CPM-Based Spread Spectrum Systems for Multi-user Communications

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***Abstract*—We propose a new spread spectrum (SS) system based on continuous-phase modulations (CPMs). The main idea is to exploit the sequence of modulation indices of a multi-h CPM as a frequency-hopping (FH) sequence. Spectral spreading, flatness and smoothness can be easily achieved by an appropriate choice of the maximum value of the modulation index and of the length of the index sequence. We will show that the proposed CPM-based spread spectrum system achieves an overall spectral efficiency larger than that of a single-user single-h CPM transmission even when a single-user detector is employed at the receiver. It also outperforms other solutions in the literature. In addition, we will also derive some suboptimal multi-user detectors.**

I. INTRODUCTION

Modern communications require modulation formats robust to nonlinearities and multiple-access interference (MAI), as well as power- and spectrally efficient. Robustness to nonlinearity is mandatory in order to use strongly saturated amplifiers, and spectral efficiency is one of the most important quality figures in any communication system. For this reason the choice of using modulation formats such as continuous-phase modulations (CPMs) comes quite naturally. CPMs are a family of very appealing modulation formats. Their robustness to nonlinearity stemming from the constant envelope is one of the main reasons of their popularity, along with excellent power and spectral efficiencies [1].

Code division multiple access (CDMA) is one of the most studied methods for multi-user communication systems. Based on the employed spread spectrum (SS) technique, CDMA schemes are grouped in two major classes, namely direct-sequence SS (DS-SS) and frequency-hopping SS (FH-SS).

DS-SS has been combined with CPMs in many different ways. Lane and Bush [2] proposed a SS multi-h CPM whose drawbacks in a multi-user scenario have already been analyzed in [3]. Giannetti *et al.* [4] studied a special subset of single-h binary CPMs, known as generalized minimum-shift keying (GenMSK), which can be approximately viewed as linear modulations. Hence, classical results of multi-user communications for linear modulations apply. Obviously, the main drawback of this approach is the strict constraint on the modulation format. Hsu and Lehnert [5] considered a multi-user system where each user transmits a SS signal that is the product between a linear modulation (for the data) and a multi-h CPM (for the spreading chips), giving up to phase continuity. This problem has been solved by Yang *et al.* [6] mapping the M-ary information symbols into M binary phase spreading sequences (PSSs) modulated by a single-h CPM modulator. The main problem of this approach is the time-consuming design of a unique set of M different and orthogonal PSSs for each user. Moreover, a simple receiver structure is not available because the data and the spreading chips are not separable. The separation between data and spreading chips has been preserved in the dual-phase technique proposed by McDowell *et al.* [7]. Chips are modulated as a multi-h CPM, data are modulated as a MSK signal, and finally multiplied. The receiver, as in the linearly-modulated DS-SS systems, is composed by an analog (and therefore expensive) despreader and a detector. Müller and Lampe proposed in [8] a DS-SS system using linear modulations with constant envelope and continuous phase. To avoid phase jumps to occur at every symbol change, they pose few constraints on the information symbol alphabet, the spreading factor, and the symbol waveform. This latter must depend on the chip sequence and the chip waveform. This solution, called continuous-phase chip modulation (CPCM), has nevertheless big spectral sidelobes, incompatible with spectral masks of most wireless communication standards. Therefore Müller recently proposed in [9] a linear DS-SS system where each user is assigned a set of very similar spreading sequences, which are chosen in a data-dependent fashion. These sequences are generated by an iterative algorithm ensuring their high stop-band attenuation, constant envelope and continuous phase.

To our knowledge, FH has never been studied as a multiple access technique in CPM-based systems. Nevertheless, FH has been used with the purpose of spreading the CPM power spectral density (PSD) for security issues in [10] and [11]. Here, a new multiple-access technique based on multi-h CPMs is proposed. The main idea is to exploit the fact that each CPM can be viewed as a frequency modulation where the frequency deviation is strictly related to the modulation index. Since in multi-h CPMs the modulation index is replaced by a sequence of indices (with the index varying every symbol period), the resulting effect is a sort of frequency hopping. This is exactly an instance of FH when applied to a continuous-phase FSK (CPFSK). So, we will use multi-h CPM not to improve bit error rate (BER) performance (as in [2], [5], and [7]) but to

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spread the PSD and allow multiple access without resorting to spreading codes or to any other DS-CDMA technique. In other words, we directly construct a modulation format with a PSD extremely flat, large and smooth at will. The corresponding single-user detector has practically the same complexity of a classical single-h CPM detector with the same number of phase states (which is a clear advantage if compared to the complexity of the receivers in [6] and [7]). In the CPM literature, the modulation index is hardly ever chosen bigger than one (except for [12] where satellite navigation systems have been addressed), even though this would not invalidate the CPM definition. Therefore, the most natural way to spread the CPM power spectral density is by using indices much bigger than one and varying in a wide range [13]. Moreover, using a long sequence of indices the CPM power spectral density will become smoother. Assigning to each user a different and randomly generated sequence of indices, we will obtain a new and efficient FH spread spectrum technique for CPM-based systems. With this approach, we will get rid of the constraints on the modulation formats (since we consider general M-ary multi-h partial response CPMs). Obviously the phase continuity and the constant envelope are guaranteed. The spreading factor, usually defined in DS-CDMA systems with linear modulations as the ratio between the bandwidth of the spread signal and the bandwidth of the signal before spreading, cannot be defined in the same way here because in the proposed system there is no "signal before spreading"—the spectral spreading effect is now embedded in the modulation format itself. On the other hand, the definition of spreading factor proposed in [14], as the ratio of the Fourier bandwidth of the spread signal to its Shannon bandwidth, could be used. However, it requires the computation of an orthonormal basis for the spread signal, not available here in closed form.

Since we are considering a multi-user scenario, we also address the multi-user detection (MUD) issue. Because the complexity of the optimal multi-user receiver grows exponentially with the number of users, suboptimal detection schemes are required. We consider different multi-user detectors, based on hard interference cancellation (HIC) [15], soft interference cancellation (SIC) [16], extended to frequency division multiplexed CPM-based systems in [17], and an algorithm derived in [17] by using factor graphs (FGs) and the sum-product algorithm (SPA) framework [18].

II. System model

A. CPM signal

The complex envelope of a generic multi-h CPM signal is [19]

$$s(t) = \sqrt{\frac{2E_s}{T}} \exp\left\{ \jmath \left[2\pi \sum_i h_{\underline{i}} \alpha_i q(t-iT) + \theta \right] \right\} \quad (1)$$

where E_s is the energy per symbol, T is the symbol period, $\{\alpha_i\}$ are the M-ary information symbols, $\{h_i\}$ is the sequence of N_h modulation indices, $\underline{i} = i \bmod N_h$, $q(t)$ is the phase-smoothing response characterizing the format, and θ is an initial phase offset. The phase-smoothing response is a continuous function satisfying the following property:

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

L being the correlation length of the signal. The frequency pulse is defined as

$$p(t) = \frac{\mathrm{d}}{\mathrm{d}t} q(t)$$

and (1) can be rewritten as

$$s(t) = \sqrt{\frac{2E_s}{T}} \exp\left\{ \jmath \left[2\pi \int_{-\infty}^{t} \sum_i h_{\underline{i}} \alpha_i p(\tau - iT) \mathrm{d}\tau + \theta \right] \right\} \quad (2)$$

which is the expression of a frequency-modulated signal using a pulse amplitude modulation (PAM) with shaping pulse $p(t)$ as modulating signal. The most used frequency pulses are the rectangular pulse (L-REC to denote its duration of L symbol periods) and the raised-cosine pulse (L-RC).

CPMs are modulations with memory. In the generic symbol interval $nT \le t < (n+1)T$, the CPM signal (1) is completely defined by symbol α_n and state $\sigma_n = (\omega_n, \varphi_n)$, where $\omega_n = (\alpha_{n-1}, \alpha_{n-2}, \ldots, \alpha_{n-L+1})$ is the correlative state and

$$\begin{aligned} \varphi_n &= \left(\pi \sum_{i=-\infty}^{n-L} h_{\underline{i}} \alpha_i \right) \bmod 2\pi \\ &= (\varphi_{n-1} + \pi h_{\underline{n-L}} \alpha_{n-L}) \bmod 2\pi \end{aligned} \quad (3)$$

is the phase state [20], [21]. The correlative state can assume M^{L-1} values, whereas the phase state can assume p values, having defined $h_i = k_i/p$ where k_i and p are positive integer numbers[1] and integer values for h_i are forbidden. Therefore, the total number of states is pM^{L-1}. The CPM signal in the symbol interval $nT \le t < (n+1)T$ can thus be expressed as

$$\begin{aligned} s(t) &= \sqrt{\frac{2E_s}{T}} e^{\jmath(\varphi_n + \theta)} \exp\left\{ \jmath 2\pi \sum_{i=0}^{L-1} h_{\underline{n-i}} \alpha_{n-i} q(t - nT + iT) \right\} \\ &= \sqrt{\frac{2E_s}{T}} e^{\jmath(\varphi_n + \theta)} \prod_{i=0}^{L-1} \left[\exp\left\{ \jmath \frac{2\pi}{p} q(t - nT + iT) \right\} \right]^{k_{\underline{n-i}} \alpha_{n-i}} \end{aligned} \quad (4)$$

B. SS-FH-CPM

In the proposed multi-user system, multiple access is guaranteed by assigning a different sequence of modulation indices to each user. We assume that each user transmits K symbols, and we denote by $\alpha_n^{(u)}$ and $\sigma_n^{(u)}$ the symbol transmitted by user u at discrete-time n and the corresponding state. We define $\boldsymbol{\alpha}^{(u)} = (\alpha_0^{(u)}, \ldots, \alpha_{K-1}^{(u)})^T$ as the vector of the K symbols transmitted by user u, and also $\boldsymbol{\alpha}_n = (\alpha_n^{(1)}, \ldots, \alpha_n^{(U)})^T$ as the vector of all symbols transmitted at discrete-time n (one

[1] A correct definition of the modulation index requires that k_i and p are relative prime to have a minimal trellis representation. As it will be clear later, the considered sequence of indices is chosen such that p is kept constant whereas k_i is chosen randomly with the only constraint that h_i cannot be integer. When k_i and p are not relative prime, we still use, for simplicity, a trellis representation with p states although it could be reduced. This allows to always use the same trellis without the need to resort to a time-varying trellis.

symbol per user), and $\boldsymbol{\alpha} = (\boldsymbol{\alpha}_0^T, \ldots, \boldsymbol{\alpha}_{K-1}^T)^T$, where $(.)^T$ denotes transpose. Similarly, we define $\boldsymbol{\sigma}_n = (\sigma_n^{(1)}, \ldots, \sigma_n^{(U)})^T$ and $\boldsymbol{\sigma} = (\boldsymbol{\sigma}_0^T, \ldots, \boldsymbol{\sigma}_{K-1}^T)^T$. We also define

$$s^{(u)}(\boldsymbol{\alpha}^{(u)}, t) = \sqrt{\frac{2E_s^{(u)}}{T}} \exp\left\{ \jmath 2\pi \sum_{i=0}^{K-1} h_{\underline{i}}^{(u)} \alpha_i^{(u)} q(t-iT) + \jmath\theta^{(u)} \right\} \quad (5)$$

the signal transmitted by user u and, without loss of generality, we assume that all users employ the same values of T, M, L, p, N_h and $h_{\max}$, $h_{\max}$ being the maximum value taken by the modulation index. We will also assume that all users employ the same phase-smoothing response $q(t)$.

We consider an asynchronous multiple-access system on an additive white Gaussian noise (AWGN) channel, so that the complex envelope of the received signal is

$$\begin{aligned} r(t) &= \sum_{u=1}^{U} s^{(u)}(\boldsymbol{\alpha}^{(u)}, t-\tau^{(u)}) + w(t) \\ &= s^{(\ell)}(\boldsymbol{\alpha}^{(\ell)}, t-\tau^{(\ell)}) + \sum_{\substack{u=1 \\ u\neq\ell}}^{U} s^{(u)}(\boldsymbol{\alpha}^{(u)}, t-\tau^{(u)}) + w(t) \,. \end{aligned} \quad (6)$$

Initial phase offsets $\theta^{(u)}$ and delays $\tau^{(u)}$ are random variables uniformly distributed in $[0, 2\pi)$ and $[0, T)$, respectively. For user ℓ, the reference user, without loss of generality we will assume $\theta^{(\ell)} = \tau^{(\ell)} = 0$. The thermal noise is a zero-mean circularly symmetric white Gaussian process with PSD $2N_0$.

Fixing the indices denominator p is mandatory to keep constant the number of the phase states, while fixing the maximum numerator allows every user to undergo the same spectral spreading. Each user has a different sequence of randomly-generated modulation indices. The spectral spreading depends only on the range of values assumed by the modulation index—the larger this range, the stronger the spreading effect. The number of modulation indices N_h plays a role only in the smoothness of the PSD. A CPM with high N_h will show a smooth PSD with small oscillations and no sidelobes (see the numerical results in Section IV).

The number of users allowed in the system depends on the total number of possible indices $\nu = ph_{\max} - \lfloor h_{\max} \rfloor$ (where $\lfloor x \rfloor$ denotes the maximum integer lower than x). If we impose the absence of overlaps, in a synchronous system the maximum number of users would coincide with the number of possible indices

$$U_{\max} = \nu \,.$$

C. Multi-user Detectors

Although not necessary in the derivation of the algorithms, since it applies unmodified independently of the employed set of sufficient statistics, we will adopt, as in practical receiver implementations, an approximated set of sufficient statistics for MAP symbol detection obtained as described in [22]. We assume the useful signal component to be band-limited with bandwidth lower than $N/2T$, where N is a proper positive integer. Although this is obviously an approximation in the case of CPM signals, whose PSD has, strictly speaking, an infinite support, the choice of a proper value of N ensures that this approximation can be made good at will. The approximated statistics can be obtained by extracting N samples per symbol interval from the received signal (6) prefiltered by means of a low-pass filter which leaves unmodified the useful signal and has a vestigial symmetry around $N/2T$ [22]. The condition on the vestigial symmetry ensures that the noise samples are independent and identically distributed complex Gaussian random variables with independent components, each with mean zero and variance $\xi^2 = N_0 N/T$ [22]. An alternative (and not approximated) set of sufficient statistics can be obtained as described in [17]. We denote by $r_{n,m}$ the m-th received sample (with $m = 0, \ldots, N-1$) of the n-th symbol interval. It can be expressed as

$$r_{n,m} = \sum_{u=1}^{U} s_{n,m}^{(u)}(\alpha_n^{(u)}, \sigma_n^{(u)}) + w_{n,m} \quad (7)$$

where, as mentioned, $\{w_{n,m}\}$ are independent and identically distributed complex Gaussian noise samples and $s_{n,m}^{(u)}(\alpha_n^{(u)}, \sigma_n^{(u)})$ (whose dependence on $\alpha_n^{(u)}$ and $\sigma_n^{(u)}$ will be omitted in the following) is the contribution of user u to the useful signal component. In the following, we will denote by $\mathbf{r}_n = (r_{n,0}, r_{n,1}, \ldots, r_{n,N-1})^T$ the vector of the received samples in the n-th symbol interval, by $\mathbf{r} = (\mathbf{r}_0^T, \mathbf{r}_1^T, \ldots, \mathbf{r}_{K-1}^T)^T$ the vector of all the received samples, and by $\mathbf{s}_n^{(u)} = (s_{n,0}^{(u)}, s_{n,1}^{(u)}, \ldots, s_{n,N-1}^{(u)})^T$ the vector collecting the samples of the signal of user u in the n-th symbol interval.

When considering coded CPM schemes where the CPM modulator is concatenated, possibly through an interleaver, with an outer encoder (as an example, see [23], [24] and references therein), the receiver is usually based on a soft-input soft-output (SISO) detector that iteratively exchanges soft information with the outer SISO decoder according to the turbo principle. Regarding single-user SISO CPM detection, little can be added to what already said in the literature (as an example, see [25] and references therein)—the adoption of multi-h CPM signals here entails only trivial modifications with respect to the case of single-h CPMs or the adoption, in case of simplified detectors, of the Laurent decomposition extended to multi-h signals [26]. As far as the optimal multi-user detector (MUD) is concerned, it has a complexity which is exponential in the number of users U and is thus infeasible.[2] Suboptimal multi-user SISO CPM detectors can also be conceived by extending those described in [17] for frequency-division-multiplexed CPM systems.

1) HIC-based receiver: The most trivial multi-user detector is that based on HIC [15]. The receiver for each user is composed by a SISO SUD, a SISO decoder, an encoder and a modulator. The SUD receiver for user u estimates its own information bits through a proper number of iterations of the soft detector and the soft decoder. If the estimated bits form a valid codeword, this is re-encoded and re-modulated. The resulting signal is then passed to the SUD detectors of all other users to allow the interference cancellation. Then,

[2]For its derivation, the reader can refer to [17, Section III.A]. In fact, although [17] deals with CPM-based frequency-division-multiplexed systems, the derivation holds unmodified in the case of SS-FH-CPM systems.

this process of iterative soft detection/decoding, interference estimation and cancellation is iterated until a valid codeword cannot be decoded.

2) SIC-based receivers: One of the reduced-complexity SIC algorithms with a very good performance available in the CDMA literature is that proposed in [16]. Being based on a Gaussian approximation of the MAI, the algorithm can be obtained by replacing the probability mass function (PMF) of the interfering symbols with a complex circularly symmetric Gaussian probability density function (PDF) with the same mean and variance. In the following, we will denote by $P(.)$ (respectively, $p(.)$) the PMF (respectively, the PDF) of a discrete (respectively, continuous) random vector.

Users will employ a SISO SUD each, and will exchange soft information to cancel out the interference. For the sake of simplicity, let us consider U synchronous users. We assume the discrete-time equivalent channel for user ℓ to be

$$r_{n,m}^{(\ell)} = s_{n,m}^{(\ell)} + z_{n,m}^{(\ell)}$$

where $z_{n,m}^{(\ell)}$ accounts for both interference and noise, that is

$$z_{n,m}^{(\ell)} = \sum_{\substack{u=1\\u\neq\ell}}^{U} s_{n,m}^{(u)} + w_{n,m}\,.$$

The vector $\mathbf{z}_n^{(\ell)} = (z_{n,0}^{(\ell)}, \ldots, z_{n,N-1}^{(\ell)})^T$ is assumed Gaussian with mean vector $\boldsymbol{\mu}_n^{(\ell)}$ and covariance matrix $\boldsymbol{\Phi}_n^{(\ell)}$, respectively, defined as

$$\boldsymbol{\mu}_n^{(\ell)} = \sum_{\substack{u=1\\u\neq\ell}}^{U} \bar{\boldsymbol{\mu}}_n^{(u)} \tag{8}$$

$$\bar{\boldsymbol{\mu}}_n^{(u)} = \sum_{(\alpha_n^{(u)},\sigma_n^{(u)})} \hat{P}(\alpha_n^{(u)}, \sigma_n^{(u)}|\mathbf{r})\mathbf{s}_n^{(u)} \tag{9}$$

$$\begin{aligned}\boldsymbol{\Phi}_n^{(\ell)} =& \sum_{\substack{u=1\\u\neq\ell}}^{U} \sum_{(\alpha_n^{(u)},\sigma_n^{(u)})} \hat{P}(\alpha_n^{(u)}, \sigma_n^{(u)}|\mathbf{r})(\mathbf{s}_n^{(u)} - \bar{\boldsymbol{\mu}}_n^{(u)})(\mathbf{s}_n^{(u)} - \bar{\boldsymbol{\mu}}_n^{(u)})^H \\ &+ 2\xi^2\mathbf{I}\end{aligned} \tag{10}$$

where $\mathbf{I}$ is the identity matrix, $(.)^H$ denotes conjugate transpose, and $\{\hat{P}(\alpha_n^{(u)}, \sigma_n^{(u)}|\mathbf{r})\}$ are the estimates of the a posteriori probabilities provided by the single-user SISO detector related to the interfering user u. The SISO detector for user ℓ, in the form of a BCJR algorithm [27], will employ the following branch metric (dependencies are omitted for the sake of notational convenience)

$$G_n^{(\ell)} \propto \exp\left\{2\Re\left[\mathbf{s}_n^{(\ell)H}\boldsymbol{\Phi}_n^{(\ell)-1}(\mathbf{r} - \boldsymbol{\mu}_n^{(\ell)})\right] - \mathbf{s}_n^{(\ell)H}\boldsymbol{\Phi}_n^{(\ell)-1}\mathbf{s}_n^{(\ell)}\right\} \tag{11}$$

where $\Re[.]$ stands for the real part operator and $\propto$ denotes a proportionality relation. Denoting by $I_n^{(\ell)}(\sigma_{n+1}^{(\ell)}, \sigma_n^{(\ell)}, \alpha_n^{(\ell)})$ the indicator function equal to one if $\alpha_n^{(\ell)}$, $\sigma_n^{(\ell)}$ and $\sigma_{n+1}^{(\ell)}$ satisfy the trellis constraint for user ℓ, and equal to zero otherwise, we define

$$C_n^{(\ell)}(\sigma_{n+1}^{(\ell)}, \sigma_n^{(\ell)}, \alpha_n^{(\ell)}) = I_n^{(\ell)}(\sigma_{n+1}^{(\ell)}, \sigma_n^{(\ell)}, \alpha_n^{(\ell)})P(\alpha_n^{(\ell)})\,.$$

The outputs of the SISO detector are the estimates of the a posteriori probabilities needed by the other users' SISO detectors to perform soft cancellation:

$$\hat{P}(\alpha_n^{(\ell)}, \sigma_n^{(\ell)}|\mathbf{r}) \propto A_n(\sigma_n^{(\ell)})B_{n+1}(\sigma_{n+1}^{(\ell)})G_n^{(\ell)}C_n^{(\ell)} \tag{12}$$

where $A_n(\sigma_n^{(\ell)})$ and $B_n(\sigma_n^{(\ell)})$ are the forward and backward messages of the BCJR algorithm.

The SIC MUD is then formed by U enhanced SISO SUDs, each of which computes the mean vector $\boldsymbol{\mu}_n^{(\ell)}$ and the covariance matrix $\boldsymbol{\Phi}_n^{(\ell)}$ for every symbol interval through (8) and (10), inverts $\boldsymbol{\Phi}_n^{(\ell)}$ and then computes the branch metric in (11). Finally, it computes the a posteriori probabilities $\{\hat{P}(\alpha_n^{(\ell)}, \sigma_n^{(\ell)}|\mathbf{r})\}$ with (12) and passes them to all the other SISO detectors for soft cancellation. In the following, this algorithm will be referred to as SIC 1. Its complexity is quadratic in the number of users [16].

This algorithm can be simplified by neglecting the off-diagonal elements of $\boldsymbol{\Phi}_n^{(\ell)}$ [16]. Consequently, the matrix inversion results to be computationally less expensive at the price of a performance degradation. This simplified detector will be referred to as SIC 2 and has a complexity that linearly depends on the number of users.

3) FG-based receiver: This algorithm, proposed in [17] for FDM-CPM systems and based on the application of the FG/SPA framework, derives from a suitable factorization of the PMF $P(\boldsymbol{\alpha}, \boldsymbol{\sigma}|\mathbf{r})$:

$$P(\boldsymbol{\alpha}, \boldsymbol{\sigma}|\mathbf{r}) \propto p(\mathbf{r}|\boldsymbol{\alpha}, \boldsymbol{\sigma})P(\boldsymbol{\sigma}|\boldsymbol{\alpha})P(\boldsymbol{\alpha})\,.$$

Each term can be further factored as follows:

$$\begin{aligned} P(\boldsymbol{\alpha}) &= \prod_{u=1}^{U}\prod_{n=0}^{K-1} P(\alpha_n^{(u)}) \\ P(\boldsymbol{\sigma}|\boldsymbol{\alpha}) &= \prod_{u=1}^{U} P(\sigma_0^{(u)}) \prod_{n=0}^{K-1} P(\sigma_{n+1}^{(u)}|\alpha_n^{(u)}, \sigma_n^{(u)}) \\ p(\mathbf{r}|\boldsymbol{\alpha}, \boldsymbol{\sigma}) &\propto \prod_{n=0}^{K-1} F_n(\boldsymbol{\alpha}_n, \boldsymbol{\sigma}_n) \prod_{u=1}^{U} H_n^{(u)}(\alpha_n^{(u)}, \sigma_n^{(u)}) \end{aligned}$$

where

$$\begin{aligned} P(\sigma_{n+1}^{(u)}|\alpha_n^{(u)}, \sigma_n^{(u)}) &\propto I_n^{(u)}(\sigma_{n+1}^{(u)}, \sigma_n^{(u)}, \alpha_n^{(u)}) \\ F_n(\boldsymbol{\alpha}_n, \boldsymbol{\sigma}_n) &= \prod_{i=1}^{U-1}\prod_{j=i+1}^{U} \exp\left\{-\frac{1}{\xi^2}\Re\left[\mathbf{s}_n^{(i)H}\mathbf{s}_n^{(j)}\right]\right\} \\ H_n^{(u)}(\alpha_n^{(u)}, \sigma_n^{(u)}) &= \exp\left\{\frac{1}{\xi^2}\Re\left[\mathbf{r}_n^H\mathbf{s}_n^{(u)}\right]\right\}\,. \end{aligned}$$

Hence, we finally have

$$\begin{aligned} P(\boldsymbol{\alpha}, \boldsymbol{\sigma}|\mathbf{r}) \propto& \left[\prod_{u=1}^{U} P(\sigma_0^{(u)})\right] \prod_{n=0}^{K-1} F_n(\boldsymbol{\alpha}_n, \boldsymbol{\sigma}_n) \\ &\cdot \prod_{u=1}^{U} H_n^{(u)}(\alpha_n^{(u)}, \sigma_n^{(u)}) I_n^{(u)}(\sigma_{n+1}^{(u)}, \sigma_n^{(u)}, \alpha_n^{(u)}) P(\alpha_n^{(u)})\,. \end{aligned} \tag{13}$$

The resulting FG, not shown here for a lack of space, has cycles of length four. As known, the application of the SPA

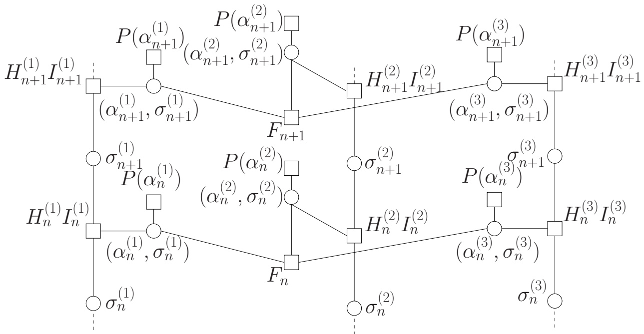

Figure 1. FG corresponding to (13) after stretching variables $\sigma_n^{(u)}$ in $(\alpha_n^{(u)}, \sigma_n^{(u)})$ and for $U = 3$. Circles and squares represent variable and function nodes, respectively.

to a FG with cycles allows an approximate (because of the presence of cycles) computation of the a posteriori probabilities $\{P(\alpha_n^{(u)}|\mathbf{r})\}$ required for the implementation of the MAP symbol detection strategy [18]. However, the presence of shortest cycles of length four makes the convergence of the SPA to good approximations of the a posteriori probabilities $\{P(\alpha_n^{(u)}|\mathbf{r})\}$ very unlikely [18]. It is possible to remove these short cycles by stretching [18] the variables $\sigma_n^{(u)}$ in $(\alpha_n^{(u)}, \sigma_n^{(u)})$. In other words, instead of representing variables $\alpha_n^{(u)}$ alone, we define a new variable given by the couple $(\alpha_n^{(u)}, \sigma_n^{(u)})$. This transformation does not involve approximations, since the resulting graph preserves all the information of the original graph. The resulting FG, shown in Fig. 1, has cycles of length twelve. Since cycles are still present, the SPA applied to this graph is iterative and still leads to an approximate computation of the a posteriori probabilities $\{P(\alpha_n^{(u)}|\mathbf{r})\}$ [18]. However, the absence of short cycles allows us to obtain very good approximations, as shown later. As the SIC 2, this algorithm has a complexity which is linear in the number of users [17].

D. Complexity considerations

With respect to the optimal detector for a single-h CPM signal, the SUD for a SS-FH-CPM signal has the same number of states (provided that the values of p, M, and L are the same) and the same number of trellis branches. In order to evaluate the branch metrics, we need the N samples $\mathbf{s}_n^{(u)}$ of all the possible waveforms that can be transmitted in a symbol period. These samples will be then correlated with the received samples in a given symbol period, i.e., the product $\mathbf{r}_n^H \mathbf{s}_n^{(u)}$ has to be computed. For a single-h signal, these waveforms are M^L and can be precomputed and stored in a look-up table (LUT). On the other hand, for a SS-FH-CPM signal the number of possible waveforms also depends on the possible L-tuple of consecutive modulation indices in the sequence of N_h modulation indices adopted by the considered user, which are $\min\left\{N_h, \binom{\nu}{L}\right\} M^L$, although not all are employed in the same trellis section. If this number is too high, it could not be convenient to store them but could be preferable to precompute and store the samples of the L waveforms $\left\{\exp\left[\jmath \frac{2\pi}{p} q(t-iT)\right]\right\}$, $i = 0, 1, \ldots, L-1$, in (4) and then use them to compute the needed waveforms in each symbol period. The same waveforms are also required to be computed every symbol epoch or precomputed and stored for the implementation of all MUDs as well.

With respect to traditional DS-SS systems based on linear modulations, a much larger number of correlations has to be computed. This is the price to be paid to have signals with constant envelope (and large spectral efficiency, as shown later). However, we point out that a significant complexity reduction can be obtained by extending the technique described in [25] for single-h CPM signals to the case of multi-h signals using the decomposition in [26] that allows to express a multi-h signal as a sum of linearly-modulated components. In this case, the number of trellis states of the SUD is reduced to p and also the branch metrics computation results to be greatly simplified.

III. Spectral efficiency

The main quality figure we consider in this work is the overall spectral efficiency η_U of the system. Since we are considering a multiple-access scenario where all users share the same bandwidth, the most intuitive way to compute η_U is to evaluate the spectral efficiency η of a reference user, and then define $\eta_U = U\eta$.

The spectral efficiency for the reference user can be computed as

$$\eta = \frac{I}{BT} \quad \text{[bit/s/Hz]} \tag{14}$$

where B is the bandwidth occupied by the CPM signal and I is the information rate of the user. CPM bandwidth is theoretically infinite because the PSD of a CPM signal has rigorously an infinite support. Hence, we consider the traditional definition of bandwidth based on the power concentration, that is the bandwidth that contains a given fraction of the overall power. Being this fraction a parameter, we choose to use the 99.9% of the overall power. This definition is coherent with systems where a limitation on the out-of-band power exists. To compute this bandwidth we need the CPM power spectral density, which cannot be evaluated analytically in closed form but only numerically. The adopted algorithm is the one proposed in [28] and [29].

To compute the information rate I for the reference user, we can use the simulation-based technique described in [30], which only requires the existence of an optimal MAP symbol detector for the considered system. Unfortunately, the complexity of the optimal MUD is exponential in U, making the evaluation of I practically infeasible. Therefore, we can evaluate an achievable lower bound by resorting to the concept of mismatched detection [31]. We can consider an approximated channel model (the auxiliary channel) for which an exact MAP symbol detection with affordable complexity exists—the more similar the auxiliary channel to the actual channel, the tighter the obtained bound on the spectral efficiency.

As done in [1], we approximate the channel model at the receiver side by modeling the interference as a zero-mean circularly symmetric white Gaussian process with PSD $2N_I$, N_I being a design parameter independent of the thermal noise.

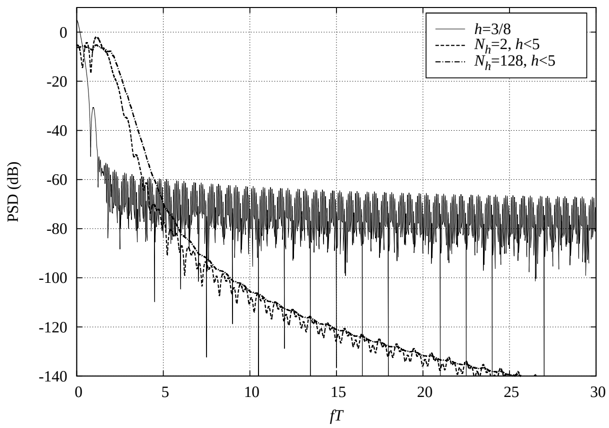

Figure 2. Power spectral densities for different single-h and SS-FH-CPM signals.

This approximation is exploited only by the receiver, while in the actual channel the interference is generated as in (6). Hence, the considered auxiliary channel model is that for which the received signal reads

$$r(t) = s^{(\ell)}(t) + \zeta(t) \tag{15}$$

where $\zeta(t)$ is a zero-mean circularly symmetric white Gaussian process with PSD $2(N_0 + N_I)$. The simulation-based method described in [30] allows to evaluate the achievable information rate for the mismatched receiver, i. e.,

$$I(\boldsymbol{\alpha}^{(\ell)}, \mathbf{r}) = \lim_{J\to\infty} \frac{1}{J} \mathrm{E}\left\{\log \frac{p(\mathbf{r}^J|\boldsymbol{\alpha}^{(\ell)J})}{p(\mathbf{r}^J)}\right\} \quad [\text{bit/ch.use}] \tag{16}$$

where we used the superscript J to remark that a sequence is truncated to its first J elements. In (16) $p(\mathbf{r}^J|\boldsymbol{\alpha}^{(\ell)J})$ and $p(\mathbf{r}^J)$ are probability density functions according to the auxiliary channel model (15), while the statistical average is with respect to the input and the output sequences evaluated according to the actual channel model (6). Both $p(\mathbf{r}^J|\boldsymbol{\alpha}^{(\ell)J})$ and $p(\mathbf{r}^J)$ can be evaluated recursively through the forward recursion of the MAP detection algorithm matched to the auxiliary channel model [30]. The mismatched receiver can assure communication with arbitrarily small nonzero error probability when the transmission rate at the CPM modulator input does not exceed $I(\boldsymbol{\alpha}^{(\ell)}, \mathbf{r})$ bits per channel use.

IV. Numerical results

A. Power spectral density

In order to describe the spectral behavior of the proposed system, we consider three different binary CPM signals using the 2-RC pulse and show their PSDs in Fig. 2, computed by using the technique described in [28], [29]. The first signal is a single-h signal with $h = 3/8$. The remaining ones are SS-FH-CPM signals with $h < 5$ and characterized by sequences of modulation indices of different length N_h. It is possible to see that increasing the number of indices the PSD becomes smoother. Moreover, the sidelobes disappear (since there are no frequency notches) and are replaced by a small ripple. This spectral behavior is not surprising, since the PSD of a CPM signal with a long index sequence is—intuitively speaking—the average of the PSDs of all the single-h signals that use as index one of the ν possible indices.

B. Overall spectral efficiency

We consider an asynchronous SS-FH-CPM system using a 2-RC frequency pulse, $N_h = 16$ and $p = 8$. Since we are not interested in a particular sequence of indices but in the average behavior of the system, we consider a packet transmission (with 1024 symbols per packet) and, for each user, we change the sequence of indices $\{h_i^{(u)}\}_{i=0}^{N_h-1}$, the time delay $\tau^{(u)}$ and the initial phase offset $\theta^{(u)}$ every packet. We generate the indices in a quasi-random way. For the first user we generate the index sequence randomly, while the sequences of the remaining users are shifted versions (modulo h_{max}) of the sequence of the first user. The shifts are chosen in order to maximize the pairwise index distance defined as

$$d = \min_{u\neq v} \left| h_i^{(u)} - h_i^{(v)} \right|$$

between each couple of users. Obviously d remains the same for $i = 0, \ldots, N_h - 1$. Using the maximum distance, the correlations of all the possible couples of users are minimized and our system becomes more similar to an orthogonal system. Finally, to remove the correlation introduced by the shift, a random interleaver is used to scramble the simultaneous indices among the users. In the following computations of spectral efficiency, we adopt as bandwidth definition the bandwidth that contains the 99.9% of the overall power. Changing this definition would obviously cause a quantitative, but not qualitative, variation of the results.

In order to make some comparisons with the proposed SS-FH-CPM system, we first consider single-user systems using binary single-h CPMs with a 2-RC frequency pulse and $h < 1$, as traditionally done in literature. There is no interest in considering single-h systems with $h > 1$ because they have a larger bandwidth than those with $h < 1$ [19], resulting in a lower spectral efficiency. For the single-h systems the signal bandwidth strongly depends on h (as shown in Table I), and so does the spectral efficiency.

h	$1/8$	$3/8$	$1/2$	$5/8$	$7/8$
BT	0.94	1.28	1.62	1.87	2.12

Table I
Bandwidths of single-h 2-RC CPMs with different modulation indices.

Hence, we chose $h = 1/8$, $h = 3/8$, $h = 1/2$, $h = 5/8$, and $h = 7/8$, and compared the corresponding spectral efficiencies versus E_b/N_0, E_b being the received mean energy per information bit, with the overall spectral efficiency of the SS-FH-CPM binary system with $h_{\text{max}} = 39/8$ and $U = 37$ asynchronous users. The number of users U has been found maximizing η_U (via numerical simulations) as a function of U and the interference variance N_I assumed at the receiver for a fixed signal-to-noise ratio (SNR) value. As it can be seen in Fig. 3, the SS-FH-CPM system has a better spectral

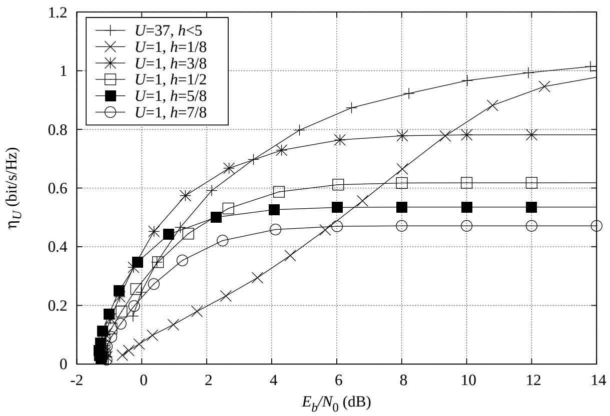

Figure 3. Spectral efficiencies of the considered 2-RC binary SS-FH-CPM with $N_h = 16$, $p = 8$, $h_{\max} = 39/8$, and of different single-h 2-RC CPMs with $h = 1/8$, $h = 3/8$, $h = 1/2$, $h = 5/8$ and $h = 7/8$, respectively. For the SS-FH-CPM signal, we used the (suboptimal) single-user detector.

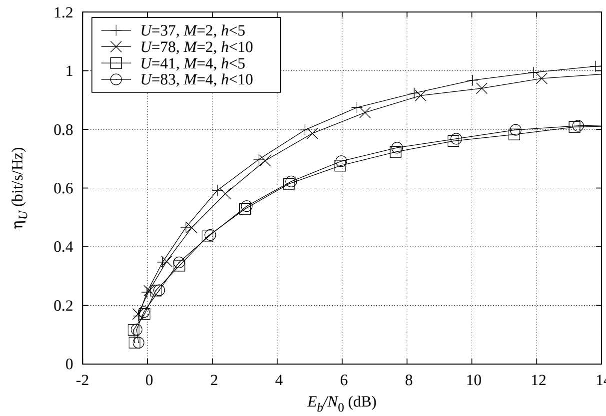

Figure 4. Spectral efficiencies of the considered 2-RC binary and quaternary SS-FH-CPM with $N_h = 16$, $p = 8$, $h_{\max} = 39/8$, and the same systems with double bandwidth ($h_{\max} = 79/8$). All curves have been obtained with a single-user detector.

efficiency than all single-user single-h systems for medium to high SNR values. At low SNR, η_U is in the same range of values as the single-h spectral efficiencies. According to the well-known results in information theory, the curve in Fig. 3 can be approached, even with $U \gg 1$, using a SUD and a proper channel code.

In traditional DS-SS systems, the number of users that maximizes the global spectral efficiency linearly depends on the total occupied bandwidth. Since in the proposed system the theoretical results obtained for linear modulations cannot be used, we will show via numerical simulations that this dependence is approximately linear also for the SS-FH-CPM system. In Fig. 4 we show the optimized η_U of the SS-FH-CPM system considered before, and the optimized η_U of a system with the same parameters but doubled bandwidth (i.e., a higher value of $h_{\max}$). For comparison, we show the same curves also for two quaternary systems. It is clear from Table II and Fig. 4 that doubling the bandwidth allows (approximately) doubling the number of users. Moreover, optimized binary systems outperform optimized quaternary systems.

M	2		4	
$h_{\max}$	$39/8$	$79/8$	$39/8$	$79/8$
BT	6.03	12.25	15.43	30.31

Table II
BANDWIDTHS OF 2-RC CPMS WITH $N_h = 16$ AND $p = 8$.

This last result is the reason why in the following we will discard higher order modulations and focus only on binary modulations. Therefore, a comparison among the SS-FH-CPM system and other binary systems, namely those proposed in [4], [8], and [9], named in the following GiLuRe, MuLa, and Mu, respectively, is needed. We fix the total bandwidth $BT \simeq 38$ for all the four systems and chose the spreading factors of GiLuRe, MuLa, and Mu systems, and the value of $h_{\max}$ for the proposed system accordingly. The resulting parameters are shown in Table III, where γ is the spreading factor and $T_c = T/\gamma$ is the chip period.

format	$h_{\max}, \gamma$	BT_c	BT
SS-FH-CPM	$311/8$	/	38.25
GiLuRe	24	1.62	38.83
MuLa	18	2.21	39.78
Mu	44	0.88	38.63

Table III
PARAMETERS USED TO COMPARE DIFFERENT SYSTEMS WITH THE SAME BANDWIDTH $BT \simeq 38$.

The number of asynchronous users has been optimized, jointly with the interference variance N_I, for all systems in order to maximize the global spectral efficiency. For the GiLuRe system we have chosen the 2-RC format (for a fair comparison with the proposed SS-FH-CPM system) and random chips as described in [4]. For the MuLa system we have chosen a roll-off factor $\alpha = 0$ since it is the value providing the best spectral efficiency [8]. Finally, for the Mu system we used the same parameters used in [9], i.e., $p = 1/3$, 10^4 primary iterations, 10^3 secondary iterations and random initial binary chips. The results reported in Fig. 5 show that our proposed system outperforms all other systems.

Finally, in order to show that it is possible to approach the performance promised by the information-theoretic analysis, we show the information rates for $U = 3$, 6, and 9 synchronous users (Fig. 6) and the corresponding BER curves (Fig. 7) obtained with rate-1/2 convolutional code with constraint length 5, generators $[2, 32]_8$ and codewords of length 64000 information bits, concatenated with the modulator through a random interleaver.[3] For both figures, the interference variance N_I has been optimized through numerical simulations. The interleavers (one for each user) used in the BER simulations have been generated randomly. At the receiver, iterative detection and decoding is performed for a maximum of 20 allowed iterations. As it can be observed, for $U = 3$, the loss with respect to the information rate curve is around 1 dB for $U = 3$,

[3]It is clear that the larger the number of users, the lower the information rate of each user (see Fig. 6). Hence, for a high number of users the information rate of each user is very low. For this reason, in order to employ codes with a rate sufficiently high, we consider a limited number of users (at most 9).

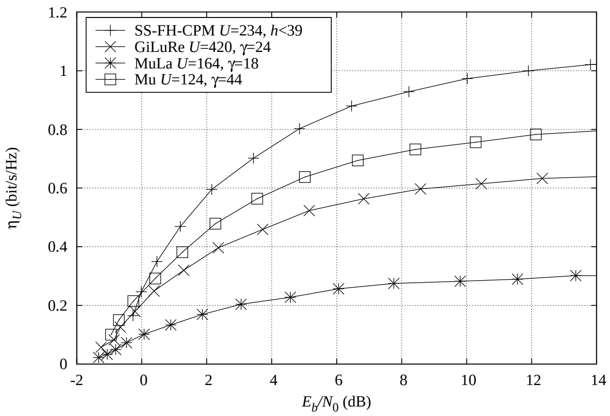

Figure 5. Spectral efficiencies of the proposed 2-RC binary SS-FH-CPM with $N_h = 16$ and $h_{\max} = 311/8$, GiLuRe 2-RC system with $\gamma = 24$, MuLa system with γ =18 and $\alpha = 0$, and Mu system with $\gamma = 44$. All curves have been obtained with a single-user detector.

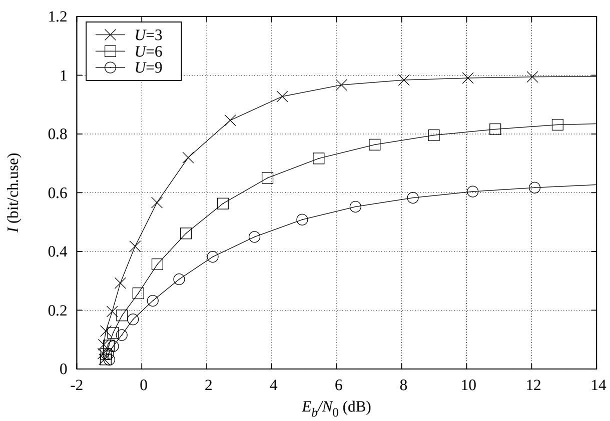

Figure 6. Information rates of the proposed 2-RC binary SS-FH-CPM with $N_h = 16$ and $h_{\max} = 39/8$. $U = 3$, $U = 6$, and $U = 9$ users have been considered. All curves have been obtained with a single-user detector.

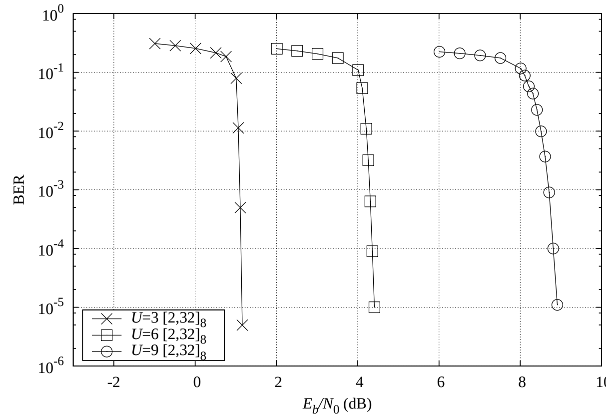

Figure 7. Bit error rate of the proposed 2-RC binary SS-FH-CPM with $N_h = 16$ and $h_{\max} = 39/8$. $U = 3$, $U = 6$, and $U = 9$ users have been considered.

2 dB for $U = 6$, and 3 dB for $U = 9$, despite the use of a very simple coding scheme [23]. An extensive search of the optimal convolutional codes for the three cases would further improve the BER performance (in particular for the system with $U = 9$).

C. BER with equal powers

In order to assess the performance of the described suboptimal MUDs, we considered a coded SS-FH-CPM system with $U = 3$ synchronous users using binary a 2-RC CPM with $p = 4$, $h_{\max} = 19/4$, and $N_h = 8$. All users have the same energy per symbol (i.e., $E_s^{(u)} = E_s$, $u = 1, 2, 3$) and employ the $(64, 51)$ extended Bose, Ray-Chaudhuri, Hocquenghem (eBCH) code with rate $R = 0.79$ and codewords of length 1024 bits described in [24], serially concatenated with the modulator through an S-random interleaver, with $S = 22$. As a benchmark, we consider the BER of a SUD with $U = 3$ users and the BER of a SUD in the absence of interference ($U = 1$ user). Again, we optimized the noise variance assumed by each detector and allowed 20 detector/decoding iterations.

For the suboptimal multi-user detector described in [17], the performance also depends on the adopted schedule. Serial or parallel schedules are usually considered in the literature. Since the difference in performance is practically negligible in this scenario of users transmitting at the same power, we only consider the parallel schedule. In this case, at each iteration all users are activated simultaneously. The computed soft-outputs are then provided to the other users for the next iteration and, after deinterleaving, to the decoders.

Since SIC 1 and SIC 2 detectors show the same performance when the users are uncorrelated (or weakly correlated) [32], we decided to introduce a correlation to point out the different behavior of the two algorithms. Therefore, we generated the index sequence for user $u = 1$ randomly, and from that we derived all the other sequences as

$$h_i^{(u)} = h_i^{(1)} + \frac{u-1}{p}\,.$$

If $h_i^{(u)}$ is an integer then we changed its value in $h_i^{(1)} + u/p$. In other words, the modulation indices of all users are close to each other as much as possible. The performance of the considered detectors is shown in Fig. 8. The HIC algorithm performs as the SUD because the interference prevents a correct bit estimation, which implies that (almost) no cancellation is done.

The SIC 2 algorithm performs much better than the HIC, but, as expected, even better does the SIC 1. However, the FG-based receiver has the best performance because the Gaussian approximation of the interference is not accurate with only two interferers. To see the SIC algorithms outperform the FG-based receiver, we should consider a much higher number of users.

D. BER with unbalanced powers

We also considered the case of unbalanced powers in a system with the same characteristics and parameters as the one described in the previous section. Without loss of generality, we chose to order the users in a decreasing way according to their energy, i.e., $E_s^{(1)} \geq E_s^{(2)} \geq \ldots \geq E_s^{(U)}$. We

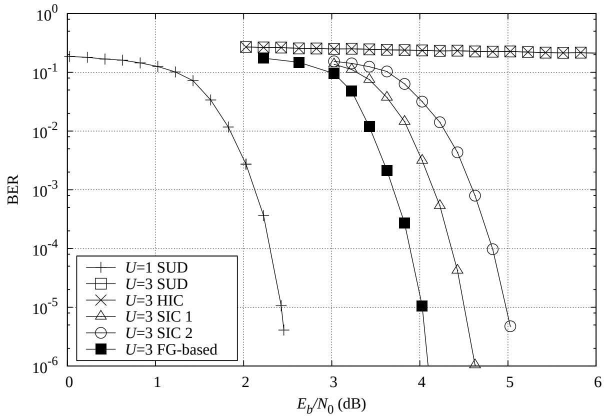

Figure 8. BER performance of the SUD and different MUDs in the case of a binary 2-RC system with $U = 1$ and $U = 3$, $N_h = 8$, $p = 4$, $h_{\max} = 19/4$, and a $(64, 51)$ eBCH code with rate $R = 0.79$.

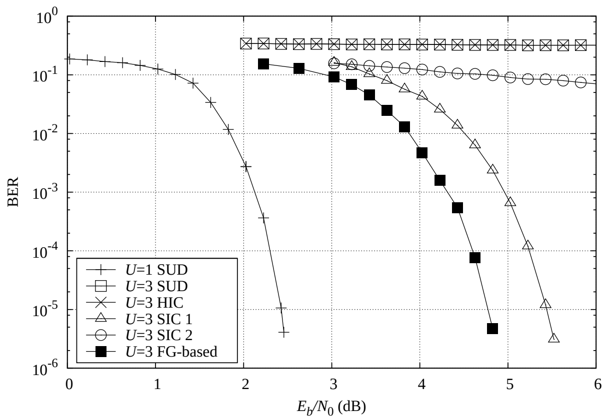

Figure 9. BER performance of the SUD and different MUDs in the case of an unbalanced binary 2-RC system with $U = 1$ and $U = 3$, $N_h = 8$, $p = 4$, $h_{\max} = 19/4$, and a $(64, 51)$ eBCH code with rate $R = 0.79$.

considered as reference user the central user ℓ and fixed its power $P^{(\ell)}$, while the powers of the other users are assumed to be $P^{(u)} = P^{(\ell)} + 2(\ell - u)$ dB. We employed S-random interleavers and we adopted a serial schedule, starting the detection from the user with the highest power. The computed soft-outputs are then provided to the users with lower powers for the interference cancellation and, after deinterleaving, to the decoders.

In Fig. 9 we show the performance of the different receivers. Again, the HIC algorithm performs as the SUD because the interference prevents a correct bit estimation. The SIC 2 has a poor performance, and again the FG-based receiver outperforms the SIC algorithms.

E. Optimization of the index sequences

In traditional linearly-modulated CDMA systems, the optimization of the spreading sequences (also called signature sequences) is a well-studied topic. Theoretical analyses have found the optimum sequences in synchronous systems, under either the condition $U \leq \gamma$ [33] or $U > \gamma$ [34], where γ is the spreading factor. In these cases, an iterative algorithm to determine the optimum sequence sets is available [35]. More recently a new approach to the optimization problem has been carried out by exploiting mathematical tools coming from the game theory [36].

Nevertheless, none of these techniques can be applied to CPM-based systems because of the nonlinearity of the modulation format. In linearly-modulated CDMA systems, the waveforms are independent of the information symbols and depend only on the signature sequence of each user. On the contrary, in CPM-based systems the waveforms depend in a nonlinear fashion not only on the index sequence, but also on all transmitted symbols because of the modulation memory. Therefore it is no longer possible to assume the orthogonality condition as an optimality criterion because symbols and waveforms are no more separable. Therefore, even though it might be possible to further investigate this issue, there is no evidence that a simple (or, at least, a practical) solution even exists.

V. Conclusions

We proposed a brand new technique allowing to use multi-h CPM in CDMA systems. Tuning the highest value the modulation index can assume, it is possible to set the spectral spreading of the CPM signal. PSD smoothness is reachable using a long enough sequence of modulation indices. Binary multi-h CPM-based system outperforms a similar quaternary system and other alternative solutions proposed in the literature.

In a multi-user scenario, the proposed system can surpass the spectral efficiency of a single-user single-h system, whereas the BER performance can be improved by a suboptimal multi-user detector.

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