Timing-free Blind Multiuser Detection for Multicarrier DS/CDMA Systems with Multiple Antennas

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Abstract

The problem of blind multiuser detection for an asynchronous multicarrier DS/CDMA system employing multiple transmit and receive antennas over a Rayleigh fading channel is considered in this paper. The solution that we develop requires prior knowledge of the spreading code of the user to be decoded only, while knowledge of the propagation delay and of the channel impulse response for the user of interest is not required at the receiver. The proposed receiver is immune to co-channel interferers transmitting with an arbitrarily large power and, as computer simulation results confirm, performs pretty close to the ideal linear minimum mean square error receiver, which requires knowledge of the spreading codes, symbol timings and channel impulse responses for all the active users.

1. Introduction

The explosive growth of the demand for high data-rates services through wireless links has posed the problem of devising modulation schemes and multiple access strategies with high spectral efficiency and intrinsically resistant to multipath distortions. Third-generation (3G) cellular systems are/will be mostly based on the Direct-Sequence Code Division Multiple Access (DS/CDMA) strategy, which is nowadays the leading technology for the realization of the air interface of many current cellular networks [1]. On the other hand, it is well-known that, due to the very short chip-duration, one of the main shortcomings of high-rate DS/CDMA systems is the need for fast electronic circuitry and channel equalizers, and the critical role played by the synchronization process at the receiver stage. In order to circumvent these drawbacks, it has been recently suggested to combine the potentiality of DS/CDMA systems with the benefits of multicarrier modulation formats: this leads to the concept of multicarrier CDMA, which was introduced in the early 90's (see [6] for a review of multicarrier CDMA techniques), and which is a serious candidate for the realization of the air interfaces of fourth-generation wireless cellular networks. Generally speaking, in a multicarrier system the available bandwidth is partitioned in many sub-bands, which are occupied by independently modulated digital signals. This partition has two positive effects: (a) the propagation channel in each sub-band is frequency-flat, i.e. there is no intersymbol interference at the receiver; and (b) the symbol duration for the data signals occupying the frequency sub-bands grows linearly with the number of sub-bands, thus implying that the need for fast electronics and high-performance synchronization schemes is less stringent. The combination of the multicarrier concept with the CDMA technology has led to the birth of three main access schemes, i.e. multitone CDMA [11, 12], multicarrier CDMA [14, 5, 15] and multicarrier DS/CDMA [7, 8, 9, 3]. This paper is concerned with the multicarrier DS/CDMA technique, even though the results herein presented can be easily extended to the other two modulation formats.

Besides the use of multicarrier modulation schemes, another interesting strategy to obtain high data-rates systems is to resort to the use of multiple transmit and receive antennas. Indeed, it has been recently shown that the capacity of a multi-antenna wireless communication system in a rich scattering environment grows with a law approximately linear in the minimum between the number of transmit and receive antennas [10]. Indeed, on one hand, the use of multiple transmit antennas permits increasing the system data-rate without increasing the system bandwidth; on the other hand, the use of multiple receive antennas leads to improved system performance due to the diversity gain. Motivated by these considerations, several papers that have recently appeared in the literature present theoretical findings and/or performance results for both single-user and multiuser multiantenna systems [4].

In this paper we consider an asynchronous multicarrier DS/CDMA system employing multiple antennas. We propose a new detection strategy, which does not require any knowledge beyond the spreading code for the user of interest. The proposed receiver has a two-stage architecture; the former stage suppresses the multiple access interference (MAI), while the latter stage is aimed at signal-to-noise ratio and BER optimization. Interestingly, the proposed receiver is immune to co-channel interferers with arbitrarily

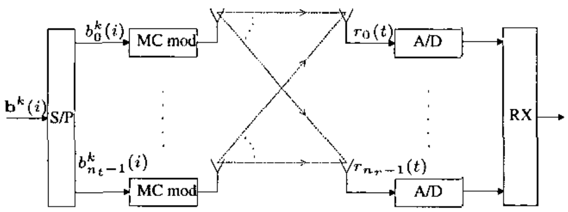

Figure 1. Scheme of a communication system with multiple transmit multiple receive.

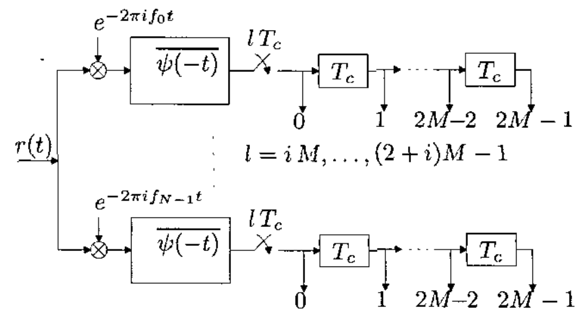

Figure 2. General A/D converter for multicarrier DS/CDMA.

large power and it performs pretty close to the ideal MMSE receiver, which requires knowledge of the spreading codes, channel states and propagation delays for all the users.

Notation: in the following, $\mathcal{M}_{m\times n}(\mathbb{C})$ is the set of all the $m \times n$-dimensional matrices with complex entries; $Im(\mathbf{A})$ is the image of $\mathbf{A}$, i.e. its column range span; $Ker(\mathbf{A})$ is the kernel of $\mathbf{A}$, i.e. the orthogonal complement of $Im(\mathbf{A})$; the superscripts $\overline{(\cdot)}$, $(\cdot)^T$ and $(\cdot)^H$ denote conjugate, transpose and conjugate transpose, respectively; the symbols $< \cdot, \cdot >$, $\otimes$ and $\odot$ denote the canonical scalar product, the Kronecker product and the Schur (i.e. component-wise) matrix product, respectively; $diag(\mathbf{a})$ denotes a diagonal matrix containing the entries of the vector $\mathbf{a}$ on its diagonal; finally $\mathbf{A}^+$ is the *Moore-Penrose generalized inverse* of $\mathbf{A}$.

2. System Model

The general diagram of a multicarrier communication system equipped with multiple transmit and receive antennas is shown in figure 1. A block of n_t symbols is converted from serial to parallel and each symbol feeds a (spatially) separate antenna. Thus, the n_t symbols are transmitted in parallel, achieving an n_t-fold increase in the data rate, and received on n_r spatially separated receive antennas, providing an n_r-th order receive diversity to combat fading (the coherence bandwidth of the channel is assumed to be smaller than the overall transmitted signals' bandwidth).

According to the above assumptions, the complex envelope of the signal received on the r-th antenna can be written as

$$r_r(t) = \sum_{l=0}^{P-1} \sum_{k=0}^{K-1} \sum_{t=0}^{n_t-1} b_t^k(l) s_{t,r}^k(t - l\,T_b - \tau^k) + n_r(t)\,, \quad (1)$$

where K is the number of active users, P is the length of the transmitted frame, $b_t^k(l)$ is the symbol transmitted by the t-th antenna of the k-th user at the l-th bit interval, T_b is the the bit duration, $\tau^k \in [0, T_b)$ is the k-th user's delay and $n_r(t)$ is the additive white Gaussian noise on the r-th receive antenna, independent for different antennas, with power spectral density equal to $2\mathcal{N}_0$. Finally, $s_{t,r}^k(\cdot)$ is the signature transmitted by the t-th antenna of the k-th user and received, after propagation, at the r-th antenna (namely it is a "spatial" signature related to the signal one through the channel impulse response). In a multicarrier DS/CDMA system with a sufficiently large number of carriers so that the channel coherence bandwidth exceeds the carrier spacing and assuming that cyclic prefix or guard interval are employed in order to avoid interference between successive transmitted symbol on each subcarrier, this signature is written as

$$s_{t,r}^k(t) = \sqrt{\frac{2\,\mathcal{E}_b^k}{NM}} \sum_{n=0}^{N-1} H_{n,t,r}^k \sum_{m=0}^{M-1} c_{t,n}^k(m)\, \psi(t - mT_c)\, e^{2\pi i f_n t}. \quad (2)$$

Here, $\mathcal{E}_b^k$ is the bit energy, N is the number of sub-carriers and M is the spreading gain along each sub-carrier. If $\mathbf{c}_t^k$ is an NM-dimensional vector representing the spreading sequence for the t-th antenna of the k-th user, then we denote by $\mathbf{c}_{t,n}^k$ the M-dimensional spreading sequence on each sub-carrier, obtained segmenting $\mathbf{c}_t^k$ into N subsequences. The quantity $T_c = T_b/M$ is the chip duration and $\psi(t)$ is a bandlimited chip waveform. Denoting by $\varphi(t) = r_\psi(t) = \int_{\mathbb{R}} \psi(x)\, \overline{\psi(x-t)}\, dx$ the cross-correlation function of $\psi(t)$, we assume that $\varphi(\cdot)$ is a bandlimited Nyquist waveform, i.e. $\varphi(nT_c) = \delta(n)$. In (2) $H_{n,t,r}^k$ is the complex Gaussian random variate representing the fading attenuation experienced by the n-th sub-carrier in its propagation from the t-th transmit antenna of the k-th user to the r-th receive one; we assume that the variates $H_{n,t,r}^k$ are independent for all n, t, r and k. Finally, f_n is the frequency of the n-th subcarrier with the subcarrier spacing exceeding the coherence bandwidth of channel in order to obtain independent fading (note, however, that to further ensure independence, it is only matter of incorporating sufficient interleaving).

At the receiver side, the signal observed on each antenna is converted to discrete-time; to illustrate further, let us focus on the k-th user signal and on a system with one transmit and receive antenna. According to the scheme in figure 2, there are N branches (i.e. as many as the number of carriers) in the A/D converter, each one consisting of a multiplier and of a chip-matched filter, whose output is sampled every T_c seconds. Since the timing of the desired user is not known, we consider for further processing $2MN$ samples of the received signal (i.e. $2M$ samples for each sub-carrier) so that one complete symbol is guaranteed to be observed. At the n-th branch, the output of the matched filter is written as

$$\int_{\mathbb{R}} r(x)\,\overline{\psi(x-t)}\,e^{-2\pi i f_n t}dx\ . \tag{3}$$

Now, if we let

$$i^k = \left\lfloor \frac{\tau^k}{T_c} \right\rfloor \quad \text{and} \quad \delta^k = \tau^k - i^k T_c \tag{4}$$

and define the following $M+3$ and $2M$-dimensional vectors:

$$\mathbf{g}^k_{sub} = \sqrt{\frac{2\,\mathcal{E}^k_b}{NM}} \begin{bmatrix} i^k \left\{ \begin{matrix} 0 \\ \vdots \\ 0 \end{matrix} \right. \\ \varphi(-T_c-\delta^k) \\ \varphi(-\delta^k) \\ \varphi(T_c-\delta^k) \\ \varphi(2T_c-\delta^k) \\ M-i^k-1 \left\{ \begin{matrix} 0 \\ \vdots \\ 0 \end{matrix} \right. \end{bmatrix}, \quad \mathbf{c}^{k,l}_n = \begin{bmatrix} l \left\{ \begin{matrix} 0 \\ \vdots \\ 0 \end{matrix} \right. \\ c^k_n(0) \\ \vdots \\ c^k_n(M-1) \\ M+2-l \left\{ \begin{matrix} 0 \\ \vdots \\ 0 \end{matrix} \right. \end{bmatrix} \tag{5}$$

along with the matrix

$$\mathbf{C}^k_n = [\ \mathbf{c}^{k,0}_n \ \cdots \ \mathbf{c}^{k,M+2}_n\ ] \in \mathcal{M}_{2M\times M+3}(\mathbb{C})\ , \tag{6}$$

the $2M$-dimensional vector containing the $2M$ samples (relative to the n-th sub-carrier) corresponding to the interval $[iT_b, (i+2)T_b)$ can be written as

$$\begin{aligned} \mathbf{r}_n(i) = \ & b^k(i-1)H^k_n \mathbf{C}^k_{n,-1}\mathbf{g}^k_{sub} + b^k(i)H^k_n\mathbf{C}^k_n\mathbf{g}^k_{sub} + \\ & + b^k(i+1)H^k_n\mathbf{C}^k_{n,+1}\mathbf{g}^k_{sub} + \mathbf{n}(i)\ . \end{aligned} \tag{7}$$

In deriving the above equation we have assumed that $|\varphi(x)| \simeq 0\ \forall |x| > 2T_c$; the matrices $\mathbf{C}^k_{n,-1}$ and $\mathbf{C}^k_{n,+1} \in \mathcal{M}_{2M\times M+3}(\mathbb{C})$, instead, are defined as

$$\mathbf{C}^k_{n,-1} = \begin{bmatrix} \mathbf{C}^k_{n\,L} \\ \mathbf{O} \end{bmatrix}, \quad \mathbf{C}^k_{n,+1} = \begin{bmatrix} \mathbf{O} \\ \mathbf{C}^k_{n\,H} \end{bmatrix}, \tag{8}$$

where $\mathbf{C}^k_{n\,H}$ and $\mathbf{C}^k_{n\,L} \in \mathcal{M}_{M\times M+3}(\mathbb{C})$ contain the M upper and M lower rows of the matrix $\mathbf{C}^k_n$, respectively.

The vectors on the N carriers can be stacked up so as to obtain the following vector:

$$\mathbf{r}(i) = \begin{pmatrix} \mathbf{r}_0(i) \\ \vdots \\ \mathbf{r}_{N-1}(i) \end{pmatrix} = b^k(i-1)\mathbf{C}^k_{-1}\mathbf{g}^k + b^k(i)\mathbf{C}^k\mathbf{g}^k + \\ + b^k(i+1)\mathbf{C}^k_{+1}\mathbf{g}^k + \mathbf{n}(i) \in \mathbb{C}^{2MN}. \tag{9}$$

In eq. (9) $\mathbf{g}^k = \mathbf{h}^k \otimes \mathbf{g}^k_{sub}$, with, in turn, $\mathbf{h}^k = (H^k_0 \cdots H^k_{N-1})^T \in \mathbb{C}^N$ denoting the vector of the Rayleigh fading coefficients; moreover, we have that

$$\mathbf{C}^k = \begin{pmatrix} \mathbf{C}^k_0 & & \mathbf{O} \\ & \ddots & \\ \mathbf{O} & & \mathbf{C}^k_{N-1} \end{pmatrix} \in \mathcal{M}_{2M\times M+3}(\mathbb{C})\ , \tag{10}$$

while the matrices $\mathbf{C}^k_{-1}$ and $\mathbf{C}^k_{+1}$ are defined similarly to eq. (10), and $\mathbf{n}(i) = (\,\mathbf{n}_0(i)\cdots\mathbf{n}_{N-1}(i))^T$ is a zero-mean Gaussian random vector with covariance matrix $2\mathcal{N}_0\mathbf{I}_{2MN}$.

In the general scenario with K active users each one equipped with multiple antennas, the discrete time signal received at the r-th antenna is represented by the following $2MN$-dimensional vector:

$$\begin{aligned} \mathbf{r}_r(i) = \sum_{k=0}^{K-1}\sum_{t=0}^{n_t-1} & \left[\mathbf{C}^k_{t,-1}\mathbf{g}^k_{t,r}b^k_t(i-1) + \mathbf{C}^k_t\mathbf{g}^k_{t,r}b^k_t(i) + \right. \\ & \left. + \mathbf{C}^k_{t,+1}\mathbf{g}^k_{t,r}b^k_t(i+1)\right] + \mathbf{n}_r(i) = b^0_0(i)\mathbf{C}^0_0\mathbf{g}^0_{0,r} + \mathbf{z}_r(i) + \mathbf{n}_r(i). \end{aligned} \tag{11}$$

Assuming that we are interested in decoding the information symbols transmitted by the 0-th antenna of the 0-th user, then in (11) $\mathbf{C}^0_0\mathbf{g}^0_{0,r}$ is the useful signature, $\mathbf{z}_r(i)$ represents the self-interference, MAI and ISI contribution and $\mathbf{n}_r(i)$ is the thermal noise. Note that the subscript "t" points out that each transmit antenna of a given user is assigned a different spreading sequence, a condition that will be shown to be necessary in blind systems. Finally the n_r observables can be stacked in the vector:

$$\begin{aligned} \mathbf{r}(i) = \begin{pmatrix} \mathbf{r}_0(i) \\ \vdots \\ \mathbf{r}_{n_r-1}(i) \end{pmatrix} = b^0_0(i) \begin{pmatrix} \mathbf{C}^0_0\mathbf{g}^0_{0,0} \\ \vdots \\ \mathbf{C}^0_0\mathbf{g}^0_{0,n_r-1} \end{pmatrix} + \begin{pmatrix} \mathbf{z}_0(i) \\ \vdots \\ \mathbf{z}_{n_r-1}(i) \end{pmatrix} + \\ + \begin{pmatrix} \mathbf{n}_0(i) \\ \vdots \\ \mathbf{n}_{n_r-1}(i) \end{pmatrix} = b^0_0(i)\mathbf{s}^0_0 + \mathbf{z}(i) + \mathbf{n}(i) \ \in \mathbb{C}^{2MNn_r}, \end{aligned} \tag{12}$$

where $\mathbf{s}^k_t = \mathbf{S}^k_t\mathbf{g}^k_t \in \mathbb{C}^{2MNn_r}$, $\mathbf{S}^k_t = \mathbf{I}_{n_r} \otimes \mathbf{C}^k_t \in \mathcal{M}_{2MNn_r\times(M+3)Nn_r}(\mathbb{C})$ and $\mathbf{g}^k_t = (\,\mathbf{g}^{k\,T}_{t,0}\cdots\mathbf{g}^{k\,T}_{t,n_r-1})^T \in \mathbb{C}^{(M+3)Nn_r}$.

3. Detector Design

The detector analyzed is linear and represented by the vector $\mathbf{m} = \mathbf{De}$, with $\|\mathbf{m}\| = 1$ (see figure 3). In what it follows, we also assume differential data encoding-decoding so as to cope with the absence of a phase reference. At the receiver side, the observables $\mathbf{r}_0(i), \ldots, \mathbf{r}_{n_r-1}(i)$ can be either processed separately and then combined or processed simultaneously; we'll refer to the former case as *non-cooperative* reception while to the latter case as *cooperative* reception. If we adopt a non-cooperative scheme, the signal at the output of the n_r antennas is processed through as many detectors. The output of the r-th detector is expressed as $\langle \mathbf{r}_r(i), \mathbf{m}_r\rangle$. A combining block then forms the n_r unquantized estimates of the differential phase $\vartheta_r(i) = \langle \mathbf{r}_r(i), \mathbf{m}_r\rangle\,\overline{\langle \mathbf{r}_r(i-1), \mathbf{m}_r\rangle}$ and combines them according to two different strategies:

1. *Hard Integration* (with a randomized offset):
$$\widehat{d}^0_0(i) = \mathrm{sgn}\Big[\sum_{r=1}^{n_t-1}\mathrm{sgn}\left[\Re\left(\vartheta_r(i)\right)\right] + u\Big], \quad u \sim U\big((-\tfrac{1}{2}, \tfrac{1}{2})\big);$$
2. *Soft Integration*: $\widehat{d}^0_0(i) = \mathrm{sgn}[\Re(\sum_{r=1}^{n_t-1}\vartheta_r(i))]$.

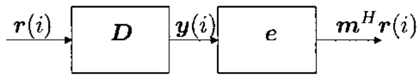

Figure 3. Linear Receiver scheme.

In the above rules, $\widehat{d}_0^0(i)$ is the detected information bit after that the differential encoding has been removed. Note that having imposed $\|\mathbf{m}_r\| = 1$, for $r = 0, \ldots, n_r - 1$, implies that all the estimates of the differential phase are uniformly weighted and therefore the soft integration is effectively an equal gain combiner technique. Conversely, with regard to the cooperative receiving scheme, the observables are first stacked in a unique vector as in (12) and then processed with a single receiver obtaining $\widehat{d}^0(i) = \text{sgn}\left[\Re\left(\langle \mathbf{r}(i), \mathbf{m}\rangle \overline{\langle \mathbf{r}(i-1), \mathbf{m}\rangle}\right)\right]$. Obviously, the cooperative scheme brings, at the price of some complexity increase, a substantial performance improvement with respect to the non-cooperative detection schemes.

Notice that eq. (12) reduces to (11) when we have one receive antenna. As a consequence, the synthesis of the receiver can be carried out once for both the situations of cooperative and non-cooperative detection considering the observables in (12):

$$\mathbf{r}(i) = b_0^0(i)\mathbf{S}_0^0\mathbf{g}_0^0 + \mathbf{z}(i) + \mathbf{n}(i) = \mathbf{q}(i) + \mathbf{n}(i) \quad \in \mathbb{C}^{2NMn_r}, \tag{13}$$

with correlation matrix:

$$\mathbf{R}_r = E\left[\mathbf{r}(i)\mathbf{r}^H(i)\right] = \mathbf{R}_q + 2\mathcal{N}_0\mathbf{I}_{2MN}. \tag{14}$$

Let us now move on to the synthesis of the two-stage receiver.

3.1. Synthesis of the first stage D

Looking at eq. (13), it is seen that the useful signature is a linear combination of the columns of $\mathbf{S}_0^0$ according to the entries of the vector $\mathbf{g}_0^0$, thus implying that it lies in $Im(\mathbf{S}_0^0) \in \mathbb{C}^{(M+3)Nn_r}$. The first stage is thus a non-invertible transformation of the observables, i.e.

$$\mathbf{y}(i) = \mathbf{D}^H\mathbf{r}(i), \tag{15}$$

where $\mathbf{D} \in \mathcal{M}_{2MNn_r \times (M+3)Nn_r}(\mathbb{C})$ matrix solving one the following constrained minimization problems

$$\begin{cases} E\left[\|\mathbf{D}^H\mathbf{r}(i)\|^2\right] = \min \\ \det\left(\mathbf{D}^H\mathbf{S}_0^0\right) \neq 0 \end{cases}, \quad \begin{cases} E\left[\|\mathbf{D}^H\mathbf{q}(i)\|^2\right] = \min \\ \det\left(\mathbf{D}^H\mathbf{S}_0^0\right) \neq 0 \end{cases}. \tag{16}$$

The first cost function is the classical one for minimum mean output energy (MMOE) while the second involves the minimization of the noise-free observables. Note that the constraint ensures that the signal of interest always survives after the non-invertible transformation.

Following the same steps as in [2], wherein a single-antenna DS/CDMA system is considered, it can be shown that under the assumption

$$\dim\left(Im(\mathbf{R}_q) \cap Im(\mathbf{S}_0^0\mathbf{S}_0^{0\,H})\right) = 1, \tag{17}$$

the solution to the above problems is written as

$$\mathbf{D} = \left(\mathbf{R} + \mathbf{S}_0^0\mathbf{S}_0^{0H}\right)^{-1}\mathbf{S}_0^0\,\mathbf{B}, \tag{18}$$

where $\mathbf{B} = \left[\left(\mathbf{C}^0\left(\mathbf{R} + \mathbf{S}_0^0\mathbf{S}_0^{0H}\right)^+\mathbf{S}_0^{0H}\right) \odot \mathbf{I}\right]^{-1} \text{diag}(\beta)$ and $\beta \in \mathcal{C}^{(M+3)Nn_r}$ is an arbitrary vector with strictly positive entries and $\mathbf{R}$ can be either $\mathbf{R}_r$ or $\mathbf{R}_q$. If $\mathbf{R} = \mathbf{R}_r$, $\mathbf{D}$ is the solution to the former problem in (16) and subsumes, as a special case of non-fading channel with known timing, the MMOE solution, equivalent to the MMSE one; thus we refer to this solution as an MMSE-like receiver. Otherwise, if $\mathbf{R} = \mathbf{R}_q$, $\mathbf{D}$ is the solution to the latter problem in (16) and subsumes in the same way the ZF solution; we thus refer to this solution as ZF-like receiver. Since scalar multiplicative constants can be shown to have no influence on the decision rule, the matrix $\mathbf{D}$ can be also expressed as:

$$\mathbf{D} = \left(\mathbf{R} + \mathbf{S}_0^0\mathbf{S}_0^{0\,H}\right)^+\mathbf{S}_0^0. \tag{19}$$

3.2. Synthesis of the second stage e

Assuming that the blocking matrix $\mathbf{D}$ has suppressed the most part of the MAI and ISI contribution (the term $\mathbf{D}^H\mathbf{z}(i)$ is very small in the MMSE-like receiver while is exactly zero in the ZF-like one) the observables at the output of the second stage can be written as

$$\mathbf{y}(i) = b_0^0(i)\,\mathbf{D}^H\mathbf{S}_0^0\mathbf{g}_0^0 + \mathbf{D}^H\mathbf{n}(i). \tag{20}$$

The vector $\mathbf{e}$ is then chosen so as to minimize the BER, i.e. it is the cascade of a whitening filter and of a filter matched to the whitened useful signal. Upon considering the following "economy size" singular value decomposition $\mathbf{D} = \mathbf{U}_D\mathbf{\Lambda}\mathbf{V}^H$, the whitening filter is $\mathbf{V}\,\mathbf{\Lambda}^{-1}$, with $\mathbf{\Lambda} \in \mathcal{M}_{(M+3)Nn_r \times (M+3)Nn_r}(\mathbb{C})$ a diagonal eigenvalues matrix and $\mathbf{V} \in \mathcal{M}_{(M+3)Nn_r \times (M+3)Nn_r}(\mathbb{C})$ a unitary square matrix. Accordingly, the whitened observables are given by

$$\mathbf{y}_w(i) = \left(\mathbf{V}\,\mathbf{\Lambda}^{-1}\right)^H\mathbf{D}^H\mathbf{r}(i) = b^0(i)\mathbf{U}_D^H\mathbf{C}^0\mathbf{g}_0^0 + \mathbf{U}_D^H\mathbf{D}^H\mathbf{n}(i) \tag{21}$$

and the matched filter is $\mathbf{U}_D^H\mathbf{C}^0\mathbf{g}^0$. Since $\mathbf{g}_0^0$ is not known, a further processing is needed to obtain an estimate of the matched filter. In particular, notice that the correlation matrix of $\mathbf{y}_w(i)$ can be written as

$$\mathbf{R}_{y_w} = \mathbf{U}_D^H\mathbf{S}_0^0\mathbf{g}_0^0\left(\mathbf{U}_D^H\mathbf{S}_0^0\mathbf{g}_0^0\right)^H + 2\mathcal{N}_0\,\mathbf{I}_{(M+3)Nn_r}, \tag{22}$$

Consequently, the eigenvector $\mathbf{u}_{max}$ corresponding to the largest eigenvalue of $\mathbf{R}_{y_w}$ is parallel to $\mathbf{U}_D^H\mathbf{S}_0^0\mathbf{g}_0^0$, and the receiver second stage is $\mathbf{e} = \mathbf{V}\,\mathbf{\Lambda}^{-1}\mathbf{u}_{max}$. Thus the complete receiver is given by:

$$\mathbf{m} = \mathbf{D}\,\mathbf{e} = \mathbf{U}_D\,\mathbf{\Lambda}\,\mathbf{V}^H\mathbf{V}\,\mathbf{\Lambda}^{-1}\mathbf{u}_{max} = \mathbf{U}_D\mathbf{u}_{max}. \tag{23}$$

In practice, the vector $\mathbf{u}_{max}$ is estimated through an eigendecomposition of the sample covariance matrix

$\widehat{\mathbf{R}}_{\mathbf{y}_w}$ of the whitened observables $\mathbf{y}_w(i)$ with $\widehat{\mathbf{R}}_{\mathbf{y}_w} = \frac{1}{Q}\sum_{i=0}^{Q-1} \mathbf{y}_w(i)\,\mathbf{y}_w(i)^H = \mathbf{U}_D^H \widehat{\mathbf{R}}_r \mathbf{U}_D$.

As a by product of the previous derivations, an estimate (up to a scalar factor) of the discrete-time channel impulse response $\mathbf{g}_0^0$ can be obtained, based on the consideration that $\mathbf{u}_{max}$ is parallel to $\mathbf{U}_D^H \mathbf{S}_0^0 \mathbf{g}_0^0$. Accordingly, the estimate $\widehat{\mathbf{g}}_0^0$ of $\mathbf{g}_0^0$ is:

$$\widehat{\mathbf{g}}_0^0 = (\widehat{\mathbf{U}}_D^H \mathbf{S}_0^0)^{-1} \widehat{\mathbf{u}}_{max} \, . \qquad (24)$$

3.3. Maximum Number of Users

Condition (17) states that $\mathbf{s}_0^0$ can be univocally determined as the common direction of the two subspaces $Im(\mathbf{R}_q)$ and $Im(\mathbf{S}_0^0)$ and then is called *identifiability condition*, common to all the detectors using a subspace approach. This condition puts a limit on the maximum rank of $\mathbf{R}_q$, and, consequently, on the maximum number of users K_{max} that the system can reliably accommodate. K_{max} can be easily found by observing that:

$$2MNn_r \geq \dim(Im(\mathbf{R}_q) \cup Im(\mathbf{S}_0^0 \mathbf{S}_0^{0H}) = 3Kn_t + (M+3)Nn_r - 1 \qquad (25)$$

and then $K_{max} = \left\lfloor \frac{(M-3)Nn_r + 1}{3n_t} \right\rfloor$.

It is finally worthwhile noticing that:

- K_{max} can be raised enlarging the processing window; in particular taking $(2m+2)MNn_r$ samples $K_{max} = \left\lfloor \frac{(2m+2)MNn_r - (M+3)Nn_r + 1}{(2m+3)n_t} \right\rfloor$.
- The cooperative detection scheme can accommodate a larger number of users than the non-cooperative scheme at the price of some complexity increasing. In fact, due to the inversion in the first stage and to the singular value decomposition in the second one, the receiver complexity is cubic with the dimension of $\widehat{\mathbf{R}}_r$, i.e. it is $\mathcal{O}\left((MNn_r)^3\right)$. So, the cooperative scheme has a complexity proportional to n_r^3, while the non-cooperative one is proportional to n_r. Note, however, that, coupling the RLS procedure with subspace tracking techniques as in [2], the overall complexity can be limited to be quadratic.
- If each user were given one signature, then the identifiability condition would not hold any longer, and the implementation of the second stage of the receiver would end up problematic. In this situation, training sequences are needed to separate the signals transmitted by the n_t antennas of each given user.

4. Numerical Results

The performance of the proposed detector is investigated through Montecarlo simulation results representing the system BER averaged over 10^4 random channels and delays realization. In the simulation each user is equipped with two transit antennas, the cross-correlation $r_\psi(t)$ of the chip waveform is a raised cosine with roll-off factor 0.22, the number of sub-carriers N is 4 and the spreading over each one M is 8 (the composite spreading gain is then 32 and the spreading sequences are PN of length 31 stretched out with a -1); finally the sample correlation matrix $\widehat{\mathbf{R}}_r$ is obtained through an estimate over $Q = 1300$ samples). The probability of error and channel estimation quality of the receiver are shown versus the ratio $\frac{\mathcal{E}_b}{\mathcal{N}_0}$ and are compared with the ones of the MMSE-like limit receiver (i.e. when the sampled correlation matrix tend to the real one), subspace-based MMSE and ZF receivers (i.e. the one proposed in [13, 9] where the orthogonality between the noise subspace and the useful signal $\mathbf{s}_0^0$ is exploited to extract the timing and the channel required to realize the canonical decorrelating and MMSE receivers: $\mathbf{m}_{MMSE} = \widehat{\mathbf{R}}_r^{-1}\widehat{\mathbf{s}}^0$, $\mathbf{m}_{ZF} = \widehat{\mathbf{R}}_r^{+}\widehat{\mathbf{s}}^0$) and ideal MMSE receiver (i.e. $\mathbf{m} = \mathbf{R}_r^{-1}\mathbf{s}^0$).

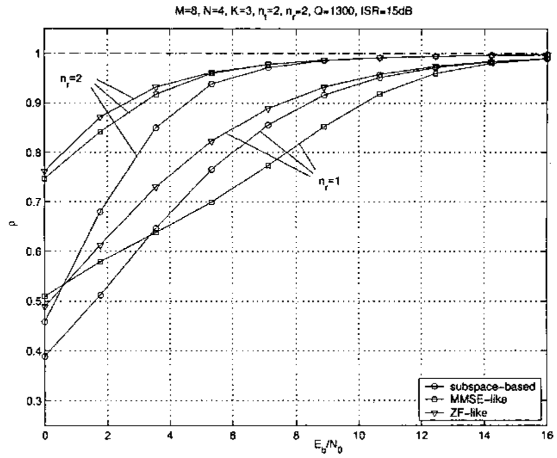

Figure 4. Channel estimation for both cooperative and non-cooperative reception.

The Monte Carlo simulation are presented in a severe near-far scenario (+15dB) and with 3 active users, for both cooperative and non-cooperative reception (note that the maximum number of user for the cooperative scheme is 6 while for the non cooperative one is 3, so, in the latter case, the network is fully loaded; note also that enlarging the processing window with $m = 2$ the maximum number of user in the cooperative case grows to 21). Figure 4 show the correlation coefficient $\rho = \langle \widehat{\mathbf{g}}_0^0, \mathbf{g}_0^0 \rangle / (\|\widehat{\mathbf{g}}_0^0\| \, \|\mathbf{g}_0^0\|)$ versus $\frac{\mathcal{E}_b}{\mathcal{N}_0}$ in order to evaluate the channel estimation quality, while figures 5 and 6 present the system BER. In the non-cooperative scheme, with the network fully loaded, best channel estimation is achieved by the ZF-like receiver, immediately followed by the subspace-based one, while for the cooperative case both the ZF-like and MMSE-like receiver outperform the subspace-based one. This trend is maintained in the error probability: ZF-like receiver performs slightly better then the ZF subspace-based one in both the cases while the MMSE-like receiver do that only in the cooperative case.

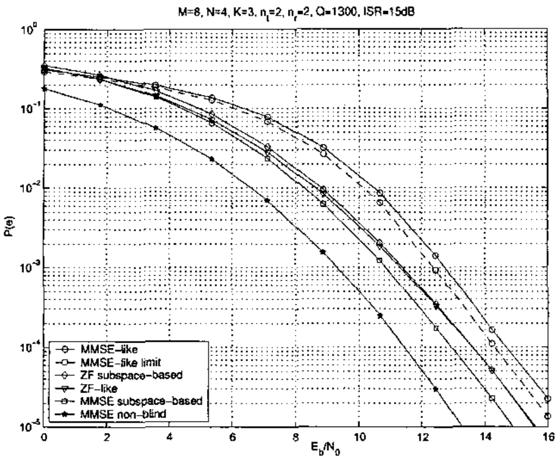

Figure 5. Error probability versus $\frac{\mathcal{E}_b}{\mathcal{N}_0}$ for non cooperative soft integration reception.

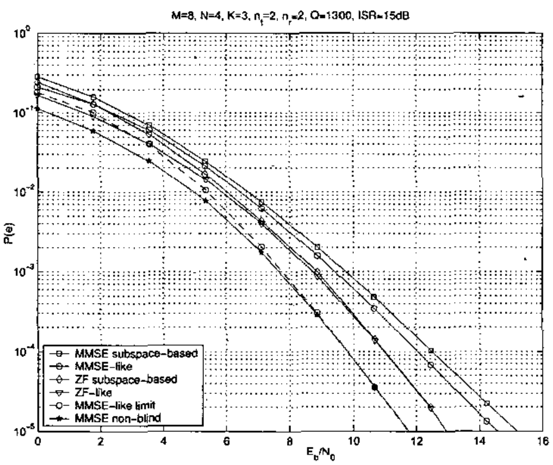

Figure 6. Error probability versus $\frac{\mathcal{E}_b}{\mathcal{N}_0}$ for cooperative reception.

Finally note that the cooperative reception outperforms the non-cooperative one and, for large Q, achieves a performance practically coincident with that of the ideal MMSE receiver.

5. Conclusions

In this paper we have considered the problem of blind multiuser detection for asynchronous multicarrier DS/CDMA systems equipped with multiple transmit and receive antennas. This is nowadays an interesting research topic, since multicarrier modulation formats coupled with the use of multiple antennas represent a suitable means to achieve high data-rates on the wireless channel at a reasonable computational and practical implementation cost. The receiver that has been proposed here is code-aided, in the sense that it requires knowledge of the spreading code for the user of interest only, while no prior knowledge on the channel state and on the timing offset is needed. Simulation results have shown that the proposed detection strategy performs pretty close to the ideal MMSE receiver, and that the use of multiple receiving antennas has a beneficial impact on the system performance. Future work on this topic comprises the consideration of space-time and space-frequency codes, as well as the extension of the proposed detection strategy to the situation in which the channel is time-dispersive, i.e. it does not remain constant over the whole transmitted frame.

References

[1] F. Adachi, M. Sawahashi, and H. Suda. Wideband ds-cdma for next-generation mobile communication systems. *IEEE Personal Commun. Magazine*, 36:56–69, Sep. 1998.

[2] S. Buzzi, M. Lops, and H. V. Poor. Timing free code aided blind adaptive joint mai and isi suppression in dispersive cdma channels. *IEEE Wireless Comm. and Networking Conf. (Orlando)*, pages 542–548, Mar. 2002.

[3] L. Fang and L. Milstein. Performance of successive interference cancellation in convolutionally coded multicarrier ds/cdma systems. *IEEE Trans. Commun.*, 49:206–2067, Dec. 2001.

[4] G. J. Foschini and M. J. Gans. On limits of wireless communication in a fading environment when using multiple antennas. *Wireless Personal Comm.*, 6:311335, Mar. 1998.

[5] X. Gui and T. S. Ng. Performance of asynchronous orthogonal multicarrier cdma system in frequency selective fading channel. *IEEE Trans. Commun.*, 47:1084–1091, Jul. 1999.

[6] S. Hara and R. Prasad. Overview of multicarrier cdma. *IEEE Commun. Magazine*, 35:126–133, Dec. 1997.

[7] S. Kondo and L. Milstein. Performance of multicarrier ds cdma systems. *IEEE Trans. Commun.*, 44:238–246, Feb. 1996.

[8] S. Miller and B. J. Rainbolt. Mmse detection of multicarrier cdma. *IEEE J. Select. Areas Commun.*, 18:2356–2362, Nov. 2000.

[9] J. Namgoong, T. F. Wong, and J. S. Lehnert. Subspace multiuser detection for multicarrier ds-cdma. *IEEE Trans. Commun.*, 48:1897–1908, Nov. 2000.

[10] E. Telatar. Capacity of multiantenna gaussian channels. *AT&T-Bell Labs. Internal Technical Memo.*, Jun. 1995.

[11] L. Vandendorpe. Multitone direct sequence cdma system in an indoor wireless environment. *Proc. of the IEEE First Symp. of Commun. and Vehic. Technology*, pages 3–14, Oct. 1993.

[12] L. Vandendorpe. Multitone spread spectrum multiple access communications system in a multipath rician fading channel. *IEEE Trans. Vehic. Technology*, 44:327–337, May 1995.

[13] X. Wang and H. V. Poor. Blind multiuser detection: a subspace approaach. *IEEE Trans. on Inform. Theory*, 44:311335, Mar. 1998.

[14] N. Yee, J. P. Linnartz, and G. Fettweis. Multicarrier cdma in indoor wireless radio network. *Proc. 'IEEE Int. Symp. on Personal, Indoor and Mobile Radio Commun. (Yokohama)*, Sep. 1993.

[15] P. Zong, K. Wang, and Y. Bar-Ness. Partial sampling mmse interference suppression in asynchronous multicarrier cdma system. *IEEE J. Select. Areas Commun.*, 19:1605–1613, Aug. 2001.