BLIND DIFFERENTIAL SCHEMES FOR CDMA ON DISPERSIVE MIMO CHANNELS

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ABSTRACT

This paper considers the problem of fully blind detection in an asynchronous dispersive Multiple-Input Multiple-Output (MIMO) Code Division Multiple Access (CDMA) channel, wherein each user is assigned one and the same spreading code to be used on all of the transmit antennas. In this scenario, differential space-time coding is needed, and conditions for blind linear user separability and channel equalization are stated. Focusing on the differential Alamouti code, two new decoding structures suitable for frequency-selective channels are then presented and discussed. Interestingly, they are amenable to a fully-blind implementation, i.e. they require only prior knowledge of the spreading code of the user of interest, while retaining a complexity only linear in the cardinality of the transmitted constellation.

1. INTRODUCTION

We consider the problem of an asynchronous slow-fading wide-band CDMA network wherein each user employs a set of t transmit antennas and the base station is equipped with r receive antennas: in this situation, Multiple-Access Interference (MAI) and Inter-Symbol Interference (ISI) elimination and bit-error-rate optimization would in principle require either receiver training, so as to extract the composite channel responses, or the adoption of blind techniques. As shown in [1], reliable channel estimates may dramatically impair the system transmission efficiency, especially in the low signal–to–noise ratio region. On the other hand, using blind techniques over dispersive channels may pose additional problems: indeed, the techniques presented in [2–4] all assume that each active user is assigned t spreading codes, one for each transmit antenna, to enable both blind linear MAI and ISI elimination and easy re-association, at the receiver end, of the demodulated data with the corresponding transmit antenna. Such an inefficient use of the spreading codes is inevitable if the multi-antenna transmitter is a pure spatial multiplexer, but represents a resource waste if space-time coding is adopted: indeed, the space-time code represents itself an additional signature that can be exploited at the receiver end for interference suppression and/or data re-association [5].

In this paper, we consider an asynchronous CDMA system – encompassing both direct sequence CDMA (DS/CDMA) and Multi-Carrier DS/CDMA (MC-DS/CDMA) – over a dispersive slow-fading MIMO channel, wherein each user transmits differentially Space-Time Block (STB) encoded data [6–8], using the same spreading code on all of the transmit antennas. At first, general conditions for blind linear user separability and ISI removal are stated. After interference elimination, focusing on the differential version of the Alamouti STB Code (STBC) [6, 7], two new differential space-time detection structures suitable for frequency-selective channels are introduced and assessed, which exhibit some interesting features. On one hand, those strategies subsume as special case the incoherent differential receiver of [6, 7]; on the other hand, they are amenable to a fully blind implementation, i.e. no prior knowledge of the timing, of the MIMO channel responses and of the encoder delay is required for both the user of interest and the interfering users, nor, obviously, any information is needed on the spreading codes of the interfering users. Merits and drawbacks of this system are demonstrated through a thorough performance assessment.

The paper is organized as follows. In next Section, a brief outline of the system model is presented, and the proposed receive strategies are derived. Numerical results are presented in Section 3, while concluding remarks are given in Section 4.

2. RECEIVER DERIVATION

Assume that there are K active users equipped with t transmit antennas and sharing the same bandwidth $2W$, in general split up into M disjoint subcarriers; each subcarrier has a bandwidth B_{sc} and a guard band B_g is inserted between adjacent subcarriers. The bit stream of each user is differentially STB encoded by transmitting $2b$ bits every t symbol intervals through a unitary $t \times t$ code matrix as in [7,8]. At epoch $p = 0, \dots, P-1$, the transmitted code matrix for user $k = 0, \dots, K-1$ is

$$\mathbf{\Sigma}_p^k = \begin{pmatrix} s_0^k(tp) & \dots & s_0^k((p+1)t-1) \\ \vdots & & \vdots \\ s_{t-1}^k(tp) & \dots & s_{t-1}^k((p+1)t-1) \end{pmatrix} = \mathbf{\Sigma}_{p-1}^k \mathbf{M}_p^k \in \mathcal{L},$$

$\mathcal{L}$ representing the set of all of the possible transmitted codewords, and $\mathbf{M}_p^k$ being a $t \times t$ unitary matrix carrying the new symbols to be transmitted at epoch p. The transmission rate is $\mathcal{R} = 2b/(tT_s)$, T_s being the duration of the symbol interval. In the following, we mostly focus on the Alamouti STBC [6], where $t = 2$ and

$$\mathbf{M}_p^k = \begin{pmatrix} \mu^k(2p) & -\mu^k(2p+1)^* \\ \mu^k(2p+1) & \mu^k(2p)^* \end{pmatrix},$$

with $\{\mu^k(q), \quad q = 0, \dots, 2P-1\}$ representing the uncoded process, belonging to any constant modulus, half-energy constellation $\mathcal{A}$ with cardinality 2^b (for example, a 2^b-PSK).

We assume that "0" is the user of interest and its unknown transmission delay τ^0 is regarded as the sum of a system delay $\tau_s^0 \in [0, T_s)$ tied to the channel, and an encoder delay $\tau_c^0 = n_{\tau_c^0} T_s$, with $n_{\tau_c^0} \in \{0, \dots, t-1\}$. We refer to [2–4] for the mathematical details concerning signal discretization. What matters here is that the discrete-time signal received at epoch $q = pt + u$, $u = 0, \dots, t-1$ and $p = 0, \dots, P-1$, can be written as

$$\begin{aligned} \mathbf{r}(q) &= \underbrace{\sum_{i=0}^{t-1} s_i^0(q - n_{\tau_c^0}) \mathbf{h}_i^0}_{\text{useful signal}} + \underbrace{\mathbf{z}(q)}_{\text{ISI+MAI}} + \underbrace{\mathbf{w}(q)}_{\text{noise}} \\ &= \mathbf{d}(q) + \mathbf{z}(q) + \mathbf{w}(q) \in \mathbb{C}^{\overline{m}NQr}. \end{aligned} \quad (1)$$

In (1), Q is the oversampling factor, whereas $N = N_{sc}M$ is the overall processing gain, N_{sc} being the spreading factor along each subcarrier (notice that for $M = 1$ the system reduces to a plain DS-CDMA); $\overline{m} \geq \overline{L} = \left(2 + \left\lceil \frac{L-Q-1}{N_{sc}Q} \right\rceil\right)$ is the length (expressed in symbol intervals) of the processing window, with L the sum of the maximum multi-path delay spread T_m of the channel and the time duration of the convolution of the transmit and receive filters (expressed in multiples of the sampling rate $\frac{T_s}{N_{sc}Q}$). Finally, the

$\overline{m}NQr$-dimensional vectors $\{\mathbf{h}_i^0,\ i=0,\ldots,t-1\}$ represent the composite signatures received at the base station, whereas $\mathbf{w}(q)$ is a Gaussian noise vector. Notice that, even though the additive thermal noise is assumed to be white with power spectral density $2\mathcal{N}_0$, using band-limited receive filters may introduce a known correlation among the noise samples.

As outlined in [4], the unknown composite signatures in (1) can be equivalently expressed as

$$\mathbf{h}_i^0 = \overline{\mathbf{S}}^0 \overline{\mathbf{g}}_i^0, \quad \text{for } i=0,\ldots,t-1, \tag{2}$$

where $\overline{\mathbf{g}}_i^0$ is a channel vector with a cluster of LMr consecutive non-zero entries whose position is dictated by the user delay τ_s^0, and whose length is determined by T_m and by the temporal extension of the transmit and receive filters. As to the matrix $\overline{\mathbf{S}}^0$, it is uniquely determined by the spreading code assigned to user "0". More generally, the composite spatial signatures appearing in (2) are formed as weighted sums of LMr consecutive columns of $\overline{\mathbf{S}}^0$, which can be cast in the following reduced-dimension matrix

$$\widetilde{\mathbf{S}} = \overline{\mathbf{S}}^0\Big(:, n_{\tau_s^0}Mr{+}1 : (n_{\tau_s^0}{+}L)Mr\Big) \in \mathcal{M}_{\overline{m}NQr\times LMr}, \tag{3}$$

where $n_{\tau_s^0} = \lfloor \tau_s^0 Q/T_c \rfloor$ and $T_c = T_s/N_{sc}$. A fully blind system may solely rely upon knowledge of $\overline{\mathbf{S}}^0$, while all of the other quantities in (1) are unknown at the receiver, and so is the position of the first non-zero column contributing to the matrix $\widetilde{\mathbf{S}}$ in (3).

We consider a two-stage receive structure. The first block is a linear system which should suppress the overall interference, while the second stage decodes the transmitted information.

We start by noticing that, due to the structure imposed by the STBC, the received signal $\mathbf{r}(q)$ in (1) is, in general, cyclostationary with period t and correlation matrices $\mathbf{R}_{rr}(u) \triangleq \mathrm{E}[\mathbf{r}(pt{+}u)\mathbf{r}(pt{+}u)^H]$, $u=0,\ldots,t-1$, which we assume either known or estimated. From [3,4], accounting for the cyclostationarity of $\mathbf{r}(q)$, the blocking stage may consist of a set of $\overline{m}NQr \times LMr$-dimensional matrices, $\{\mathbf{U}(u),\ u=0,\ldots,t-1\}$ say, obtained by cascading a noise-whitening filter $\mathbf{W}(u)$ to an interference–blocking-matrix $\mathbf{D}(u)$, which solves the following generalized minimum mean output energy (MMOE) problem:

$$\min_{\mathbf{D}(u)} \mathrm{E}\left[\left\|\mathbf{D}^H(u)\mathbf{v}(pt+u)\right\|^2\right], \text{ with } \det\Big(\mathbf{D}^H(u)\widetilde{\mathbf{S}}\Big) \neq 0, \tag{4}$$

for $u=0,\ldots,t-1$. In (4), $\mathbf{v}(q)$ may be chosen equal to $\mathbf{r}(q)$ (MMSE-like solution) or to $\mathbf{d}(q)+\mathbf{z}(q)$ (Zero-Forcing-like solution). Evidently, if the structure of the STBC is such that the observations are wide-sense stationary (this is, for example, the case for the differential Alamouti STBC), $\mathbf{D}$ becomes time-invariant, i.e. $\mathbf{D}(0)=\ldots=\mathbf{D}(t-1)=\mathbf{D}$. Requiring that the system performance be noise-limited amounts to requiring that, choosing $\mathbf{v}(q)=\mathbf{d}(q)+\mathbf{z}(q)$, the mean output energy be completely nullified [2–4]. To this end, we give the following:

Proposition 1. *Let $\mathbf{R}_{vv}(u) = \mathrm{E}[\mathbf{v}(pt+u)\mathbf{v}(pt+u)^H]$. The filter $\mathbf{D}(u) = \Big(\mathbf{R}_{vv}(u) + \widetilde{\mathbf{S}}\widetilde{\mathbf{S}}^H\Big)^\dagger \widetilde{\mathbf{S}}$ solves the problem (4). Moreover, it is able to completely suppress (asymptotically for the MMSE solution) the overall interference if the following conditions are met:*

C1. $\mathrm{Im}(\mathrm{E}[\mathbf{z}(q)\mathbf{z}(q)^H]) \cap \mathrm{Im}(\widetilde{\mathbf{S}}) = \emptyset$, *i.e. the useful signal subspace and the interference subspace are disjoint;*

C2. $\mathrm{E}\left[\mathbf{d}(q)\mathbf{z}(q)^H\right] = \mathbf{O}_{\overline{m}NQr,\overline{m}NQr}$, *i.e. the useful signal and the interferers are uncorrelated.*

In the above Proposition, $(\cdot)^\dagger$ denotes pseudo-inversion, while $\mathrm{Im}(\cdot)$ denotes column-span. [C1] is a generalization to the assumed scenario of the so-called *identifiability condition* [2,3], and leads to the following upper-bound to the maximum number of active users

$$K \leq \min\left(\left\lfloor \frac{(\overline{m}NQ - LM)r + t}{(\overline{m}+\overline{L}-1)t} \right\rfloor, \phi(N)\right), \tag{5}$$

where $\phi(N)$ is the number of available spreading codes, usually tied to the processing gain N. Notice that, for a fixed N, this bound can be relaxed both enlarging the processing window size $\overline{m}T_s$ and, more effectively, increasing the number of receive antennas.

Condition [C2], instead, poses some limitations on the structures of the STBC's that can be actually used. Indeed, [C2] requires that the useful signal be uncorrelated with all of the interferers, i.e. with both MAI and ISI. Uncorrelation with MAI is always satisfied provided that different users transmit independent bit-streams, i.e. in the absence of cooperative encoding. Uncorrelation with ISI, instead, deserves special attention. Assuming that the bit-stream of each user can be described as an i.i.d. process, uncorrelation is always fulfilled in uncoded (whether single antenna or multi-antenna) systems [3, 4]; conversely, if the transmitted information is either temporally encoded (in single antenna systems) or STB encoded (in multiple antenna systems), uncorrelation is no longer ensured, unless interleaving is adopted. Interestingly, it can be proven that the differential Alamouti format satisfies [C2] without interleaving, provided that the uncoded stream is an i.i.d. zero-mean proper process, a requirement that can be fulfilled adopting phase modulations with cardinality greater or equal to 4. Due to its wide-spread application, and since it simplifies the design of the suppression stage ($\mathbf{D}$ is time-invariant and no interleaving is needed), in the remaining part of this paper we focus only on the Alamouti code.

Let us now move on to the space-time decoder. Assuming $n_{\tau_c}=0$ and $n_{\tau_s^0}$ perfectly known (deferring to the next sub-section the problem of their estimation) and assuming differential Alamouti encoding, the signal at the output of the first stage is

$$\mathbf{y}(2p+u) = s_0^0(2p+u)\overline{\mathbf{h}}_0 + s_1^0(2p+u)\overline{\mathbf{h}}_1 + \mathbf{n}(2p+u), \tag{6}$$

where $\overline{\mathbf{h}}_i = \mathbf{U}^H\mathbf{S}^0\mathbf{g}_i^0$, $\mathbf{n}(q)=\mathbf{U}^H\mathbf{w}(q)$ is white noise, and we have neglected the residual interference (this is rigorously true if the ZF-like solution is adopted, while only asymptotically true for the MMSE-like solution [3]).

Relying upon the signals received in four consecutive symbol intervals, namely $\{\mathbf{y}(2p+u),\ u=-2,-1,0,1\}$, a differential space-time decoding rule may be derived paralleling [7]. Direct application of the results of [7] to the assumed dispersive scenario is rigorously not feasible, in that the noise vectors $\{\mathbf{n}(2p+u),\ u=-2,-1,0,1\}$ exhibit a non-zero cross-correlation, since we are considering overlapping processing windows. Nevertheless, neglecting at the design stage such a time correlation and assuming the entries of the unknown equivalent channels $\{\overline{\mathbf{h}}_i,\ i=0,1\}$ in (6) to be independent and Rayleigh distributed, the following differential detection rule can be derived:

$$\begin{pmatrix}\widehat{\mu}^0(2p)\\ \widehat{\mu}^0(2p+1)\end{pmatrix} = \arg\min_{\nu_0,\nu_1\in\mathcal{A}} \left\| \begin{pmatrix}\nu_0\\ \nu_1\end{pmatrix} - \begin{pmatrix}\mathbf{y}^H(2p-2)\mathbf{y}(2p)+\mathbf{y}^H(2p+1)\mathbf{y}(2p-1)\\ \mathbf{y}^H(2p-1)\mathbf{y}(2p)-\mathbf{y}^H(2p+1)\mathbf{y}(2p-2)\end{pmatrix}\right\|. \tag{7}$$

A different decoding strategies can be, instead, derived by following a non-Baeysian approach. In particular, we propose to parallel the design strategy usually adopted in differential $M-$ary phase-shift keying with soft decoding, where *soft maximum-likelihood* estimates of the transmitted uncoded process are first derived and, then, the corresponding hard estimates are obtained by using a minimum distance classification rule [2]. In this case, neglecting the time correlation between the noise vectors $\{\mathbf{n}(2p+u),\ u=-2,-1,0,1\}$, the following differential decoder is obtained:

$$\begin{aligned}\begin{pmatrix}\widehat{\mu}^0(2p)\\ \widehat{\mu}^0(2p+1)\end{pmatrix} = \arg\min_{\nu_0,\nu_1\in\mathcal{A}} \Bigg\| &\begin{pmatrix}\nu_0\\ \nu_1\end{pmatrix} - \\ &\begin{pmatrix}\mathbf{y}^H(2p-2)\mathbf{R}_1\mathbf{y}(2p)+\mathbf{y}^H(2p+1)\mathbf{R}_1\mathbf{y}(2p-1)\\ \mathbf{y}^H(2p-1)\mathbf{R}_1\mathbf{y}(2p)-\mathbf{y}^H(2p+1)\mathbf{R}_1\mathbf{y}(2p-2)\end{pmatrix} - \\ &\begin{pmatrix}\mathbf{y}^H(2p-2)\mathbf{R}_2\mathbf{y}^*(2p+1)+\big(\mathbf{y}^H(2p)\mathbf{R}_2\mathbf{y}^*(2p-1)\big)^*\\ \mathbf{y}^H(2p-1)\mathbf{R}_2\mathbf{y}^*(2p+1)-\big(\mathbf{y}^H(2p)\mathbf{R}_2\mathbf{y}^*(2p-2)\big)^*\end{pmatrix}\Bigg\|, \end{aligned} \tag{8}$$

where $\mathbf{R}_1 = \overline{\mathbf{h}}_0\overline{\mathbf{h}}_0^H + \overline{\mathbf{h}}_1\overline{\mathbf{h}}_1^H$, $\mathbf{R}_2 = \overline{\mathbf{h}}_0\overline{\mathbf{h}}_1^T - \overline{\mathbf{h}}_1\overline{\mathbf{h}}_0^T$.

Comparing (7) and (8), some remarks are in order: a) In both cases, the decisions on the two transmitted symbols have been decoupled, whereby the decoding complexity grows linearly with the constellation size. b) While (7) ignores any information about the useful signal directions $\{\overline{\mathbf{h}}_0, \overline{\mathbf{h}}_1\}$, (8) exploits some knowledge of $\text{Im}([\overline{\mathbf{h}}_0, \overline{\mathbf{h}}_1])$ contained in $\mathbf{R}_1, \mathbf{R}_2$, and the transmitted information is sought for not in the entire space $\mathbb{C}^{LMr}$, but, more effectively, inside the subspace spanned by $\{\overline{\mathbf{h}}_0, \overline{\mathbf{h}}_1\}$. Thus, we expect that this latter decoding rule outperforms the former, since less noise is actually integrated. c) Finally, for synchronous systems equipped with a single receive antenna and operating on frequency flat channels, the observables become scalar quantities: as a consequence, (8) and (7) become equivalent, reducing to the differential decoding rule in [6, 7]

2.1 Blind Implementation

To make the previous schemes fully blind, we have to show that all of the involved parameters can be blindly extracted from the observations. We split the discussion in three parts, i.e., [a] blind implementation of the blocking stage; [b] blind acquisition of the encoder delay; [c] blind implementation of the decoding rule.

As to [a], while $\mathbf{R}_{vv}(u)$ may be easily estimated starting upon $\mathbf{R}_{rr}(u)$ [2, 3], estimation of the matrix $\widetilde{\mathbf{S}}$ in (3) requires more discussion. The problem here is the extraction of the parameter $n_{\tau_s^0} = \lfloor \tau_s^0 Q/T_c \rfloor$, which in turn takes on values in the set $\{0, \dots, N_{sc}Q - 1\}$. Thus, given the $N_{sc}Q$ matrices

$$\widetilde{\mathbf{S}}_\ell = \overline{\mathbf{S}}^{0,0}(:, \ell Mr{+}1 : (\ell{+}L)Mr), \quad \ell = 0, \dots, N_{sc}Q{-}1,$$

as many blocking stages, $\{\mathbf{U}_\ell(u),\ \ell = 0, \dots, N_{sc}Q - 1\}$, can be constructed and the following maximum mean output energy test can be performed: $\mathbf{U} = \arg\max_{\mathbf{U}_\ell} \{\lambda_1(\mathbf{U}_\ell) + \lambda_2(\mathbf{U}_\ell)\}$, where $\{\lambda_i(\mathbf{U}_\ell),\ i = 0, 1\}$ denote the 2 largest eigenvalues of $\mathbf{U}_\ell^H \mathbf{R_{rr}} \mathbf{U}_\ell$.

Let us now move on to the problem [b] of recovering the encoder synchronism. The vectors $\mathbf{y}(q)$ have to be processed in group of $t = 2$ in order to space-time decode the transmitted information; now, there are 2 different ways of casting these vectors together. A systematic approach to the problem is outside the scope of the paper and we just give an *ad hoc* test, relying on the noticeable symmetry properties of the Alamouti STBC shown in the following

Observation 1. *For the Alamouti code, the following relationship holds:*

$$\boldsymbol{\Sigma}_p^0(:,1)\left(\boldsymbol{\Sigma}_p^0(:,2)\right)^T - \boldsymbol{\Sigma}_p^0(:,2)\left(\boldsymbol{\Sigma}_p^0(:,1)\right)^T = \begin{pmatrix} 0 & 1 \\ -1 & 0 \end{pmatrix}, \quad (9)$$

where $\boldsymbol{\Sigma}_p^0(:,i)$ denotes the i-th column of $\boldsymbol{\Sigma}_p^0$. If $\tau_c^0 = 0$, $\boldsymbol{\Sigma}_p^0$ is conveyed by $\mathbf{y}(2p)$ and $\mathbf{y}(2p+1)$; in this case from (9) we have $\text{E}[\mathbf{y}(2p)\mathbf{y}^T(2p{+}1) - \mathbf{y}(2p{+}1)\mathbf{y}^T(2p)] = \overline{\mathbf{h}}_0\overline{\mathbf{h}}_1^T - \overline{\mathbf{h}}_1\overline{\mathbf{h}}_0^T = \mathbf{R}_2$, *and*

$$\text{tr}\left\{\mathbf{R}_2\mathbf{R}_2^H\right\} = \|\overline{\mathbf{h}}_0\|^2\|\overline{\mathbf{h}}_1\|^2 - |\overline{\mathbf{h}}_0^H\overline{\mathbf{h}}_1|^2 \geq 0,$$

where the equality holds with probability zero in our setup (i.e. for a dispersive channel). On the other hand, if $\tau_c^0 = T_s$, $\boldsymbol{\Sigma}_p^0$ is contained in $\mathbf{y}(2p+1)$ and $\mathbf{y}(2p+2)$ and we have: $\text{E}[\mathbf{y}(2p)\mathbf{y}^T(2p+1) - \mathbf{y}(2p+1)\mathbf{y}^T(2p)] = \mathbf{O}_{LMr,LMr}$.

Thus, forming the two sample estimates

$$\mathbf{F}_0 = \frac{1}{B}\sum_{n=0}^{B}\left[\mathbf{y}(2n)\mathbf{y}^T(2n+1) - \mathbf{y}(2n+1)\mathbf{y}^T(2n)\right],$$

$$\mathbf{F}_{T_s} = \frac{1}{B}\sum_{n=0}^{B}\left[\mathbf{y}(2n+1)\mathbf{y}^T(2n+2) - \mathbf{y}(2n+2)\mathbf{y}^T(2n+1)\right],$$

B denoting the estimation sample size, we have that $\mathbf{F}_0$ and $\mathbf{F}_{T_s}$ represent an estimate of $\mathbf{R}_2$ and $\mathbf{O}_{LMr,LMr}$, respectively, if $\tau_c^0 = 0$; on the other hand, if $\tau_c^0 = T_s$, they give an estimate of $\mathbf{O}_{LMr,LMr}$ and $\mathbf{R}_2$, respectively. Thus, task [b] can be easily accomplished through a trace test on $\mathbf{F}_0$ and $\mathbf{F}_{T_s}$.

Finally, task [c] requires a blind estimation of the matrices $\mathbf{R}_1$ and $\mathbf{R}_2$ in (8), which can be obtained as:

$$\widehat{\mathbf{R}}_1 = 2\left(\mathbf{U}^H\widehat{\mathbf{R}}_{rr}\mathbf{U} - \widehat{2\mathcal{N}_0}\mathbf{I}_{LMr}\right),\ \widehat{\mathbf{R}}_2 = \arg\max_{\mathbf{F}\in\{\mathbf{F}_0, \mathbf{F}_{T_s}\}} \text{tr}\left\{\mathbf{F}\mathbf{F}^H\right\}.$$

Notice that, while the batch blind estimates of the matrices $\mathbf{D}$, $\mathbf{R}_1$ and $\mathbf{R}_2$ have to be updated at the beginning of each new data packet (whose length depends upon the Doppler bandwidth), the symbol and the encoder delays may be acquired *una tantum* at the beginning of the transmission and kept as far as the user is active. Thus, a batch fully-blind implementation of the proposed receive strategies has a main complexity $\mathcal{O}\left((\overline{m}NQr)^3\right)$ due to the interference blocking-stage, which is comparable to the complexity of competing blind linear procedures for dispersive single-input multiple-output (SIMO) CDMA systems [4, 5].

	r	1	2	3	4
K_{max}	MIMO ($t = 2$)	9	17	26	31
	SIMO ($t = 1$)	17	31	31	31

Table 1: Maximum users number given by relationship (5) for the proposed blind differentially Alamouti encoded MIMO CDMA system ($t = 2$) and for a blind SIMO CDMA scheme where no transmit diversity is pursued ($t = 1$) [4].

3. NUMERICAL RESULTS

We consider a system with a total bandwidth constraint of $2W = 1.25N/T_s$ and $T_m = 3T_s/N$. The channel linking the i-th transmit antenna of the k-th user to the j-th receive antenna is modeled as $c_{i,j}^k(\tau) = \sum_{l=0}^{\nu} \alpha_{i,j,l}^k \delta(\tau - l/2W)$, with $\nu \simeq \lfloor 2WT_m \rfloor = 3$. Slow Rayleigh fading is assumed and the complex path gains are independently generated according to an exponentially decreasing profile, namely $E[|\alpha_{i,j,l}^k|^2] = 0.65, 0.25, 0.08, 0.02$ for $l = 0, 1, 2, 3$ and $\forall\, i, j, k$. The user and the encoder delays $\{\tau_s^k,\ k = 0, \dots, K-1\}$ and $\{\tau_c^k,\ k = 0, \dots, K-1\}$ are uniformly and independently generated in the interval $[0, T_s)$ and in the set $\{0, T_s, \dots, (t-1)T_s\}$, respectively. We consider $M = 2$ subcarriers, separated by a guard band $B_g = 0.05B_{sc}$, resulting in a subcarrier bandwidth extension of $B_{sc} \simeq 0.61N/T_s$. Notice that since the spacing between the subcarriers $\Delta f = (1{+}0.05)B_{sc} \simeq 0.64N/T_s$ exceeds the coherence bandwidth $B_c \simeq 1/T_m \simeq 0.33N/T_s$ of the channel, each subcarrier substantially experiences independent frequency-selective fading. At the transmitter/receiver side, raised cosine chip waveforms with roll-off factor 0.17, truncated to include the main lobe only, are employed. A minimum processing window size $\overline{m} = \overline{L} = 3$ and $Q = 1$ are selected. The processing gain is $N = 32$ and the spreading sequences are PN sequences of length 31 stretched out with a ± 1. The Alamouti code with a 4-PSK modulation format is adopted, giving a spectral efficiency of $\mathcal{R}/(2W) = 1.6/N$ bits/(sHz). The results are expressed as a function of the *energy contrast* per symbol $\gamma = \mathcal{E}_s/\mathcal{N}_0$, $\mathcal{E}_s$ being the total received energy per symbol. Finally, an MMSE solution is adopted for the first stage and, for the sake of simplicity, the covariance matrix of the received signal is assumed perfectly estimated.

In Fig. 1, we analyze the probability of correct encoder synchronization, P_{sync} versus the estimation-sample size B: $r = 1$, the Interference-to-Signal Ratio (ISR) is 15 dB and $\gamma = 14, 20$ dB. Notice that, as far as the first stage is able to suppress the overall interference (see (5) and Table 1), P_{sync} is fairly close to one, even for moderate values of B. In Fig. 2, we report the Symbol Error Rate (SER) of the proposed system for the two decoding rules in (7) and (8), referred to as *strategy I* and *strategy II*, respectively, for ISR= 0, 15 dB. As expected, the decoding rule (8) always outperforms the one in (7), since more information about the useful signal subspace is exploited. Moreover, notice that, as long as $K < K_{max}$

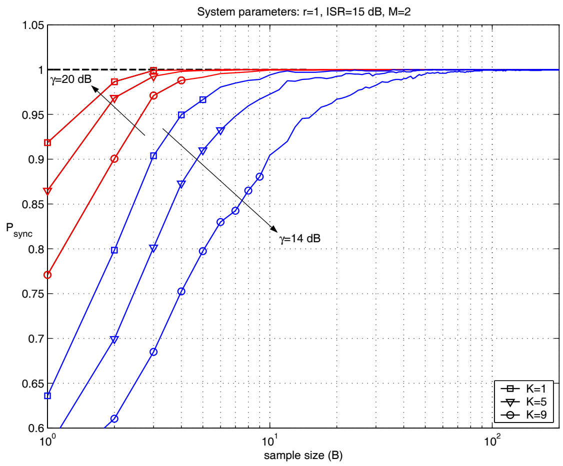

Figure 1: Probability of correct encoder synchronization versus B for $K = 1, 5, 9$ and $\gamma = 14, 20$ dB.

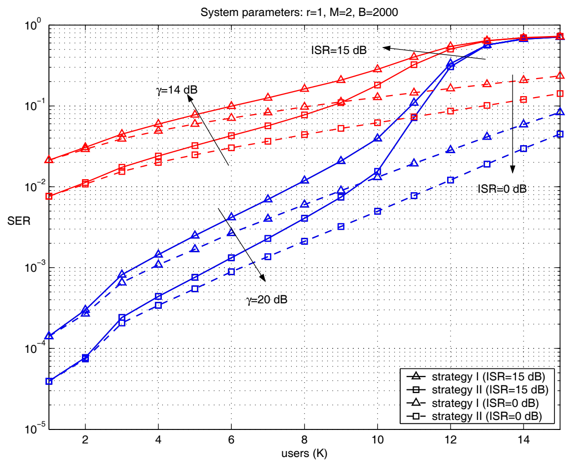

Figure 2: SER versus K for ISR= $0, 15$ dB and $\gamma = 14, 20$ dB. Both the decoding rules in (7) and (8) – referred to as *strategy I* and *strategy II*, respectively – are considered.

(see Table 1), the proposed interference-blocking stage is substantially immune to the presence of strong interfering signals, in agreement with the results in [3, 4]. Finally, Fig. 3 shows the SER of the proposed MIMO CDMA scheme compared to the SER of a fully blind SIMO CDMA system, where no transmit diversity is pursed (i.e., $t = 1$) and the two-stage decoding strategy of [4] is considered: ISR=0 dB, $\gamma = 14, 20$ dB and $r = 1, 2$. It can be seen that, while in lightly-loaded networks the MIMO system can take advantage of the additional transmit spatial diversity to outperform the SIMO one, as K increases, the shortage of interference-free directions severely impairs the system performance, nullifying the transmit diversity advantage and suggesting - eventually - the use of only one transmit antenna. On the other hand, it can be seen that increasing r, i.e. enlarging the signal representation space, has the beneficial effect of moving forward the limitation imposed on the user number by condition [C1] in (5).

4. CONCLUSIONS

The problem of blind decoding in dispersive CDMA MIMO channels has been addressed: the proposed approach unifies in a unique model DS/CDMA and MC-DS/CDMA transmission formats. Each user is assigned one and the same signature to be employed on all of the transmit antennas. The new context rules out any form of uncoded transmission, since the association of the decoded symbols to the corresponding transmitting antenna requires the availability of distinct signatures, labeling the different channels: in the proposed framework, those signatures are provided, in the space-time domain, through STBC. General conditions for blind linear user separability and ISI removal are stated. Focusing on the differential Alamouti code, we show that the encoder synchronism can be acquired by simply resorting to sub-optimum second-order tests, while suboptimum differential decoding rules with complexity only linear in the constellation size can be derived. The performance analysis shows merits and drawbacks of the proposed system. On one hand, when the network load is low, a quasi-full diversity gain is retained, irrespective of the number of receive antennas. Conversely, as the number of users increases, the enhancement of co-channel interference may end up with eating out the whole transmit diversity gain, unless the number of receive antennas is made conveniently large, so as to prevent any signal-space saturation.

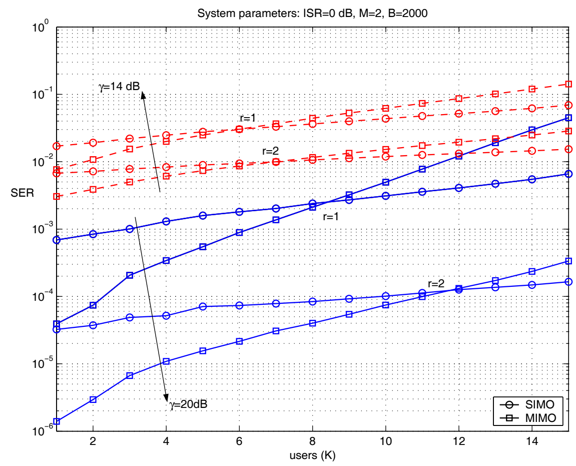

Figure 3: SER versus K, $\gamma = 14, 20$ dB. For the SIMO system, the fully-blind decoding strategy in [4] is adopted. For the MIMO system, differentially Alamouti encoding is assumed ($t = 2$), and the proposed fully-blind decoding rule (8) is employed.

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