Low Complexity Group Detectors for Multirate Transmission in TD-CDMA 3G Systems

Piero Castoldi and Hisashi Kobayashi

Dip. di Ing. dell'Informazione, Università di Parma
Parco Area delle Scienze, 181/A
43100 Parma - ITALY
castoldi@tlc.unipr.it

Department of Electrical Engineering
Princeton University
Princeton, NJ 08544, USA
hisashi@ee.princeton.edu

Abstract— **In this paper, we address the problem of the downlink detection in a mobile radio system with hybrid time-division and code-division multiple access, wherein the deployment of orthogonal variable spreading factor codes allows for multirate communication. Accounting for a multipath propagation channel, we focus on the detection of data carried by a group of codes (ranging from a single code, to all active intracell codes) destined to the same mobile station and on the simultaneous rejection of the interference due to the complementary set of active intracell codes.**

We tackle both the problems of Intersymbol Interference (ISI) suppression, and Multiple Access Interference (MAI) mitigation, and show when it is convenient to treat these issues together. Specifically, the feasibility of linear detectors based on the zero forcing and the minimum mean square error (MMSE) criterion is investigated, with the aim to mitigate both ISI and MAI. A unified low-complexity formulation based on a sliding window algorithm is proposed for this class of linear detectors.

The numerical results validate the proposed receiver structure, indicating that the attainable performance is very close to the single user lower bound up to moderate system loads.

I. Introduction

The 3rd Generation (3G) mobile radio system named International Mobile Communications 2000 (IMT-2000) will be based on Direct Sequence Code Division Multiple Access (DS-CDMA). One of the proposed duplexing techniques to be employed for high data rate applications with picocell coverage is the Time Division Duplex (TDD) mode, which uses separate time slots for the uplink and downlink streams [1].

One of the most interesting features of the 3g system will be the capability to handle services with different specifications in terms of data rates. CDMA is particularly flexible to handle data streams with different bit rates, if the system is designed according to a multirate access strategy. Multirate CDMA is a relatively new topic for research and some efforts have recently started to investigate receivers for such systems [2], [3].

The focus of this paper is on the downlink detection in a system where the chip rate is kept constant (in order to meet a fixed bandwidth requirement) which causes the symbol rate to be variable according to the spreading factor. Since different codes support different services, which may be delivered to the same mobile user, the latter must be able to detect data carried by a subset of all the active codes in the current time slot.

Detection of a synchronous CDMA signal in the presence of a known stationary channel with multipath propagation has already been studied by many authors for the uplink [6]-[8] and for the downlink [9], [10]. Some drawbacks of these previous studies are: (i) a system model which assumes an ideal pulse and no oversampling; and (ii) their implementation as a block multiuser detector [8], [9]. On one hand, in a real system the non-ideal chip pulse and the delay spread of the physical channel may cause significant InterSymbol Interference (ISI) especially when the spreading factor is small; on the other hand, block detection is computationally prohibitive even for burst transmission with short data fields. Some recursive formulations are presented in [8] for detecting all codes, using a decision feedback interference cancellation.

The present work extends previous studies [6]-[10]. We focus on detection of data carried by the group of codes (whose size varies from one to the totality of all active codes) destined to the same mobile user. We investigate four different types of linear group detectors which combat ISI by equalization and MAI by interference mitigation. We should note that two of them, i.e. the Zero Forcing (ZF) detector for suppression of both ISI and MAI and the ZF detector for ISI only, have been already partially investigated in [10]. The structures of the MMSE detector for mitigation of both ISI and MAI and mitigation of ISI only are, on the other hand, proposed here for the first time. When the detector pursues complete (i.e. ZF) or partial (i.e. MMSE) cancellation of ISI only, the decision variable for data detection must be produced by another stage, a conventional detector, which relies on the orthogonality of the spreading codes. Finally, a unified sliding window formulation which holds for any of the proposed linear detectors is derived. The algorithm is implemented by using a bank of FIR filters and symbol by symbol decision devices. A detailed investigation on the software-radio architecture is also provided as well as numerical results which assess the receiver performance.

II. System and channel model

The TDD system specifications envisions a constant chip interval T_c, such that denoting by Q_j the spreading factor for the j-th code, the symbol interval $T_j = Q_j T_c$ may be different for different data streams. We denote

This work was supported, in part, by CNR in the framework of Progetto Finalizzato *Madess II*, subproject "CDMA Interference Mitigation Receiver (CIMR)", and by the New Jersey Center for Wireless Telecommunications (NJCWT).

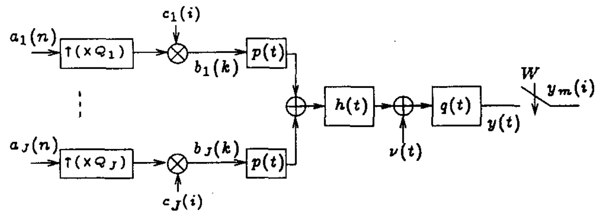

Fig. 1. Model of the considered transmission system

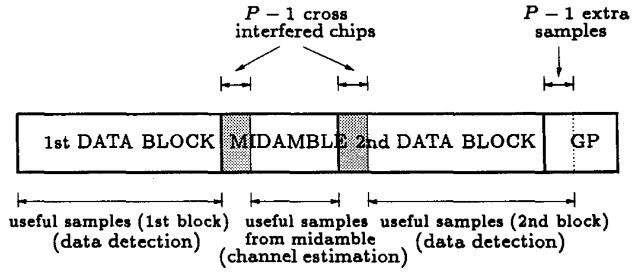

Fig. 2. Identification of the useful samples for detection in the data format

by $\{a_j(n)\}_{n=1}^{N_j}$ the data sequence spread by the j-th code ($j = 1, 2, \ldots, J$) where N_j is the data sequence length. The product $Q_j N_j = L_d$, which measures the length of the data field in multiple of the chip interval T_c, is constant.

The spreading sequence (Walsh-Hadamard) code for the j-th user is denoted by

$$\mathbf{c}_j = [c_j(0), c_j(1), \ldots, c_j(Q_j - 1)]^T, \quad j = 1, \ldots, J, \quad (1)$$

and is repeated each symbol interval. The base-station specific scrambling code superimposed to the spreading codes is not considered in this paper since we are focusing only on intracell interference rejection and we assume that the interference coming from the other cells is negligible.

The low-pass equivalent of the transmission system we consider is shown in Fig. 1. The J synchronous intracell data stream, spread with possible different orthogonal spreading factors, are sent over a slowly-varying multipath fading transmission channel. The received signal is further corrupted by Additive White Gaussian Noise (AWGN). We denote the transmitter and receiver filter impulse response by $p(t)$ and $q(t)$, respectively. Assuming the sampling rate of $W = \frac{\beta}{T_c}$, where β is the oversampling factor, the receiving filter $q(t)$ is chosen as a low pass filter, whose squared frequency response has vestigial symmetry around $W/2$.

The time-continuous complex baseband signal after the low pass filter can be expressed as

$$y(t) = \sum_{j=1}^{J} \sum_{n=1}^{N_j} a_j(n) \sum_{i=0}^{Q_j-1} c_j(i) f(t-(n-1)T_j - iT_c) + v(t) \quad (2)$$

where $v(t)$ is the result of the low pass filtering of $\nu(t)$ by $q(t)$, while $f(t)$ is the total channel impulse response given by the convolution of the chip-shaping pulse $p(t)$ common to all users, with the multipath channel response $h(t) = \sum_l h_l \delta(t - \tau_l)$. In the definition of $h(t)$ we have implicitly assumed that the channel impulse response is constant over the burst duration. Because of time-spreading in both the chip pulse (root raised cosine) and the physical channel, the overall channel response $f(t)$ exhibits a delay spread equal to P chip intervals, where $P > 1$.

We further modify (2) by defining a chip sequence $\{b_j(k)\}$ corresponding to the data sequence $\{a_j(n)\}$ of the j-th user as in [10] (see also Fig. 1):

$$b_j(k) \triangleq a_j(n) c_j(|k-1|_{Q_j}) \quad k = 1, 2, \ldots; \quad n = \left\lceil \frac{k}{Q_j} \right\rceil; \quad (3)$$

where $\lceil x \rceil$ is the smallest integer greater than or equal to x. With this definition, Eq. (2) can be written as follows:

$$y(t) = \sum_{j=1}^{J} \sum_{k=1}^{N_j Q_j} b_j(k) f(t - (k-1)T_c) + v(t) \quad (4)$$

The received signal $y(t)$ is sampled at rate $W = \beta/T_c$. By defining the polyphase representation of a sampled function $g(t)$ by $g_m(i) = g((i\beta - m)/W)$ ($m = 1, 2, \ldots, \beta$ and $i = 1, 2, \ldots$), we have

$$y_m(i) = \sum_{j=1}^{J} \sum_{k=0}^{P-1} f_m(k+1) b_j(i-k) + v_m(i), \quad (5)$$

which holds since the function $f(t)$ is causal and has a finite support. The resulting scalar observation model (5) describes both ISI and MAI.

As a final remark we examine the data format of the TDD standard, which is shown in Fig. 2. It consists of two data blocks with a midamble in between, which is used for channel estimation and other purposes [4], [5]. A silent Guard Period (GP) terminates the data block. From (5), we can see that InterChip Interference (ICI) is introduced after sampling, and consequently ISI among the symbols of the data block arises. The presence of ICI also causes some border effects, namely the overall pulse carrying the last few symbols of the first data block extends into the midamble block causing interference in the first $(P-1)$ chip intervals (shaded intervals in Fig. 2). Similarly, the midamble sequence bumps into the first $(P-1)$ chip intervals of the second data block (the interval is again shaded in Fig. 2). At the end of the second data block the impulse response extends for $(P-1)$ chip intervals into the (silent) GP without being subject to interference. Hence, we have βL_d useful samples for detection in the two data blocks.

A. Discrete-time observation model

We stack βN observation $y_m(i)$ (starting at the i-th chip interval and spanning N chip intervals backward) to form the vector:

$$\mathbf{y}^{(N)}(i)=[y_1(i-N+1),...,y_\beta(i-N+1),...,y_1(i),...,y_\beta(i)]^T. \quad (6)$$

In order to express (6) in terms of samples of the channel, spreading codes and data we define a $\beta N \times (P+N-1)$-dimensional matrix

$$\mathbf{F}^{(N)}=\begin{bmatrix} \mathbf{f}(P-1) & \cdots & \mathbf{f}(0) & & & 0 \\ & \mathbf{f}(P-1) & \cdots & \mathbf{f}(0) & & \\ & & \ddots & & \ddots & \\ 0 & & & \mathbf{f}(P-1) & \cdots & \mathbf{f}(0) \end{bmatrix}, \quad (7)$$

where $\mathbf{f}(k)=[f_1(k+1),\ldots,f_\beta(k+1)]^T, (k=0,...,P-1)$. Then we let $n_j=\left\lceil\frac{P+N+Q_j-2}{Q_j}\right\rceil$, and introduce a symbol-rate ticking index $n=\left\lceil\frac{i}{Q_j}\right\rceil$. We then define vectors $\tilde{\mathbf{c}}_j(n-n_j+1)=[c_j(|i-N-P+1|_{Q_j}),\ldots,c_j(Q_j-1)]^T$, $\tilde{\mathbf{c}}_j(n)=[c_j(0),\ldots,c_j(|i-1|_{Q_j})]^T$, and matrix

$$\mathbf{C}_j^{(N)}(i)\triangleq\begin{bmatrix} \tilde{\mathbf{c}}_j(n-n_j+1) & & & 0 \\ & \mathbf{c}_j & & \\ & & \ddots & \\ & & & \mathbf{c}_j \\ 0 & & & \tilde{\mathbf{c}}_j(n) \end{bmatrix} \quad (8)$$

whose dimension is $(P+N-1)\times n_j$, and $\mathbf{c}_j$ is given by (1). Accordingly, we define the following data string

$$\mathbf{a}_j^{(N)}(i)\triangleq[a_j(n-n_j+1),\ldots,a_j(n-l_j+1),\ldots,a_j(n)]^T. \quad (9)$$

The n_j-dimensional vector $\mathbf{a}_j^{(N)}(i)$ accounts for the j-th user-specific involved sequence in the observation $\mathbf{y}^{(N)}(i)$. The parameter $l_j=\left\lceil\frac{P+Q_j-1}{Q_j}\right\rceil$ highlighted in (9) accounts for the maximum ISI which affects each scalar observation (in fact $l_j=n_j$ when $N=1$). Let us observe that ISI span is larger for a smaller spreading factor and vice versa.

By defining a βN-dimensional vector of noise samples

$$\mathbf{v}^{(N)}(i)=[v_1(i-N+1),...,v_\beta(i-N+1),...,v_1(i),...,v_\beta(i)]^T$$

we can write

$$\mathbf{y}^{(N)}(i)=\sum_{j=1}^{J}\mathbf{F}^{(N)}\cdot\mathbf{C}_j^{(N)}(i)\cdot\mathbf{a}_j^{(N)}(i)+\mathbf{v}^{(N)}(i) \quad (10)$$

Finally defining a $(P+N-1)\times n_t$-dimensional matrix and an n_t-dimensional vector

$$\mathbf{C}^{(N)}(i) \triangleq \left[\mathbf{C}_1^{(N)}(i),\mathbf{C}_2^{(N)}(i),\ldots,\mathbf{C}_J^{(N)}(i)\right], \quad (11)$$

$$\mathbf{a}^{(N)}(i) \triangleq \left[\mathbf{a}_1^{(N)^T}(i),\mathbf{a}_2^{(N)^T}(i),\ldots,\mathbf{a}_J^{(N)^T}(i)\right]^T \quad (12)$$

we obtain

$$\mathbf{y}^{(N)}(i)=\mathbf{F}^{(N)}\cdot\mathbf{C}^{(N)}(i)\cdot\mathbf{a}^{(N)}(i)+\mathbf{v}^{(N)}(i) \quad (13)$$

By defining a $(P+N-1)$-dimensional vector $\mathbf{b}^{(N)}(i)=\mathbf{C}^{(N)}(i)\cdot\mathbf{a}^{(N)}(i)$, and a $(\beta N\times n_t)$-dimensional matrix $\mathbf{G}^{(N)}(i)=\tilde{\mathbf{F}}^{(N)}\cdot\mathbf{C}^{(N)}(i)$, as suggested in [10], we can derive two other useful expressions for the observation vector

$$\begin{aligned}\mathbf{y}^{(N)}(i) &= \mathbf{F}^{(N)}\cdot\mathbf{b}^{(N)}(i)+\mathbf{v}^{(N)}(i) \quad (14)\\ &= \mathbf{G}^{(N)}(i)\cdot\mathbf{a}^{(N)}(i)+\mathbf{v}^{(N)}(i). \quad (15)\end{aligned}$$

The noise term $\mathbf{v}^{(N)}(i)$ is regarded as strictly AWGN since we are focusing on a single cell environment.

III. Proposed Linear Detectors

In this section we present four types of linear detectors which can perform under the assumption that all J spreading codes that are actively used during the detection are known to the mobile user — its own spreading code(s) are notified via the control channel [1], the others being directly detected by a specific processing of the midamble.

The task of the four detectors is as follows: given the observation vector $\mathbf{y}^{(N)}(i)$ (with length βN) whose first element is in the i-th chip and which extends backward up to chip $(i-N+1)$, find the best (according to some criterion) linear detector that detects all the data on which $\mathbf{y}^{(N)}(i)$ depends.

Two types of interference arises in the considered system: ISI due to oversampling and MAI inherent in the CDMA strategy. The former can be alleviated by equalization, the latter by interference mitigation. Equalization is always necessary and to this purpose the channel is estimated using the midamble. The orthogonality of the codes is destroyed only when the channel exhibits multipath propagation.

In order to simplify the notation, we drop the superscript $^{(N)}$ in all vectors and matrices, in the rest of this section.

In two of the proposed detectors, we simultaneously mitigate both MAI and ISI utilizing the ZF or MMSE criterion. Through a linear transformation described by a proper matrix $\mathbf{T}(i)$ (an $n_t\times\beta N$ matrix), we obtain an n_t-dimensional vector of soft decision $\mathbf{x}_\mathrm{a}(i)$ for the transmitted data $\mathbf{a}(i)$

$$\mathbf{x}_\mathrm{a}(i)=\mathbf{T}(i)\cdot\mathbf{y}(i). \quad (16)$$

In the other two schemes, in order to reduce the complexity, we mitigate ICI only (using the ZF or MMSE criterion). Consequently, ISI only is mitigated. Making use of a different transformation $\mathbf{T}(i)$ (a $(P+N-1)\times\beta N$ matrix), we obtain an $(N+P-1)$-dimensional vector of soft decision $\mathbf{x}_\mathrm{b}(i)$ for the chip sequence $\mathbf{b}(i)$:

$$\mathbf{x}_\mathrm{b}(i)=\mathbf{T}(i)\cdot\mathbf{y}(i) \quad (17)$$

Then, an additional linear transformation, which relies on the orthogonality of the spreading sequence, is required to mitigate the MAI. This step is a conventional detection scheme that uses the code matrix $\mathbf{C}(i)$:

$$\mathbf{x}_{\mathbf{a}}(i) = \mathbf{C}^H(i) \cdot \mathbf{x}_{\mathbf{b}}(i) \tag{18}$$

where ^{H} means transpose and complex conjugate. This last step (18) relies on the fact that

$$\mathbf{C}^H(i) \cdot \mathbf{C}(i) = \mathbf{I}_{n_t} \tag{19}$$

due to the orthogonality of the Walsh-Hadamard codes. Note that (19) does not strictly hold except when the detection symbols span the whole block. Otherwise, there are border effects, predictable from the structure of $\mathbf{C}(i)$, which introduce residual MAI into the decision variables for "peripheral" symbols of each user. These variables are excluded from the sliding window detection algorithm.

Eventually, a symbol by symbol decision device produces hard decisions for the transmitted data

$$\hat{\mathbf{a}}(i) = \text{quant}[\mathbf{x}_{\mathbf{a}}(i)]. \tag{20}$$

We now present the interference mitigation schemes which rely on well known criteria.

• **The ZF detector for both MAI and ISI.** The soft decision $\mathbf{x}_{\mathbf{a}\ ZF,MI}(i)$ is determined according to the following minimization [7]:

$$\mathbf{x}_{\mathbf{a}\ ZF,MI}(i) = \arg\min_{\mathbf{x}(i)} ||\mathbf{y}(i) - \mathbf{G}(i) \cdot \mathbf{x}(i)||^2, \tag{21}$$

which can be interpreted according to (16) as a linear transformation described by $\mathbf{T}_{ZF,MI}(i) = \mathbf{G}^{\#}(i)$, where $\mathbf{G}^{\#}(i)$ is the pseudo-inverse of $\mathbf{G}(i)$.

• **ZF detector for ISI only.** The preliminary soft decision $\mathbf{x}_{\mathbf{b}\ ZF,I}(i)$ for the transmitted chip sequence is obtained according to the following minimization

$$\mathbf{x}_{\mathbf{b}\ ZF,I}(i) = \arg\min_{\mathbf{x}(i)} ||\mathbf{y}(i) - \mathbf{F} \cdot \mathbf{x}(i)||^2, \tag{22}$$

which, again, can be interpreted according to (17) as a linear transformation described by $\mathbf{T}_{ZF,I}(i) = \mathbf{F}^{\#}$, where $\mathbf{F}^{\#}$ is the pseudo-inverse of $\mathbf{F}$. Then conventional matched filtering described by (18) is required to derive the final decision variable $\mathbf{x}_{\mathbf{a}\ ZF,I}(i)$.

• **The MMSE detector for mitigation of both MAI and ISI.** The MMSE detector performs the following minimization

$$\mathbf{T}_{MMSE,MI}(i) = \arg\min_{\mathbf{T}(i)} \underset{\substack{\mathbf{a}(i)\\ \mathbf{v}(i)}}{E}\left[\ ||\mathbf{a}(i) - \mathbf{T}(i) \cdot \mathbf{y}(i)||^2\ \right], \tag{23}$$

which leads to the linear transformation described by

$$\mathbf{T}_{MMSE,MI}(i) = [\mathbf{G}^H(i) \cdot \mathbf{G}(i) + \sigma_{\mathbf{v}}^2 \mathbf{R}_{\mathbf{a}}^{-1}(i)]^{-1} \cdot \mathbf{G}^H(i), \tag{24}$$

where $\mathbf{R}_{\mathbf{a}}(i) = E[\mathbf{a}(i) \cdot \mathbf{a}^H(i)]$, and $\mathbf{x}_{\mathbf{a}\ MMSE,MI}(i)$ is determined according to (16).

• **The MMSE detector for mitigation of ISI only.** The following minimization has to be performed:

$$\mathbf{T}_{MMSE,I}(i) = \arg\min_{\mathbf{T}(i)} \underset{\substack{\mathbf{b}(i)\\ \mathbf{v}(i)}}{E}\left[\ ||\mathbf{b}(i) - \mathbf{T}(i) \cdot \mathbf{y}(i)||^2\ \right], \tag{25}$$

which leads to the linear transformation described by

$$\mathbf{T}_{MMSE,I}(i) = [\mathbf{F}^H \cdot \mathbf{F} + \sigma_{\mathbf{v}}^2 \mathbf{R}_{\mathbf{b}}^{-1}(i)]^{-1} \cdot \mathbf{F}^H, \tag{26}$$

where $\mathbf{R}_{\mathbf{b}}(i) = E[\mathbf{b}(i) \cdot \mathbf{b}^H(i)]$. Then conventional matched filtering described by (18) is required to produce the final decision variable $\mathbf{x}_{\mathbf{a}\ MMSE,I}(i)$

The receiving structures derived above seem to imply that data of all the active codes must be detected. However, if only a subset of the active codes must be detected, only relevant FIR filters (i.e. relevant rows of the matrix $\mathbf{T}$) need to be used. Hence, the linear transformation can be defined for an appropriate subspace.

IV. Sliding window detector and software-radio architecture

A processing window definition and a sliding window formulation of the detectors proposed in the previous section are necessary in order to employ them in practical receivers. The following derivation holds for any of the proposed receiver schemes. Also detectors which mitigate only ISI can be accommodated, since the "peripheral" decision variables affected by the border effects are not utilized by the algorithm.

The width of the processing window must be large enough to warrant proper interference mitigation for the data stream with the largest spreading factor Q currently used by the system (which may be smaller than Q_{max}). This design strategy yields a *low rate* linear detector [3], meaning that the detection steps tick with the symbol rate of the slowest user. Since we plan to employ a symbol-by-symbol detection strategy, we select the processing window of size $N = Q + P - 1$ chip intervals such that all the observation samples that carry energy of a symbol spread with the largest spreading factor, are collected and used for the detection. In the following we describe the sliding window algorithm for the first (left) data block (see Fig 2). The algorithm proceeds backward starting from the symbols closest to the midamble.

We first let $M \triangleq \frac{L_d}{Q}$, which represents the number of symbols in the associated data stream with the largest spreading factor, and define :

$$\tilde{\mathbf{y}}(k) = \mathbf{y}^{(Q+P-1)}(kQ + |P-1|_Q), \quad k = M, \ldots, 1 \tag{27}$$

$$\tilde{\mathbf{a}}(k) = \mathbf{a}^{(Q+P-1)}(kQ + |P-1|_Q), \quad k = M, \ldots, 1. \tag{28}$$

Due to the cyclostationarity of the received signal we determine, once per burst, the desired receiver matrix $\tilde{\mathbf{T}}$ by setting $i = Q + |P-1|_Q$. We then obtain $\tilde{\mathbf{x}}_a(k)$, the soft-estimate of $\tilde{\mathbf{a}}(k)$, as

$$\tilde{\mathbf{x}}_a(k) = \tilde{\mathbf{T}} \cdot \tilde{\mathbf{y}}(k), \qquad k = M, M-1, \ldots, 1. \tag{29}$$

For the sake of clarity, note that the argument of $\mathbf{y}(\cdot)$ runs over chip intervals while that of $\tilde{\mathbf{y}}(\cdot)$, $\tilde{\mathbf{a}}(\cdot)$ and $\tilde{\mathbf{x}}_a(\cdot)$ runs over the detection step (wich ticks according to the symbol rate of the slowest user). Denoting $\tilde{n}_j = \left\lceil \frac{2(P-1)+Q_j+Q-1}{Q_j} \right\rceil$, we obtain the length of $\tilde{\mathbf{x}}_a(k)$ as $\tilde{n}_t = \sum_j \tilde{n}_j$, which is the number of potential decision variables for all symbols embraced by the processing window. Nevertheless, for a given processing window $\tilde{\mathbf{y}}(k)$, there are code-specific proper subvectors of $\tilde{\mathbf{x}}_a(k)$ with length $\frac{Q}{Q_j}$, yielding the optimal decision variables for a part of the symbols of the j-th code data stream embraced by $\tilde{\mathbf{y}}(k)$ — the other symbols being optimally detected in the next processing window. This concept will be also clarified by Fig. 3. The value of $\frac{Q}{Q_j}$ can be regarded as the equalizer delay, which is different from code to code.

V. Receivers performance

In this section we present some numerical results to assess the receivers performance. The assumed modulation scheme is binary phase shift keying (BPSK), orthogonal variable spreading factor (OVSF) complying with those suggested in [1] are used. The chip pulse shaping is a root raised cosine with roll-off $\alpha = 0.22$ and the oversampling factor is $\beta = 2$ with respect to the chip rate equal to 3.840 Mchip/s. The multipath channel is a 3 ray channel where the relative power of the three taps are $\sigma^2_{h_0} = 1$ dB, $\sigma^2_{h_1} = -13$dB, and $\sigma^2_{h_2} = -25$ dB and the delays are $\tau_0 = 0$, $\tau_1 = 240$ ns, and $\tau_2 = 490$ ns, respectively. The delay spread of the overall channel (convolution of the multipath channel and chip pulse) is truncated to the value $P = 7$. Due to the short burst used in the transmission, the channel can be regarded as constant over the burst.

A. Multicode detection performance

In order to validate the proposed detection strategy, we have analyzed the the signal to noise and interference ratio (SNIR) affecting the $\tilde{n}_t$ components of the vector $\tilde{\mathbf{x}}_a(k)$ given by (29), which are the decision variables for as many symbols. To this purpose we have considered a case with three active codes, with spreading factors $Q_1 = 4, Q_2 = Q_3 = 8$, the latter being the maximum spreading factor among all the active codes ($Q = 8$). Accordingly, the processing window has size $N = Q+P-1 = 14$, which is the length of vector $\tilde{\mathbf{y}}(k)$. The corresponding decision variables are $\tilde{n}_t = 12$, more precisely $\tilde{n}_1 = 6$ for the 1st data stream and $\tilde{n}_2 = \tilde{n}_3 = 3$ for the 2nd and 3rd data streams.

In Fig. 3 we report the signal to noise and interference ratio (SNIR) on each of the 12 decision variable for two

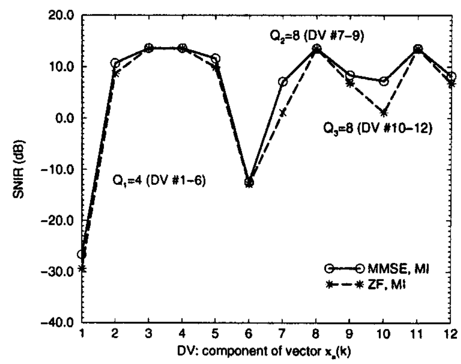

Fig. 3. SNIR as a function of the decision variables (DV) available for a detection scenario with 3 active codes ($Q_1 = 4$, $Q_2 = Q_3 = 8$).

of the proposed detectors, the ZF,MI and the MMSE,MI. The variance of the background noise is set to the value of $\sigma^2_v = -10$dB. Considering the 1st stream ($Q_1 = 4$) it can be seen that the $n_1 = 6$ decision variables exhibit a poor SNIR performance for symbols # 1, 2, 5, 6. The low SNIR is easily justified by considering that the observation vector $\tilde{\mathbf{y}}(k)$ does not collect all the received energy available for the detection of those symbols. Only detection on the $\frac{Q}{Q_1} = 2$ decision variables # 3 and # 4, which has maximum SNIR, is performed. A similar behavior occurs for the two other data streams with $Q_2 = Q_3 = 8$ where only $\frac{Q}{Q_2} = \frac{Q}{Q_3} = 1$ decision variables # 8 and # 11 are retained for detection.

The performance of the ZF,MI is, in general, slightly worse than that of the MMSE,MI.

B. BER performance

In this section we present the evaluation of the average Bit Error Rate (BER) versus E_b/N_0, E_b being the average received energy per bit. The results has been obtained by computer simulation and validated by a theoretical analysis (not reported here), both in closed form and using the Gaussian approximation, as discussed in [11].

In Fig. 4, we present the BER performance of the four types of receiver operating in a scenario where three codes with spreading factor $Q_1 = 4$, $Q_2 = Q_3 = 8$ are active. The BER performance has been evaluated for the data stream of the fastest user ($Q_1 = 4$). The detection algorithms have been tested under two extreme situations: (i) a pure Rice channel with deterministic amplitude of the echoes and (ii) a pure Rayleigh fading channel where the echo amplitudes are assumed to be stochastic but remain unchanged along the whole burst. A perfect channel estimation has been assumed, hence the performance curve can be regarded as a lower bound for any practical receiver.

For a deterministic channel (curves marked with circles) no difference in performance can be seen among the

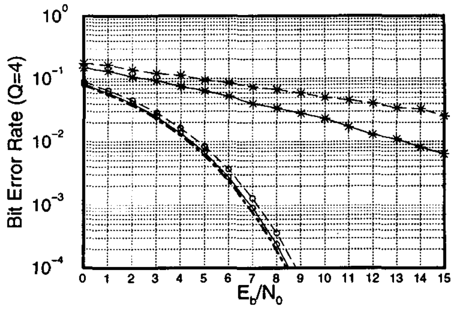

Fig. 4. BER performance of the proposed receivers as a function of $\frac{E_b}{N_0}$ for a deterministic channel and a random channel

ZF,MI, MMSE,MI and MMSE,I and a slight performance degradation is exhibited by the ZF,I receiver. We can conclude that in the presence of a direct propagation path and for the considered multipath profile an equalizer for ISI only could suffice to restore the orthogonality of the codes. Note that all curves are very close to the standard single user lower bound. On the contrary, in the presence of a pure Rayleigh channel the performance is significantly worse than the previous one. The ZF,MI, MMSE,MI and MMSE,I receiver perform approximately the same, equal to the single user bound for fading channel, while the ZF,I performs moderately worse. In this case the orthogonality of the codes is not completely recovered.

In Fig. 5, we present the performance of the the ZF,MI and MMSE,MI receivers, for a constant deterministic channel, under various system loads. The normalized system load is defined as $\rho = \sum_i Q_i^{-1}$ where i runs over all active codes ($0 \leq \rho \leq 1$). The performance is almost independent of the combination of spreading factors used by the system for light up to moderate loads ($\rho \leq 0.5$) and it is almost equivalent to that of the single user lower bound. A performance degradation appears when the load is larger: see for example the case of $\rho = 0.875$.

VI. Conclusions

We have proposed four linear receiver schemes for 3G TD-CDMA radio system to be employed in the downlink detection. The receivers mitigate both ISI and MAI, using the ZF and MMSE criterion. The proposed receivers are grouped into two classes: (i) the first type of receivers restore separation of users by mitigating both ISI and MAI simultaneously; (ii) the second attempts to restore the orthogonality of the codes by simply equalizing the common channel (to eliminate ISI) and relies on a conventional detector for user separation. The complexity of the second class is lower but its performance is significantly degraded when a random channel is encountered. In a deterministic channel the performance of two detectors can be declared

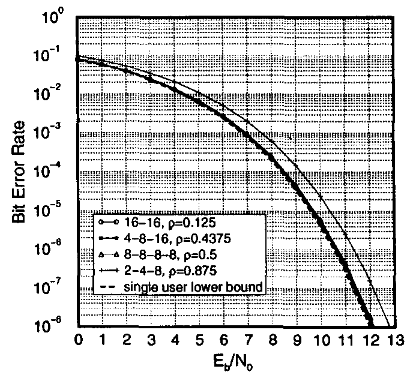

Fig. 5. BER performance of the ZF,MI and MMSE,MI receivers as a function of $\frac{E_b}{N_0}$, for various system loads ρ.

equivalent if perfect channel state information is assumed. The performance approaches that of the single user lower bound, and it is nearly independent of the set of active codes up to moderate system loads.

Beyond the numerical performance assessment a detailed investigation on the software structure of the detection schemes is provided. A relevant sliding window algorithm synchronized with the symbol rate of the slowest data stream is proposed and its fine structure is illustrated.

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