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New Iterative Frequency-Domain Detectors for IA-Precoded MC-CDMA Systems

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Abstract The aim of this paper is to design new multi-user receivers based on the iterative block decision feedback equalization concept for MC-CDMA systems with closed-form interference alignment (IA) at the transmitted side. IA is a promising technique that allows high capacity gains in interfering channels. On the other hand, iterative frequency-domain detection receivers based on the IB-DFE concept can efficiently exploit the inherent space-frequency diversity of the MIMO MC-CDMA systems. In IA-precoded based systems the spatial streams are usually separated by using a standard linear MMSE equalizer. However, for MC-CDMA based systems, linear equalization is not the most efficient way of separating spatial streams due to the residual inter-carrier interference (ICI). Therefore, we design new non-linear iterative receiver structures to efficiently remove the aligned interference and separate the spatial streams in presence of residual ICI. Two strategies are considered: in the first one the equalizer matrices are obtained by minimizing the mean square error (MSE) of each individual data stream at each subcarrier, while in the second approach the matrices are computed by minimizing the overall MSE of all data streams at each subcarrier. We also propose an accurate analytical approach for obtaining the performance of the proposed receivers. Our schemes achieve the maximum degrees of freedom provided by the IA precoding, while allowing close-to-optimum space-diversity gain, with performance approaching the matched filter bound.

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1 Introduction

To achieve high bit rates, needed to meet the quality of service requirements of future multimedia applications, multi-carrier code division multiple access (MC-CDMA) has been considered as an air-interface candidate, especially for downlink [1, 2]. This scheme combines efficiently orthogonal frequency division Multiplex (OFDM) and code division multiple access (CDMA). Therefore, MC-CDMA benefits from OFDM characteristics such as high spectral efficiency and robustness against multi-path propagation, while CDMA allows a flexible multiple access with good interference properties for cellular environments [3–5].

Conventional frequency domain equalization (FDE) schemes usually employ a linear FDE optimized under the minimum mean square error (MMSE) criterion [6]. However, the residual interference levels might be too high, leading to performance that is still several dB from the matched filter bound (MFB) [7, 8]. For this reason, there has been significant interest in the design of nonlinear FDEs in general and iterative FDEs in particular. IB-DFE was originally proposed in [9] and was extended for a wide range of scenarios over the last 10 years, ranging from diversity scenarios, MIMO systems, MC-CDMA and ultra-wideband (UWB) systems, among many other [10–17].

Recent results on the analysis of interference channels (IC) have shown that capacity of an interference channel for a given user is one half the rate of its interference-free capacity in the high transmit power regime, for any number of users [18]. One interesting scheme to efficiently eliminate the co-channel interference and achieve a linear capacity scaling is interference alignment (IA) [18, 19]. With this strategy, more interferers can be completely cancelled than with other interference cancellation methods, thus achieving the maximum degrees of freedom (DoF) [18]. An explicit formulation of the precoding vectors for 3-users achieving IA for time or frequency selectivity channels was presented in [18, 19]. A closed-form solution for constant channels is still unknown, except for 3 users. Due to the difficulty in obtain a closed-form solution for constant channels (with more than 3 users), some iterative algorithms were proposed [20–22]. In [20] several iterative algorithms, designed with only local channel state information (CSI) knowledge, at each node, were presented. A MMSE-based iterative IA scheme was proposed in [21]. Several iterative linear precoding designs using alternating minimization were proposed in [22]. Low complexity feedback strategies for iterative IA-precoded MIMO-OFDM based system were proposed in [23]. Most of these iterative algorithms require significant number of iteration to align the inter-user interference and exchange of information between the transmitter–receiver pairs at each step [22]. An overview of practical issues including performance in realistic propagation environments, the role of CSI at transmitter, and the practicality of IA in large networks was given in [24, 25].

As discussed, IA is a promising technique that allows high capacity gains in interfering channels. On the other hand, iterative frequency-domain detection receivers based on the IB-DFE concept can efficiently exploit the inherent space-frequency diversity of the MIMO MC-CDMA systems. Therefore, in this paper we consider closed form IA precoding at transmitter with IB-DFE at the receiver for MIMO MC-CDMA systems, since this combination allows us to design a system that is able to achieve maximum DoF (number of spatial stream per sub-carrier) and efficiently remove the aligned interference in the presence of ICI, allowing to exploit the high diversity order inherent to these systems. Two strategies are considered: in the first one the equalizer matrices are obtained by minimizing the MSE of each individual data stream at each subcarrier, referred as

iterative successive cancelation (SIC) approach; while in the second one the matrices are computed by minimizing the overall MSE of all data streams at each subcarrier, referred as parallel interference cancellation (PIC) approach. We also propose an accurate analytical approach for obtaining the performance of the proposed receiver structure. Our proposed scheme achieves the maximum degrees of freedom, while allowing a close-to-optimum diversity gain, with only a few iterations at the receiver side.

The remainder of the paper is organized as follows: Sect. 2 presents the system model of the K-user IC MIMO for MC-CDMA systems. In Sect. 3, we briefly start by describing the considered closed-form IA algorithm and then a detail formulation of the proposed receiver structures is presented. Section 4 presents the main results (analytical and numerical). The conclusions will be drawn in Sect. 5.

Notation: Throughout this paper, we will use the following notations. Lowercase letters, bold face lower case letters and boldface uppercase letters are used for scalars, vectors and matrices, respectively. $(.)^H$, $(.)^T$ and $(.)^*$ represents the complex conjugate transpose, transpose, and complex conjugate operators, $\mathbb{E}[.]$ represents the expectation operator, $\mathbf{I}_N$ is the identity matrix of size $N \times N$, $\mathcal{CN}(.,.)$ denotes a circular symmetric complex Gaussian vector, $\{a\}$ represents a L-length block, tr($\mathbf{A}$) is the trace of matrix $\mathbf{A}$ and $\mathbf{e}_p$ is an appropriate row vector with 0 in all positions except the pth position that is 1.

2 System Characterization

We consider a K-user MIMO interference channel with constant coefficients on a per-chip basis. It comprises K transmitter–receiver pairs sharing the same physical channel, where a given transmitter only intends to have its signal decoded by a single receiver. Without loss of generality, we consider a symmetric case where all transmitters and receivers have M antennas, which is denoted by an (M, M, K) interference channel. Since each transmitter is allowed to transmit $P = M/2$ data symbols on each subcarrier, this system has $KM/2$ DoF per-subcarrier. Figure 1 shows the proposed kth MC-CDMA based transmitter. As can be seen, each one of the P L-length data symbols block, $\{d_{k,p,l}; k = 1, \ldots K,$
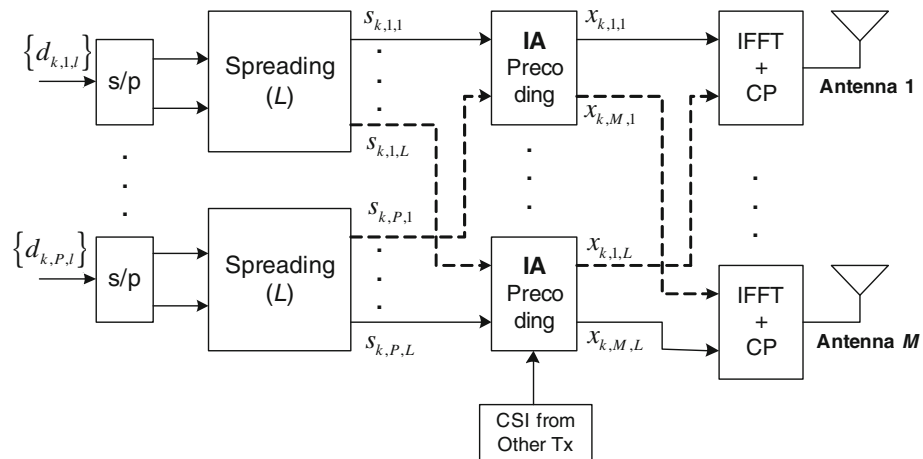

Fig. 1 Proposed MC-CDMA transmitter with interference alignment

$p = 1, \ldots, P$, $l = 0, \ldots, L - 1\}$, where the constellation symbol $d_{k,p,l}$ selected from the data according to given mapping rule (we assume $\mathbb{E}\left[\left|d_{k,p,l}\right|^2\right] = \sigma_d^2$), is spread into L chips using orthogonal Walsh-Hadamard codes, leading to the block $\{s_{k,p,l}; p = 1, \ldots, P,$ $l = 0, \ldots, L - 1\}$. Then, a set of P chips (one of each block) is weighted by an IA-precoding matrix. Note that here the IA-precoding is applied on a chip level instead of data level as in the conventional IA systems. The signal after the IA precoding at the kth transmitter and subcarrier l can be written as

$$\mathbf{x}_{k,l} = \mathbf{W}_{k,l}\mathbf{s}_{k,l}, \tag{1}$$

where $\mathbf{W}_{k,l} \in \mathbb{C}^{MxP}$ is the linear precoding matrix computed at the kth transmitter on subcarrier l, constrained to $\left\|\mathbf{W}_{k,l}\right\|_F^2 \leq T_p$, where T_p is the transmit power at the transmitters, with $\mathbf{s}_{k,l} = [s_{k,1,l} \ldots s_{k,P,l}]^T$. Finally, the precoding signals are mapped into the OFDM symbol and a suitable cyclic prefix (CP) is inserted.

The received frequency-domain signal (i.e., after cyclic prefix removal and FFT operation) for the kth receiver and the lth subcarrier is given by

$$\bar{\mathbf{y}}_{k,l} = \mathbf{H}_{k,k,l}\mathbf{W}_{k,l}\mathbf{s}_{k,l} + \sum_{\substack{j=1 \\ j \neq k}}^{K} \mathbf{H}_{k,j,l}\mathbf{W}_{j,l}\mathbf{s}_{j,l} + \mathbf{n}_{k,l}, \tag{2}$$

provided that the cyclic prefix is long enough to account for different overall channel impulse responses between the transmitters and the receivers (i.e., including transmit and receive filters, multipath propagation effects and differences in the time-of-arrival for different transmitter-to-receiver links).

The size-$M \times M$ matrix $\mathbf{H}_{k,j,l}$ denotes the overall channel between the transmitter j and receiver k on subcarrier l and $\mathbf{n}_{k,l}$ is the additive white Gaussian noise (AWGN) vector at receiver k on subcarrier l, i.e., $n_{k,l} \sim \mathcal{CN}\left(0, \sigma_n^2 I_M\right)$.

3 Receivers Design

In this section we start by briefly reviewing the closed-form IA precoding algorithm for $K \leq 3$ and the standard MMSE based equalizer used to separate the spatial streams. Then the proposed iterative frequency domain equalizers, based on IB-DFE principle, are derived in detail.

3.1 Close-form IA precoder for K = 3

For the three user interference channel it is possible to find a closed-form solution to $\mathbf{W}_{k,l}, k = 1, 2, 3$. It is shown in [18] that the solution for subcarrier l is given by

$$\begin{aligned} \mathbf{W}_{1,l} &= \text{eigvec}_{M/2}\left(\mathbf{H}_{3,1,l}^{-1}\mathbf{H}_{3,2,l}\mathbf{H}_{1,2,l}^{-1}\mathbf{H}_{1,3,l}\mathbf{H}_{2,3,l}^{-1}\mathbf{H}_{2,1,l}\right) \\ \mathbf{W}_{2,l} &= \mathbf{H}_{3,2,l}^{-1}\mathbf{H}_{3,1,l}\mathbf{W}_{1,l} \\ \mathbf{W}_{3,l} &= \mathbf{H}_{2,3,l}^{-1}\mathbf{H}_{2,1,l}\mathbf{W}_{1,l} \end{aligned} , \tag{3}$$

where $\text{eigvec}_{M/2}$ are the $M/2$ eigenvectors of matrix $\left(\mathbf{H}_{3,1,l}^{-1}\mathbf{H}_{3,2,l}\mathbf{H}_{1,2,l}^{-1}\mathbf{H}_{1,3,l}\mathbf{H}_{2,3,l}^{-1}\mathbf{H}_{2,1,l}\right)$.

Up to now, there are no closed-form solutions for MIMO channels with $K > 3$, and some iterative solutions were derived in [20–22]. In this work we consider closed form IA precoding schemes, since the design of a joint iterative IA precoding with iterative frequency domain equalizers is too complex.

3.2 Standard Linear IA MMSE Equalizer

As discussed above, usually a standard MMSE linear equalizer is employed to mitigate the aligned user's interference and separate the spatial streams. Thus, the signal after the equalization process is given by

$$\tilde{y}_{k,l} = \mathbf{G}_{k,l}\mathbf{H}_{k,k,l}\mathbf{W}_{k,l}\mathbf{s}_{k,l} + \mathbf{G}_{k,l}\sum_{\substack{j=1\\j\neq k}}^{K}\mathbf{H}_{k,j,l}\mathbf{W}_{j,l}\mathbf{s}_{j,l} + \mathbf{G}_{k,l}\mathbf{n}_{k,l}, \tag{4}$$

where $\mathbf{G}_{k,l} \in \mathbb{C}^{PxM}$ denotes the linear receiving filter. To compute it, we first need to obtain the matrix

$$\bar{\mathbf{G}}_{k,l} = \left(\mathbf{H}_{t,k,l}^{H}\mathbf{H}_{t,k,l} + \sigma_n^2\mathbf{I}_{3P}\right)^{-1}\mathbf{H}_{t,k,l}^{H}, \tag{5}$$

where $\mathbf{H}_{t,k,l} = \left[\mathbf{H}_{k,1,l}\mathbf{W}_{1,l}\mathbf{H}_{k,2,l}\mathbf{W}_{2,l}\mathbf{H}_{k,3,l}\mathbf{W}_{3,l}\right] \in \mathbb{C}^{Px3M}$. The left pseudo-inverse exist since $\mathbf{H}_{t,k,l}$ has M column rank due to the IA processing. From (5), the linear filter used at the kth receiver on lth sub-carrier is given by

$$\mathbf{G}_{k,l} = \left[g_{k,\frac{(k-1)M}{2}+1,l}^{T} \quad \mathbf{g}_{k,\frac{(k-1)M}{2}+2,l}^{T}\cdots\mathbf{g}_{k,\frac{(k-1)M}{2}+\frac{M}{2},l}^{T}\right]^{T}, \tag{6}$$

where $\mathbf{g}_{k,j,l}$ is the jth row vector of $\bar{\mathbf{G}}_{k,l}$.

3.3 Iterative Equalizers Design

It is well known that for MC-CDMA based systems, standard linear equalization is not the best strategy to separate the spatial streams due to the residual inter-carrier interference. To overcome this problem, we design new non-linear frequency domain equalizers based on IB-DFE principle, which also takes into account both the inter-user aligned interference and residual ICI. Two approaches are considered: IB-DFE PIC and IB-DFE SIC.

3.3.1 IB-DFE PIC Approach

Figure 2 shows the main blocks of the IB-DFE PIC based procedure. For each iteration we detect all P L-length data block of the kth receiver, in a parallel way, using the most updated estimated of the transmit data symbols to cancel the residual interference, which it could not be cancelled in the first equalizer block. Therefore, our receiver can be regarded as an iterative parallel interference cancellation. However, as with conventional IB-DFE based receivers, we take into account the reliability of the block data estimates for each detection procedure.
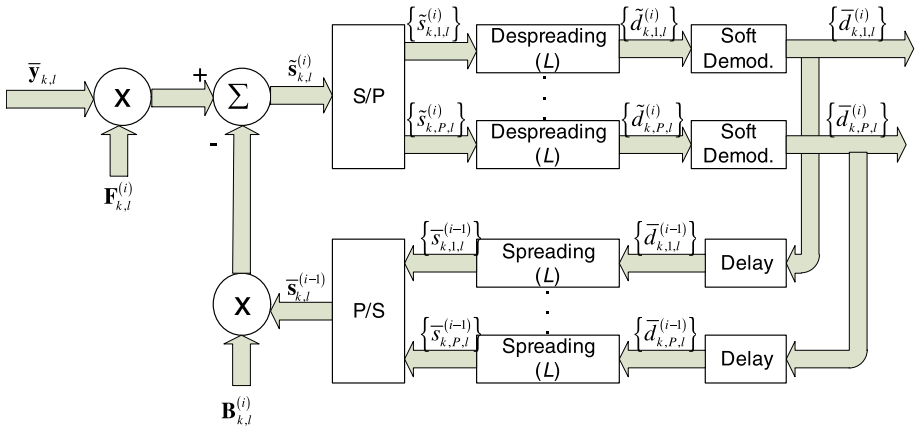
Fig. 2 Iterative receiver PIC equalizer based on IB-DFE principle

At the ith iteration, the signal at kth receiver on lth subcarrier, before the despreading operation is given by

$$\tilde{\mathbf{s}}_{k,l}^{(i)} = \mathbf{F}_{k,l}^{(i)}\bar{\mathbf{y}}_{k,l} - \mathbf{B}_{k,l}^{(i)}\bar{\mathbf{s}}_{k,l}^{(i-1)} \tag{7}$$

where $\mathbf{F}_{k,l}^{(i)} \in \mathbb{C}^{P \times M}$ denoting the feedforward matrix, $\mathbf{B}_{k,l}^{(i)} \in \mathbb{C}^{P \times P}$ is the feedback matrix and $\bar{\mathbf{s}}_{k,l}^{(i-1)} = \left[\bar{s}_{k,1,l}^{(i-1)} \ldots \bar{s}_{k,P,l}^{(i-1)}\right]^T$, where the block $\left\{\bar{s}_{k,p,l}^{(i-1)}; l = 0, \ldots, L-1\right\}$ is the spreading of the pth de-spreading block average values conditioned to the detector output $\left\{\bar{d}_{k,p,l}^{(i)}; l = 0, \ldots, L-1\right\}$ for each iteration i. It can be shown that $\bar{\mathbf{s}}_{k,l}^{(i-1)} \approx \boldsymbol{\Psi}_k^{(i-1)2}\mathbf{s}_{k,l} + \boldsymbol{\Psi}_k^{(i-1)}\boldsymbol{\Delta}_{k,l}$, where $\boldsymbol{\Delta}_{k,l}$ is a zero mean error vector. $\boldsymbol{\Psi}_k^{(i)} = diag\left[\psi_{k,1}^{(i)} \ldots \psi_{k,P}^{(i)}\right]$ is a diagonal correlation matrix computed at ith iteration, where the correlation coefficient computed for the pth L-length data block is

$$\psi_{k,p}^{(i)} = \frac{\mathbb{E}\left[\hat{d}_{k,p,l}^{(i)}d_{k,p,l}^*\right]}{\mathbb{E}\left[\left|d_{k,p,l}\right|^2\right]}, \quad l = 1, \ldots, L, \tag{8}$$

and represents a measure of the reliability of the estimates of the pth L-length data block associated to the ith iteration that can be obtained as described in [13] for QPSK modulations and in [16] for larger constellations. Considering QPSK modulations, the hard decision $\hat{d}_{k,p,l}^{(i)}$ associated to the data symbol $d_{k,p,l}$ is $\hat{d}_{k,p,l}^{(i)} = sign\left(\text{Re}\left(\tilde{d}_{k,p,l}^{(i)}\right)\right) + jsign\left(\text{Im}\left(\tilde{d}_{k,p,l}^{(i)}\right)\right)$.

For a given iteration and at each receiver, the iterative non-linear equalizer is characterized by the coefficients $\mathbf{F}_{k,l}^{(i)}$ and $\mathbf{B}_{k,l}^{(i)}$. These coefficients are computed to minimize the average bit error rate (BER) of all P streams and for a QPSK constellation with Gray mapping the BER of the kth receiver, can be approximately given

$$\text{BER}_k^{(i)} \approx Q\left(\sqrt{\frac{P}{\frac{1}{L}\sum_{l=0}^{L-1}\text{MSE}_{k,l}^{(i)}}}\right), \tag{9}$$

where $Q(x)$ denotes the well-known Gaussian function and $\text{MSE}_{k,l}^{(i)}$ is the overall mean square error of the spreading samples given by,

$$
\begin{aligned}
\text{MSE}_{k,l}^{(i)} &= \mathbb{E}\left[\left\|\tilde{\mathbf{s}}_{k,l}^{(i)} - \mathbf{s}_{k,l}\right\|^2\right] \\
&= \mathbb{E}\left[\text{tr}\left(\left(\tilde{\mathbf{s}}_{k,l}^{(i)} - \mathbf{s}_{k,l}\right)\left(\tilde{\mathbf{s}}_{k,l}^{(i)} - \mathbf{s}_{k,l}\right)^H\right)\right].
\end{aligned}
\tag{10}
$$

After some straightforward but lengthy mathematical manipulations, it can be shown that (10) is reduced as,

$$
\begin{aligned}
\text{MSE}_{k,l}^{(i)} &= \text{tr}\left(\mathbf{F}_{k,l}^{(i)}\mathbf{R}_{k,l}^{\bar{\mathbf{y}}_{k,l}}\mathbf{F}_{k,l}^{(i)H}\right) + \text{tr}\left(\mathbf{B}_{k,l}^{(i)}\mathbf{R}_k^{(i-1)\bar{\mathbf{s}}_k,\mathbf{s}_k}\mathbf{B}_{k,l}^{(i)H}\right) + P\sigma_d^2 - 2\text{tr}\left(\text{Re}\left\{\mathbf{R}_{k,l}^{\bar{\mathbf{y}}_{k,l},\mathbf{s}_k}\mathbf{F}_{k,l}^{(i)H}\right\}\right) \\
&\quad + 2\text{tr}\left(\text{Re}\left\{\mathbf{R}_k^{(i-1)\bar{\mathbf{s}}_k,\mathbf{s}_k}\mathbf{B}_{k,l}^{(i)H}\right\}\right) - 2\text{tr}\left(\text{Re}\left\{\mathbf{B}_{k,l}^{(i)}\mathbf{R}_{k,l}^{(i-1)\bar{\mathbf{s}}_k,\mathbf{y}_{k,l}}\mathbf{F}_{k,l}^{(i)H}\right\}\right)
\end{aligned}.
\tag{11}
$$

The different correlation matrices of (11) are given by,

$$
\begin{cases}
\mathbf{R}_{k,l}^{\bar{\mathbf{y}}_{k,l}} = \mathbb{E}\left[\bar{\mathbf{y}}_{k,l}^*\bar{\mathbf{y}}_{k,l}^T\right] = \mathbf{H}_{k,k,l}^{eq}\mathbf{R}_d\mathbf{H}_{k,k,l}^{eqH} + \sum_{j=1,j\neq k}^{K}\mathbf{H}_{k,j,l}^{eq}\mathbf{R}_d\mathbf{H}_{k,j,l}^{eqH} + \mathbf{R}_n \\
\mathbf{R}_k^{(i-1)\bar{\mathbf{s}}_k,\mathbf{s}_k} = \mathbb{E}\left[\bar{\mathbf{s}}_k^{(i-1)*}\bar{\mathbf{s}}_k^{(i-1)T}\right] = \mathbf{\Psi}_k^{(i-1)^2}\mathbf{R}_d \\
\mathbf{R}_{k,l}^{\bar{\mathbf{y}}_{k,l},\mathbf{s}_k} = \mathbb{E}\left[\bar{\mathbf{y}}_{k,l}^*\mathbf{s}_k\right] = \mathbf{R}_d\mathbf{H}_{k,k,l}^{eqH} \\
\mathbf{R}_k^{(i-1)\bar{\mathbf{s}}_k,\mathbf{s}_k} = \mathbb{E}\left[\bar{\mathbf{s}}_k^{(i-1)*}\mathbf{s}_k\right] = \mathbf{\Psi}_k^{(i-1)^2}\mathbf{R}_d \\
\mathbf{R}_{k,l}^{(i-1)\bar{\mathbf{s}}_k,\mathbf{y}_{k,l}} = \mathbb{E}\left[\bar{\mathbf{s}}_k^{(i-1)*}\bar{\mathbf{y}}_{k,l}\right] = \mathbf{\Psi}_k^{(i-1)^2}\mathbf{R}_d\mathbf{H}_{k,k,l}^{eqH}
\end{cases}, \tag{12}
$$

where $\mathbf{H}_{k,j,l}^{eq} = \mathbf{H}_{k,j,l}\mathbf{W}_{j,l} \in \mathbb{C}^{MxP}$ represents the equivalent channels after the IA procedure. $\mathbf{R}_d = \sigma_d^2\mathbf{I}_P$ and $\mathbf{R}_n = \sigma_n^2\mathbf{I}_M$, represent the correlation matrices of data symbols and residual noise, respectively.

From (9), it is clear that to minimize the average BER at each receiver we need to minimize the overall MSE at each subcarrier. However, only considering the MSE minimization may lead to biased estimates and thus to avoid it we force the overall received amplitude to a constant, i.e. $\frac{1}{L}\sum_{l=0}^{L-1}\text{tr}\left(\mathbf{F}_{k,l}^{(i)}\mathbf{H}_{k,l}^{eq}\right) = P$. Thus constrained optimization problem can be formulated as,

$$\min_{\mathbf{F}_{k,l}^{(i)},\mathbf{B}_{k,l}^{(i)}}\text{MSE}_{k,l}^{(i)} \text{ s.t } \frac{1}{L}\sum_{l=0}^{L-1}\text{tr}\left(\mathbf{F}_{k,l}^{(i)}\mathbf{H}_{k,l}^{eq}\right) = P. \tag{13}$$

We use the Karush–Kuhn–Tucker (KKT) conditions to solve the optimization at each step with all but one variable fixed [26]. The Lagrangian associated with this problem can be written by

$$\mathbb{L}\left(\mathbf{F}_{k,l}^{(i)}, \mathbf{B}_{k,l}^{(i)}, \mu_k^{(i)}\right) = \mathrm{MSE}_{k,l}^{(i)} - \mu_k^{(i)}\left(\frac{1}{L}\sum_{l=0}^{L-1}\mathrm{tr}\left(\mathbf{F}_{k,l}^{(i)}\mathbf{H}_{k,l}^{eq}\right) - P\right), \qquad (14)$$

where μ_k is the Lagrangian multiplier [27]. The KKT conditions are

$$\begin{cases} \nabla_{\mathbf{F}_{k,l}^{(i)}}\mathbb{L}\left(\mathbf{F}_{k,l}^{(i)}, \mathbf{B}_{k,l}^{(i)}, \mu_k^{(i)}\right) = 0 \\ \nabla_{\mathbf{B}_{k,l}^{(i)}}\mathbb{L}\left(\mathbf{F}_{k,l}^{(i)}, \mathbf{B}_{k,l}^{(i)}, \mu_k^{(i)}\right) = 0 \\ \dfrac{1}{L}\sum_{l=0}^{L-1}\mathrm{tr}\left(\mathbf{F}_{k,l}^{(i)}\mathbf{H}_{k,l}^{eq}\right) - P = 0 \end{cases} \qquad (15)$$

After a lengthy but straightforward mathematical manipulation we obtain the feedforward and feedback matrices with the iterative index dependence, given by

$$\mathbf{F}_{k,l}^{(i)} = \mathbf{\Omega}_k^{(i)}\mathbf{H}_{k,k,l}^{eqH}\left(\mathbf{H}_{k,k,l}^{eq}\left(\mathbf{I}_P - \mathbf{\Psi}_k^{(i-1)^2}\right)\mathbf{H}_{k,k,l}^{eqH} + \sum_{j=1,j\neq k}^{K}\mathbf{H}_{k,j,l}^{eq}\mathbf{R}_d\mathbf{H}_{k,j,l}^{eqH} + \frac{\sigma_n^2}{\sigma_d^2}\mathbf{I}_M\right)^{-1}, \qquad (16)$$

and

$$\mathbf{B}_{k,l}^{(i)} = \mathbf{F}_{k,l}^{(i)}\mathbf{H}_{k,k,l}^{eq} - \mathbf{I}_P, \qquad (17)$$

with

$$\mathbf{\Omega}_k^{(i)} = \left(\mathbf{I}_P - \mathbf{\Psi}_k^{(i-1)^2}\right) - \frac{\mu_k^{(i)}}{\sigma_d^2 L}\mathbf{I}_P. \qquad (18)$$

The Lagrangian multiplier is selected, at each iteration i, to ensure that the constraint $\frac{1}{L}\sum_{l=0}^{L-1}\mathrm{tr}\left(\mathbf{F}_{k,l}^{(i)}\mathbf{H}_{k,k,l}^{eq}\right) = P$ is fulfilled. It should be pointed out that for the first iteration $(i = 1)$, $\mathbf{\Psi}_k^{(0)}$ is a null matrix and $\bar{\mathbf{s}}_{k,l}^{(0)}$ is a null vector and thus this iterative receiver reduces to the standard linear MMSE discussed in Sect. 3.2).

3.3.2 IB-DFE SIC Approach

Figure 3 shows the main blocks of the IB-DFE SIC based procedure. For each iteration we detect all P L-length data block of the kth receiver, in a successive way, using the most updated estimated of the transmit data symbols to cancel the residual interference. Now, at the ith iteration, the signal at kth receiver on lth subcarrier associated to the pth$L$-length data block, before the despreading operation is given by

$$\tilde{s}_{k,p,l}^{(i)} = \mathbf{f}_{k,p,l}^{(i)}\bar{\mathbf{y}}_{k,l} - \mathbf{b}_{k,p,l}^{(i)}\bar{\mathbf{s}}_{k,p,l}^{(i-1)} \qquad (19)$$

where $\mathbf{f}_{k,p,l}^{(i)} \in \mathbb{C}^{1xM}$ and $\mathbf{b}_{k,p,l}^{(i)} \in \mathbb{C}^{1xP}$ denoting the feedforward and feedback vectors coefficients of the pth data block applied on the lth subcarrier at the kth user, and $\bar{\mathbf{s}}_{k,p,l}^{(i-1)} = \left[\bar{s}_{k,1,l}^{(i)}, \ldots, \bar{s}_{k,p-1,l}^{(i)}, \bar{s}_{k,p,l}^{(i-1)}, \ldots, \bar{s}_{k,P,l}^{(i-1)}\right]^T$.
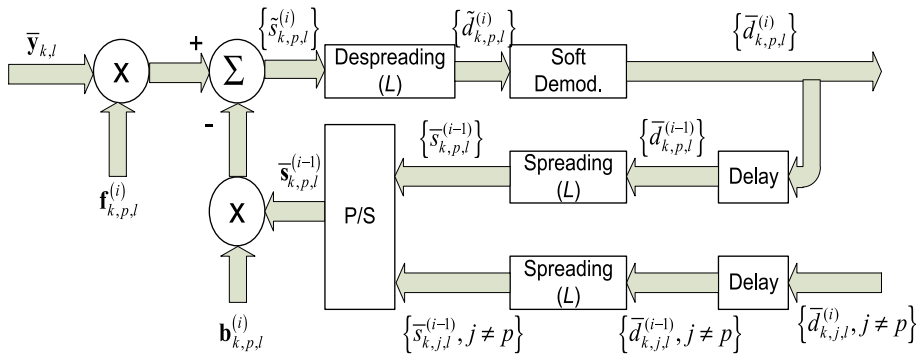
Fig. 3 Iterative receiver SIC equalizer based on IB-DFE principle

Contrarily to the PIC approach, to compute these vectors, we need to minimize the $\text{MSE}_{k,p,l}^{(i)} = \mathbb{E}\left[\left|\tilde{s}_{k,p,l}^{(i)} - s_{k,p,l}\right|^2\right]$ of the spreading samples associated to the pth L-length data block on lth subcarrier. After some mathematical manipulations, it can be shown that $\text{MSE}_{k,p,l}^{(i)}$ is reduced as,

$$\begin{aligned}
\text{MSE}_{k,p,l}^{(i)} = \; & \mathbf{f}_{k,p,l}^{(i)}\mathbf{R}_{k,l}^{\bar{\mathbf{y}}_{k,l}}\mathbf{f}_{k,p,l}^{(i)H} + \mathbf{b}_{k,p,l}^{(i)}\mathbf{R}_k^{(i-1)\bar{\mathbf{s}}_k,\mathbf{s}_k}\mathbf{b}_{k,p,l}^{(i)H} + \sigma_d^2 - 2\text{Re}\left\{\mathbf{R}_{k,l}^{\bar{\mathbf{y}}_{k,l},\mathbf{s}_k}\mathbf{f}_{k,p,l}^{(\hat{i})H}\right\} \\
& + 2\text{Re}\left\{\mathbf{R}_k^{(i-1)\bar{\mathbf{s}}_k,\mathbf{s}_k}\mathbf{b}_{k,p,l}^{(i)H}\right\} - 2\text{Re}\left\{\mathbf{b}_{k,p,l}^{(i)}\mathbf{R}_{k,l}^{(i)\bar{\mathbf{s}}_k,\mathbf{y}_{k,l}}\mathbf{f}_{k,p,l}^{(i)H}\right\}
\end{aligned} \tag{20}$$

The different correlation matrices in (20) are given by,

$$\begin{cases}
\mathbf{R}_{k,l}^{\bar{\mathbf{y}}_{k,l}} = \mathbb{E}\left[\bar{\mathbf{y}}_{k,l}^*\bar{\mathbf{y}}_{k,l}^T\right] = \mathbf{H}_{k,k,l}^{eq}\mathbf{R}_d\mathbf{H}_{k,k,l}^{eqH} + \displaystyle\sum_{j=1,j\neq k}^{K}\mathbf{H}_{k,j,l}^{eq}\mathbf{R}_d\mathbf{H}_{k,j,l}^{eqH} + \mathbf{R}_n \\
\mathbf{R}_k^{(i-1)\bar{\mathbf{s}}_k,\mathbf{s}_k} = \mathbb{E}\left[\bar{\mathbf{s}}_k^{(i-1)*}\bar{\mathbf{s}}_k^{(i-1)T}\right] = \boldsymbol{\Psi}_k^{(i-1)^2}\mathbf{R}_d \\
\mathbf{R}_{k,l}^{\bar{\mathbf{y}}_{k,l},\mathbf{s}_k} = \mathbb{E}\left[\bar{\mathbf{y}}_{k,l}^*\mathbf{s}_k\right] = \mathbf{e}_p\mathbf{R}_d\mathbf{H}_{k,k,l}^{eqH} \\
\mathbf{R}_k^{(i-1)\bar{\mathbf{s}}_k,\mathbf{s}_k} = \mathbb{E}\left[\bar{\mathbf{s}}_k^{(i-1)*}\mathbf{s}_k\right] = \mathbf{e}_p\boldsymbol{\Psi}_k^{(i-1)^2}\mathbf{R}_d \\
\mathbf{R}_{k,l}^{(i-1)\bar{\mathbf{s}}_k,\mathbf{y}_{k,l}} = \mathbb{E}\left[\bar{\mathbf{s}}_k^{(i-1)*}\bar{\mathbf{y}}_{k,l}\right] = \boldsymbol{\Psi}_k^{(i-1)^2}\mathbf{R}_d\mathbf{H}_{k,k,l}^{eqH}
\end{cases} \tag{21}$$

The optimization problem for this approach can be formulated as

$$\min_{\mathbf{f}_{k,p,l}^{(i)},\mathbf{b}_{k,p,l}^{(i)}} \text{MSE}_{k,p,l}^{(i)} \; \text{s.t} \; \frac{1}{L}\sum_{l=0}^{L-1}\mathbf{f}_{k,p,l}^{(i)}\mathbf{h}_{k,k,p,l}^{eq} = 1, \tag{22}$$

where $\mathbf{h}_{k,k,p,l}^{eq} \in \mathbb{C}^{Mx1}$ is the equivalent channel associated to the pth data block. Following the same procedure as for the PIC approach, the feedforward and feedback vectors can be obtained

$$\mathbf{f}_{k,p,l}^{(i)} = \mathbf{\Omega}_{k,p}^{(i)} \mathbf{H}_{k,k,l}^{eqH} \left(\mathbf{H}_{k,k,l}^{eq} \left(\mathbf{I}_P - \mathbf{\Psi}_k^{(i-1)^2} \right) \mathbf{H}_{k,k,l}^{eqH} + \sum_{j=1,j\neq k}^{K} \mathbf{H}_{k,j,l}^{eq} \mathbf{R}_d \mathbf{H}_{k,j,l}^{eqH} + \frac{\sigma_n^2}{\sigma_d^2} \mathbf{I}_M \right)^{-1}, \quad (23)$$

and

$$\mathbf{b}_{k,p,l}^{(i)} = \mathbf{f}_{k,p,l}^{(i)} \mathbf{H}_{k,k,l}^{eq} - \mathbf{e}_p, \quad (24)$$

with

$$\mathbf{\Omega}_{k,p}^{(i)} = \mathbf{e}_p \left(\mathbf{I}_P - \mathbf{\Psi}_k^{(i-1)^2} \right) - \frac{\mu_{k,p}^{(i)}}{\sigma_d^2 L} \mathbf{e}_p. \quad (25)$$

The Lagrangian multiplier is selected, at each iteration i, to ensure that the constraint $\frac{1}{L}\sum_{l=0}^{L-1} \mathbf{f}_{k,p,l}^{(i)} \mathbf{h}_{k,k,p,l}^{eq} = 1$ is fulfilled. It should be pointed out that for the first iteration ($i = 1$) and for the first L-length data block to be detected, $\mathbf{\Psi}_k^{(0)}$ is a null matrix and $\bar{\mathbf{s}}_{k,p,l}^{(0)}, p = 1$ is a null vector.

4 Performance Results

In this section we present a set of performances results, analytical and numerical, for the proposed receiver structures. We consider 3 ($K = 3$) transmitter–receiver pairs, each equipped with 2, 4 and 6 antennas ($M = 2, 4, 6$). Thus, each transmitter can transmit simultaneously 1, 2 and 3 ($P = 1, 2, 3$) L-length data blocks (with $L = 128$), respectively. The FFT size is also set to 128 and a QPSK constellation under Gray mapping rule is considered.

The channel between each transmitterand receiver pair are uncorrelated and severely time-dispersive, each one with rich multipath propagation and uncorrelated Rayleigh fading for different multipath components. Specifically, we assume a $L_p = 32$-path frequency-selective block Rayleigh fading channel with uniform power delay profile (i.e., each path with average power of $1/L_p$). The same conclusions could be drawn for other multipath fading channels, provided that the number of separable multipath components is high. We also assume perfect channel state information and synchronization. Our performance results are presented in terms of the average bit error rate (BER) as a function of E_b/N_0, with E_b denoting the average bit energy and N_0 denoting the one-sided noise power spectral density. In all scenarios we present the theoretical and simulated average BER performance for the proposed receiver structures (IB-DFE PIC and IB-DFE SIC based approaches). For the sake of comparisons we also include the MFB performance.

Figure 4 shows the performance results for the $M = 2$. In this case only 1 block is transmitted simultaneously and thus both the IB-DFE PIC and IB-DFE SIC approaches are equivalents. We present results for 1, 2 and 4 iterations of the IB-DFE based equalizer. From this figure, it is clear that the proposed analytical approach is very precise for the first iteration, where the IB-DFE based equalizer reduces to a standard linear IA MMSE-based frequency domain equalizer (discussed in section III.b), since $\mathbf{\Psi}_k^{(0)}$ is a null matrix and $\bar{\mathbf{s}}_{k,l}^{(0)}$ is a null vector. Although there is a small difference between theoretical and simulated results for the subsequent iterations, our analytical approach is still very accurate, with differences of just a few tenths of dB. The difference is slightly higher for the second
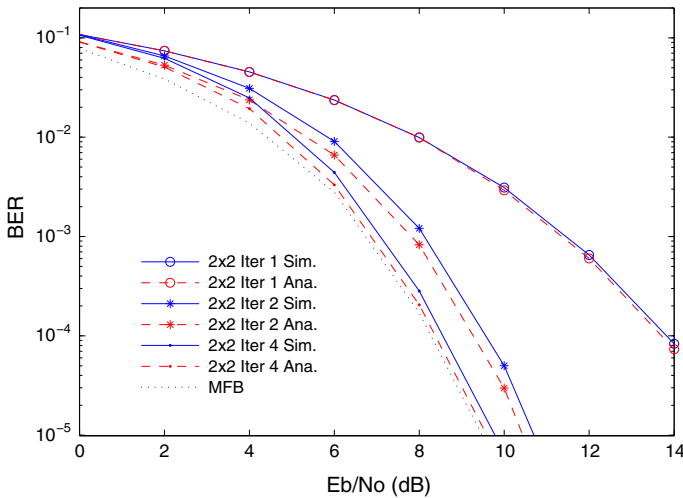
Fig. 4 Performance evaluation of the proposed receiver scheme for $M = 2$

iteration, decreasing as we increase the number of iterations. This behavior is a consequence of the accuracy of the Gaussian approximation that is behind our theoretical results. In fact, for a severely time-dispersive channel with rich multipath propagation the inter symbol interference (ISI) is high, which validates the Gaussian approximation of the residual interference after the first iteration. For the second iteration we reduce significantly the ISI, which makes the Gaussian approximation less accurate. However, as we increase the number of iterations we remove almost entirely the ISI (especially for large E_b/N_0 and, consequently, small BER) and we converge to a noise-only scenario where, once again, the Gaussian approximation is valid. Errors in the computation of the reliability of the estimates employed in the feedback loop (which are only accurate for the first iteration (zero reliability) and when we have very low BER (which means that the estimates are accurate, i.e., its reliability is 1)) also contribute to higher differences between theoretical and simulated results especially at the second iteration. As expected, the BER performance improves with the iterations and its can be observed that for the 4th iteration the performance is close the one obtained by the MFB.

Figures 5 and 6 shows the performance results for $M = 4$ with an IB-DFE PIC and an SIC, respectively. We also present results for 1, 2 and 4 iterations of the IB-DFE based equalizer. Clearly, the BER performance improves with the iterations for both IB-DFE SIC and PIC approaches, and it can be seen that for the 4th iteration the performance is close the one obtained by the MFB. Note that for this scenario the IB-DFE based equalizers must deal with inter-block interference, since 2 length-128 blocks are transmitted simultaneously by each transmitter, and residual inter-carrier interference. Therefore, the proposed schemes are quite efficient to separate the spatial streams and achieve the higher diversity order inherent to this scenario, with only a few iterations. Comparing the IB-DFE SIC and the IB-DFE PIC approach, it is clear that for the first and second iterations the IB-DFE SIC approach outperforms the IB-DFE PIC one. This is because the IB-DFE SIC based structure to detect a given user takes into account the previous detected ones, with the exception for the first user. It should be emphasized that, contrarily to the IB-DFE PIC approach, the first iteration of the IB-DFE SIC is not equivalent to the standard IA MMSE
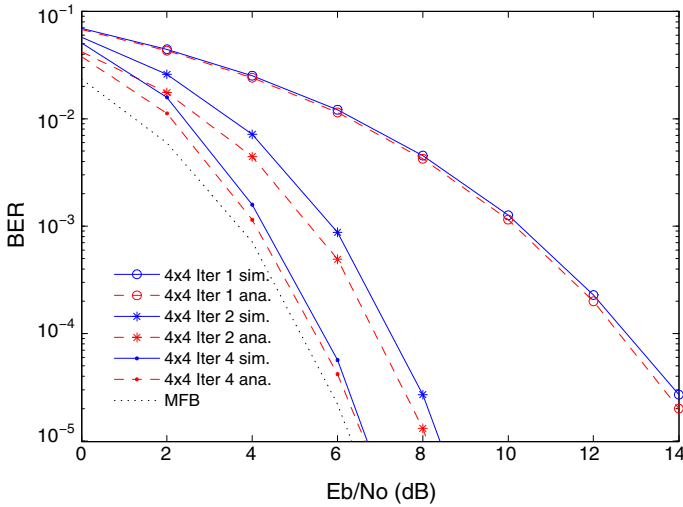
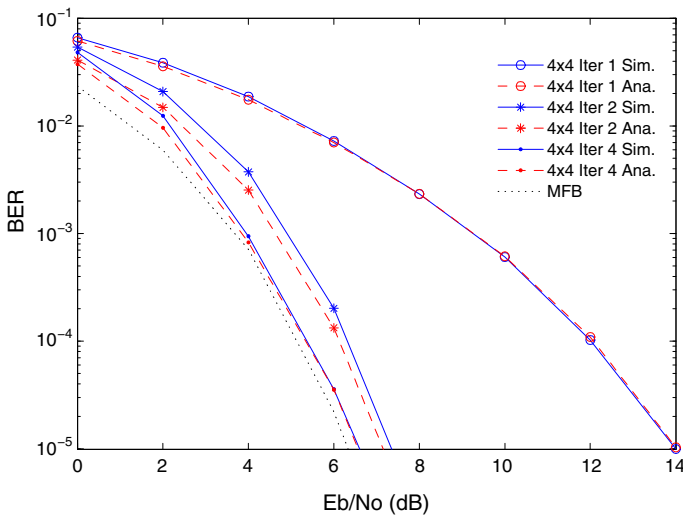**Fig. 5** Performance evaluation of the proposed IB-DFE PIC scheme for $M = 4$

Fig. 6 Performance evaluation of the IB-DFE SIC scheme for $M = 4$

equalizer. This is because to detect the second data block we took into account the information of the first detected data block, even for the first iteration.

Figures 7 and 8 shows the performance results for $M = 6$ with an IB-DFE PIC and an IB-DFE SIC, respectively. In this scenario 3 L-length data block are transmitted in parallel, increasing the inter-block interference. However, even in this case the performance of the proposed IB-DFE PIC and SIC approaches tends to the one achieved by the MFB as the number of iterations increases. As in the previous scenario, we can see that 4 equalizers iterations are enough to achieve the MFB performance.
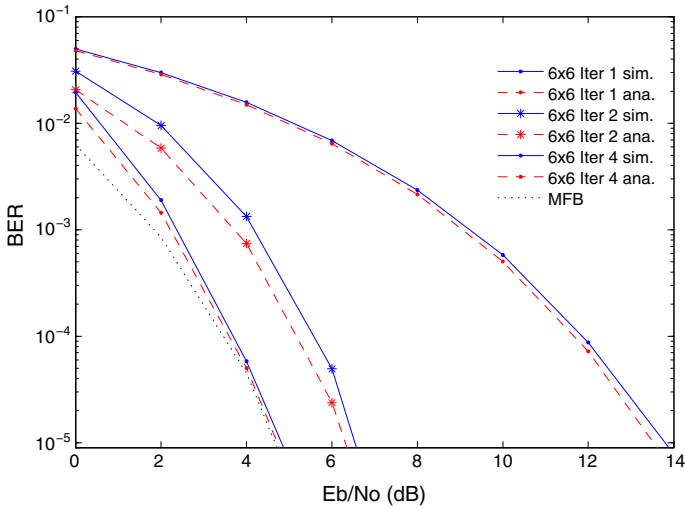
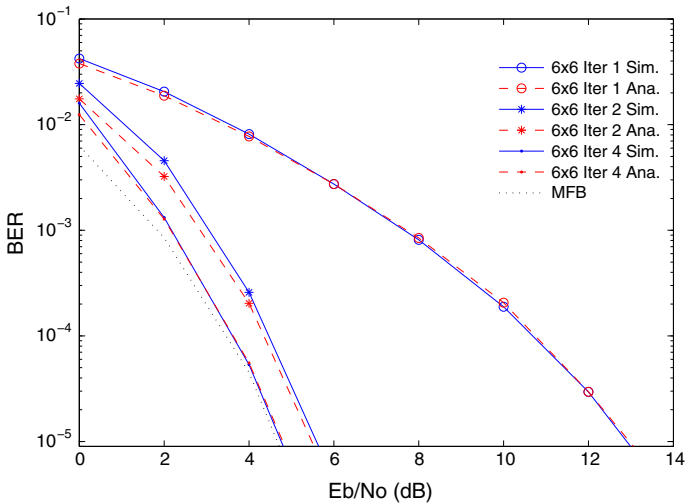**Fig. 7** Performance evaluation of the proposed IB-DFE PIC schemefor $M = 6$

Fig. 8 Performance evaluation of the IB-DFE SIC scheme for $M = 6$

5 Conclusions

In this paper we considered closed form IA precoding at the transmitter with IB-DFE based processing at the receiver for MIMO MC-CDMA systems. Two strategies were considered: IB-DFE SIC and PIC approaches. In the first one the equalizer matrices were obtained by minimizing the MSE of each individual data stream at each subcarrier; while in the second one the matrices were computed by minimizing the overall MSE of all data streams at each

subcarrier.We also proposed a simple, yet accurate analytical approach for obtaining the performance of our receiver structure.

The results have shown that the proposed receiver structures are robust to the residual inter-user interference (contrarily to the standard linear IA MMSE based equalizer) and quite efficient to separate the spatial streams, while allowing a close-to-optimum space-diversity gain, with performance close to the MFB with only a few receiver iterations. Also, the performance of both IB-DFE PIC and IB-DFE SIC receiver structures is basically the same after three or four equalizer iterations.To conclude we can clearly state that the proposed receiver structure are an excellent choice for the IA-precoded MC-CDMA based systems.

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