where $\mathbf{y} = (\mathbf{y}_n)_{n=1}^{N}$ is the sequence of symbols at the sampled output of the PC filter bank and N is the observation length.

Recently, the RD with PC orthogonalization has been exploited for the design of coded CPM systems: a scheme consisting of an outer convolutional encoder (CE) whose coded bits enter an interleaver and then the CPE has been proposed [24]. This way, the outer CE and the CPE form a serial concatenation similar to what is known in the literature as a serially-concatenated convolutional encoder (SCCC) [25]. Iterative decoding can thus be performed, and it yields rather good performance as shown in [24], [26], [27]. Fig. 4 shows a scheme of the SCCPM iterative decoder. $\tilde{\boldsymbol{\xi}}^{(i)}$ is the *extrinsic information* computed at the ith iteration by the SISO block that operates on the CPE trellis.

III. System Model

We consider a system consisting of a set of $2K + 1$ users transmitting their coded CPM signals $x_k(t)$ to a base station or satellite system equipped with a multiuser receiver. The channel is affected by additive white Gaussian noise (AWGN) with two-sided power spectral density $G_n(f) = N_0/2$. To adequately model such *uplink* system, signals received from the K users are assumed to be asynchronous and not in phase. Moreover, since our goal is the assessment of the achievable spectral efficiency and error rate performance, we assume that power control is ideal, hence the received signals have equal energy.

The received signal model has the following complex baseband representation:

$$y(t) = \sum_{k=-K}^{K} x_k(t - \tau_k) e^{j(2\pi k \Delta_f t + \varphi_k)} + n(t) \tag{7}$$

where τ_k and φ_k are, respectively, the delay and phase affecting the kth user's signal. Moreover, Δ_f is the intercarrier frequency spacing and $n(t)$ is the additive white Gaussian noise.

The signal received from the kth user is a continuous phase, constant envelope modulated carrier

$$x_k(t) = \sqrt{\frac{2E_s}{T}} e^{j\psi_k(t)} \tag{8}$$

where E_s is the symbol energy, T is the symbol interval and

$$\psi_k(t) = 2\pi h \sum_{n=-\infty}^{\infty} a_{n,k} q(t - nT) \tag{9}$$

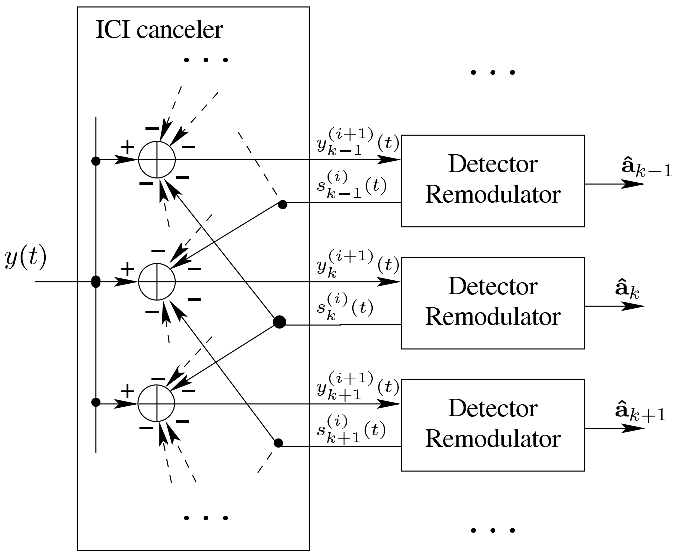


Fig. 5. Block diagram of the multiuser receiver for uncoded CPM systems.

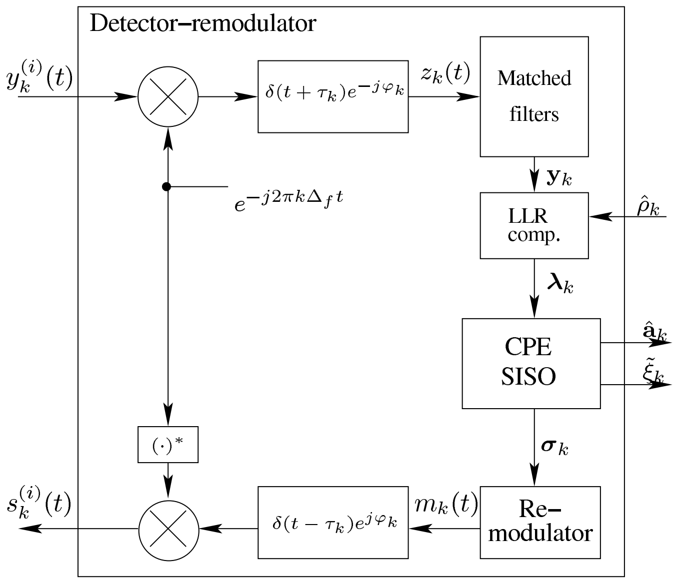


Fig. 6. Block diagram of the single-user detector-remodulator for user k.

The phase process $\psi_k(t)$ depends on the input information symbols $a_{n,k} \in \{\pm 1, \pm 3, \dots, \pm(M-1)\}$, where $M = 2^m$ is the size of the input alphabet.

We assume that all users adopt the same CPM scheme, hence the CPM parameters h and $q(t)$ do not depend on k. Although this assumption does not match a real scenario, it becomes reasonable in the absence of ICI resulting from perfect cancellation.

In the following, we will use the shorthand notation

$$x_k(t) = \mathcal{M}(\mathbf{a}_k)$$

where $\mathbf{a}_k$ denotes the symbol vector of user k and $\mathcal{M}(\cdot)$ denotes the CPM modulation function.

In [28], it has been observed that CPM schemes with rectangular (REC) frequency pulses exhibit good performance when used in FDM-CPM systems. Therefore, in this paper we will consider REC frequency pulses.

IV. The multiuser receiver

In the model defined in Sec. III, the capacity of the kth user link is determined by its signal energy to noise ratio E_S/N_0 and by the ICI. The capacity reduction due to ICI may be significant when, attempting to achieve higher spectral efficiency (SE), the intercarrier frequency spacing is set to low values. In [28], it has been shown that, in such case, the optimal frequency spacing that maximizes the SE when single-user receivers are employed depends on the parameters of the chosen CPM scheme.

The multiuser receiver we propose improves the system SE through iterative ICI cancellation. It consists of a bank of single-user receivers that iterate with an interference-cancellation block which performs simple linear operations on the received signal. The complexity of this receiver grows only linearly with the number of users. Moreover, as shown later, the proposed receiver exhibits significantly improved SE both for uncoded and coded CPM systems.

A scheme of the proposed receiver is shown in Fig. 5. Its main building block is a single-user MAP detector and remodulator (detailed in Fig. 6). Such detector computes the kth user's *a-posteriori signal* LLRs $\boldsymbol{\sigma}_{nk}$ as in (4). To compute the *a-posteriori* LLRs, the detector relies on the assumption that the signal being detected is corrupted by AWGN and on the knowledge $\hat{\rho}$ of the signal-to-noise ratio $\rho = E_s/N_0$.

Clearly, $\hat{\rho} = \rho$ yields the optimal MAP detector for a single-user communication (i.e., no ICI) through the AWGN channel, and bank of $2K+1$ single-user detectors yields optimal performance for $\Delta_f \to \infty$ since ICI becomes negligible in such case. In the latter case, however, the resulting SE becomes very low due to the large intercarrier frequency spacing. Since we are interested in *spectrally efficient* systems, we will consider the case of *tight* normalized intercarrier frequency spacing, i.e. $\Delta_f T \lessapprox 1$. As shown in [28], in such case the strong ICI significantly reduces the information rate with respect to a single-user system with no ICI. However, the low $\Delta_f T$ values compensate for the lower information rate, resulting in a possibly large SE.

The kth single-user detector and remodulator is fed with the following input signal at the ith iteration

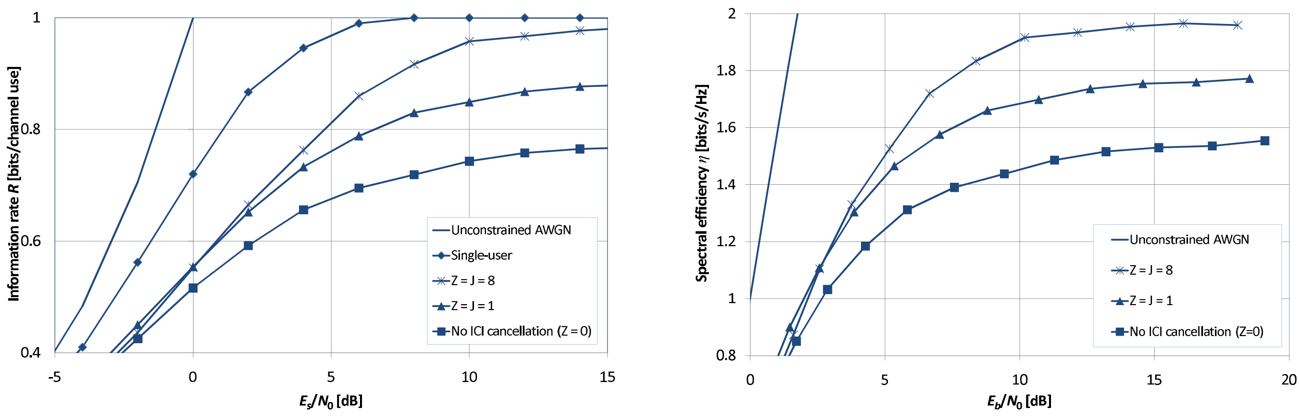
$$z_k^{(i)}(t) = y_k^{(i)}(t+\tau_k)e^{-j[2\pi k\Delta_f(t+\tau_k)+\varphi_k]} \tag{10}$$

where $y_k^{(i)}(t)$ is the received signal after interference cancellation, defined as

$$y_k^{(i)}(t) = y(t) - \sum_{j=-J, j\neq 0}^{J} s_{k+j}^{(i)}(t) \tag{11}$$

where J is a positive integer parameter related to the number of adjacent interfering signals being canceled. The signal $s_l^{(i)}(t)$ is

$$s_l^{(i)}(t) = m_l^{(i)}(t-\tau_l)e^{j[2\pi l\Delta_f(t-\tau_k)+\varphi_l]} \tag{12}$$



(a) Information rates (b) Spectral efficiencies

Fig. 7. Information rates of MSK systems (a) and spectral efficiencies achievable by coded MSK systems (b) with $\Delta_f T = 0.5$.

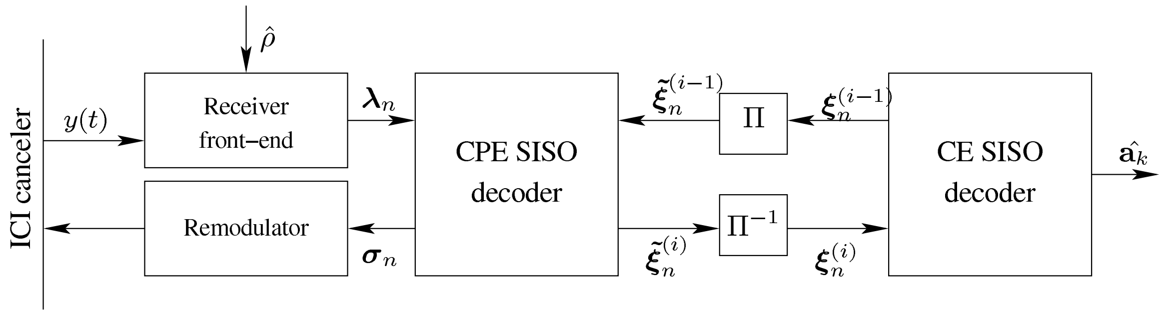


Fig. 8. Block diagram of the multiuser coded CPM receiver with ICI cancellation for user k. Here, $\hat{\mathbf{b}}_k$ is the estimated information word.

where $m_l^{(i)}(t)$ is the *re-modulated* signal for lth user at the ith iteration.

Several techniques can be applied to compute the remodulated signal. In [29], a soft technique has been proposed. It consists in the following steps: starting from the *a-posteriori signal* LLRs σ_{nk} of (4), it is possible to compute the *a-posteriori* probability distribution $\mathbf{p}_{nk}$ of the MM signal set Σ_{MM} for user k and symbol interval n as

$$\begin{aligned}
(\mathbf{p}_{n,k})_r &= P\{s_{n,k} = s_r|\mathbf{y}_k\} = e^{\sigma_{n,k}(r)} P\{s_{n,k} = s_0|\mathbf{y}_l\} \\
(\mathbf{p}_{n,k})_0 &= P\{s_{n,k} = s_0|\mathbf{y}_k\} = \left[\sum_{r=0}^{|\Sigma_{MM}|-1} e^{\sigma_{n,k}(r)}\right]^{-1}
\end{aligned}$$

The re-modulated signal is then computed as the *average signal*

$$m_l^{(i)}(t) = \sum_{n=-\infty}^{\infty} \sum_{r=0}^{|\Sigma|-1} (\mathbf{p}_{n,k})_r s_r(t - nT) \qquad (13)$$

where i denotes the iteration number.

Results show that this ICI cancellation technique improves the SE of uncoded CPM systems. As an example, we show in Fig.7 the information rates and spectral efficiencies achieved by an MSK system with normalized intercarrier frequency spacing $\Delta_f T = 0.5$. We note that ICI cancellation significantly improves the SE.

V. Coded CPM Multiuser Systems

Following the SCCPM approach [24], we connect an outer CE to the CPM modulator through an interleaver. A 4-state, rate 1/2 systematic recursive CE with generators $(7, 5)_8$ is adopted (see Fig. 10). Moreover, in order to achieve lower coding rates, the rate 1/2 CE is extended adding a further feedforward connection to obtain a rate 1/3 encoder. In this case, the generators are $(7, 5, 3)_8$.

The CE output is punctured using a rate-matching algorithm to achieve variable coding rates: the puncturing algorithm selects all systematic bits and some coded bits according to a regular pattern in order to achieve the desired overall rate.

The interleaver that connects the outer encoder to the CPM modulator is a spread (S-random) interleaver [30]. The spread parameter is set according to the code word length.

The adopted structure of channel decoder results in two iteration loops (see Fig. 8): an *inner* loop formed by the ICI canceler, the CPE SISO decoder, the receiver front-end and the remodulator, and an *outer* loop formed by the CPE SISO decoder, the CE SISO decoder, the interleaver (Π) and the deinterleaver (Π^{-1}).

The decoder starts decoding a received code word executing N_{IC} inner iterations. Then, it executes N_D outer iterations. Each outer iteration is followed by an inner iteration. This way, ICI cancellation is performed as part of the decoding iterations and it results in an improved ICI cancellation. In fact, as shown in Fig. 9, perfect cancellation and hence an

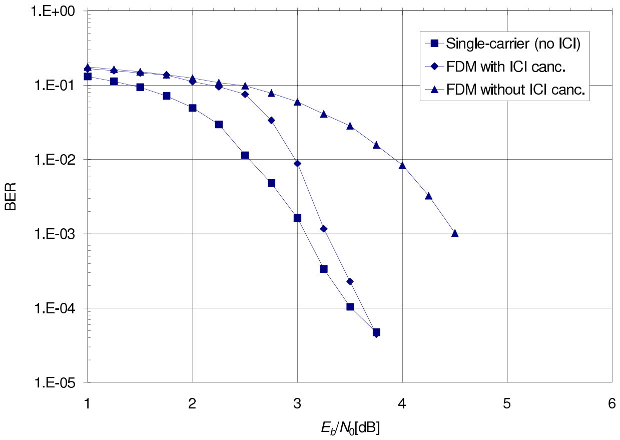


Fig. 9. Comparison of bit error rates of SC-CPM system M1-C4. Iterative decoding with 10 iterations is performed.

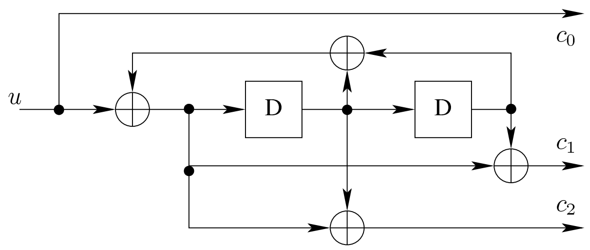


Fig. 10. Convolutional encoder.

almost single-carrier performance is achieved in some cases. The E_b/N_0 gain due to ICI cancellation is about 1.2 dB at BER = 10^{-3} and the curve is as close as 0.2 dB to the single-carrier BER curve.

VI. Choice of the Modulation and Coding Schemes

The choice of the modulation and coding (ModCod) schemes for an adaptive communication system is driven by the target SE requirements. Assuming a normalized inter-carrier frequency spacing $\Delta_f T \lessapprox 1$, then the size M of the CPM input alphabet determines the maximum achievable $SE = \log_2 M$ of the uncoded system . A quaternary scheme ($M = 4$) results therefore in a maximum achievable SE larger than 2 bits/s/Hz. Moreover, the channel encoder reduces the SE proportionally to its rate. As a result the *asymptotic* ($E_b/N_0 \to \infty$) SE is

$$\mathrm{SE}_\infty \triangleq \lim_{E_b/N_0 \to \infty} \mathrm{SE} = \frac{R_C \log_2 M}{\Delta_f T} \tag{14}$$

where $R_C = k_0/n_0$ is the rate of the punctured convolutional code.

The 4-state recursive CE of Fig. 10 features a systematic output (c_0) and two coded outputs (c_1 and c_2), therefore its rate is 1/3. Its output connection c_2 is punctured when coding rates larger than 1/2 are needed.

At the output of the CE, puncturing is applied to the coded bits. The rate of puncturing is defined as $R_P = N_P/D_P$.

TABLE I
CPM modes.

Mode	M	Q	P	L	Pulse
M1	2	1	2	1	REC
M2	4	1	6	1	REC

TABLE II
Coding modes.

Mode	k_0	n_0
C1	1	3
C2	1	2
C3	2	3
C4	5	6

TABLE III
Asymptotic spectral efficiencies.

	M1	M2
C1	0.5	1
C2	0.75	1.5
C3	1	2.
C4	1.25	2.5

Systematic bits are never eliminated, while coded bits are selectively eliminated according to a periodic pattern to achieve the desired rate.

As for the modulation scheme, *full-response* systems are preferred. Indeed, the use of partial-response systems is motivated by the improved spectral compactness needed in systems with tight bandwidth constraints when ICI is not canceled. Moreover, a large L results in a an increased receiver complexity. In our case, ICI is reduced through an appropriate cancellation algorithm. Moreover, a low CPM complexity compensates for the increased complexity due to ICI cancellation.

One binary and one quaternary CPM modulation scheme have been chosen. Their parameters are shown in Tab. I. The binary scheme is aimed at increasing the robustness of the ModCod set at low E_b/N_0, while the quaternary scheme provides the high number of signals needed to achieve a high SE when the SNR is larger. The binary CPM is a simple MSK scheme, while for the quaternary case a full response scheme with appropriate modulation index has been chosen.

As for the coding schemes, rate values from 3/8 to 5/6 have been considered. This range has been determined in order to achieve SE_∞ from 0.5 bits/s/Hz to 2.5 bits/s/Hz with an intercarrier frequency spacing $\Delta_f T = 2/3$. The parameters of each scheme are shown in Tab. II. Tab. III shows the achieved SE_∞ values with the chosen modulation and coding schemes.

VII. Results

The main results of this work are shown in Fig. 11 and Fig. 12. Bit error rates (BER) and frame error rates (FER) of the selected ModCod schemes are provided. A range of E_b/N_0 from 2 dB to 18 dB is spanned at BER = 10^{-4} with a spectral efficiency spanning the range from 0.5 bits/s/Hz to 2.5 bits/s/Hz. We observe that the curves corresponding to lower spectral efficiencies - those that use M1 as the modulation - exhibit a steep waterfall behavior, while those corresponding to higher spectral efficiencies, and hence modulation M2, exhibit

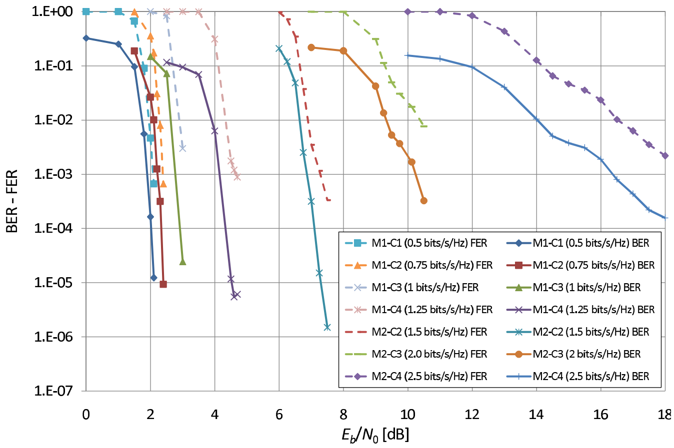


Fig. 11. Error rates of the selected coded modulation modes. Dashed lines refer to frame error rates (FER) and solid lines refer to bit error rates (BER).

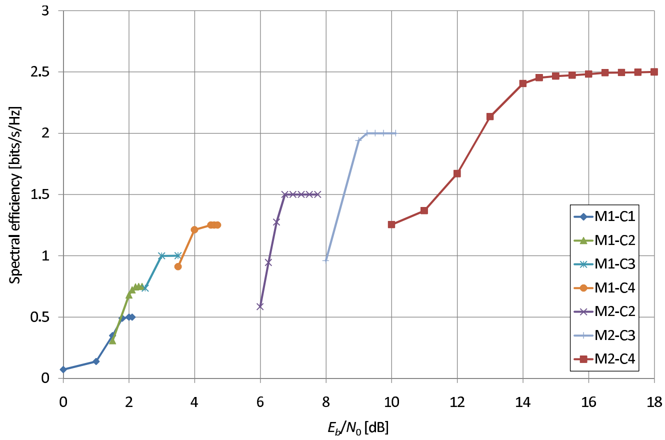


Fig. 12. Spectral efficiency of the selected coded modulation modes.

TABLE IV
RECEIVER PARAMETERS GIVEN AS (N_{IC}, J, N_D).

	M1	M2
C1	(1, 1, 10)	-
C2	(1, 1, 10)	(5, 4, 10)
C3	(1, 1, 10)	(8, 4, 10)
C4	(1, 1, 10)	(8, 4, 10)

a slower decrease with E_b/N_0. This degradation is mainly due to the much stronger ICI experienced when M2 is used. In such case, using a partial-response system would perhaps improve the performance, although resulting in an increased complexity. Another solution would be the increase of $\Delta_f T$, but this would reduced the SE.

Tab IV shows the parameter values used in the simulations. We compare the obtained results with those presented in [31], where SCCPM systems with SE ranging from 0.75 bits/s/Hz to 2.25 bits/s/Hz are designed. The outer codes used therein are convolutional codes with number of states from 4 to 64. Moreover, extended BCH codes and extended Hamming codes are also used as outer codes. In this case, a suboptimal Chase-Pyndiah decoding algorithm [32] is employed at the receiver. Moreover, the CPM schemes used therein are quaternary partial-response schemes with higher complexity.

The information word length used in [31] is 128 bytes = 1024 bits, a value very close to the length used in this paper, which is 1000 bits. Moreover, in [31] a single-user system is considered and the decoder performs 30 iterations, while our decoder performs only 10 iterations.

The best results reported in [31] show that, for a FER = 10^{-3}, SE = 0.75 bits/s/Hz is achieved at E_b/N_0 close to 2 dB and SE = 2.25 bits/s/Hz is achieved at E_b/N_0 close to 10.5 dB. Our scheme achieves SE = 0.75 bits/s/Hz at E_b/N_0 = 2.5 dB and SE = 2 bits/s/Hz at E_b/N_0 = 10.5 dB with FER = 10^{-2}.

Although our receiver operates in presence of strong ICI, its performance loss is only slight. Moreover, its complexity is lower since CPM schemes are simpler and the number of iterations is significantly lower.

VIII. CONCLUSION

We have shown that it is possible to design adaptive coded modulation systems employing CPM to mitigate the nonlinearity degradations of wireless transceivers. A high spectral efficiency can be achieved by performing a suitable design of the coding and modulation modes.

The proposed ModCod set spans a range of spectral efficiencies from 0.5 bits/s/Hz to 2.5 bits/s/Hz over a range of E_b/N_0 from 2 dB to 18 dB at BER = 10^{-4}. The adaptive coded modulation system includes an ICI cancellation algorithm, which improves the SE while resulting in a simplified adaptivity algorithm.

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Alberto Perotti (S'01-M'03) received the Laurea in Ingegneria delle Telecomunicazioni in 1999 and the Ph.D. degree in Electronics and Communications Engineering in 2003, both from Politecnico di Torino, Turin, Italy. In 2002, he spent seven months as a visiting scholar at the University of California, Los Angeles, working on the design of serially concatenated convolutional codes. Since 2003, he has been working as an associate researcher in the Electronics Department of Politecnico di Torino. His research interests are in the area of wireless communications. In particular, he is interested in channel coding and modulation techniques for wireless systems.

Piotr Remlein received the M.Sc. and the Ph.D. degrees from Poznan University of Technology (PUT), Poznan, Poland in 1991 and 2002 respectively. He has been employed at PUT since 1992, currently as an Assistant Professor at the Chair of Wireless Communications. His scientific interests cover wireless networks, communication theory, error control coding, cryptography, digital modulation, trellis coded modulation, continuous phase modulation, mobile communications, digital circuits design. Dr. Remlein is the author of more than 60 papers, which have been presented at national and international conferences, and published in the communication journals. He also acts as a reviewer for international and national conference and journal papers. He is the member of The Institute of Electrical and Electronics Engineers (IEEE) Communications Society and IEEE Information Theory Society.

Sergio Benedetto (M'76-SM'90-F'97) is a Full Professor of Digital Communications at Politecnico di Torino, Italy since 1981. He has been a Visiting Professor at University of California, Los Angeles (UCLA), at University of Canterbury, New Zealand, and is an Adjoint Professor at Ecole Nationale Superieure de Telecommunications in Paris. He has co-authored two books on probability and signal theory (in italian), the book "Digital Transmission Theory" (Prentice-Hall, 1987), "Optical Fiber Communications" (Artech House, 1996), and "Principles of Digital Communications with Wireless Applications" (Plenum-Kluwer, 1999), and over 250 papers in leading journals and conferences. He has taught several continuing education courses on the subject of channel coding for the UCLA Extension Program and for the CEI organisation. He has been Chairman of the Communications Theory Symposium of ICC 2001 and CTW2004, and has organized numerous sessions in major conferences worldwide. Sergio Benedetto has been for five years the Area Editor for the IEEE Transactions on Communications for Modulation and Signal Design, a Distinguished Lecturer of the IEEE Communications Society, and the Chairman of the Communication Theory Committee of IEEE and a Fellow of the IEEE. He is Vice President for Publications of the IEEE Communications Society. S. Benedetto is a Fellow of the IEEE, a member of the Science Academy of Torino, and has received the "Premio Siemens per le Telecomunicazioni" at the International Conference on Communications of Genova in 1973, the "Premio Bianchi" of AEI in 1974, the "Premio Bonavera dell'Accademia delle Scienze di Torino" in 1976, the "Gold medal Award of Siemens Telecomunicazioni" for the years 1993 and 1995, the "Italgas International Prize for Research and Technological Innovation" in 1998, the Cristoforo Colombo International Award for Communications in 2006, and the IEEE Communications Society Edwin Howard Armstrong in 2008.