IB-DFE SIC BASED RECEIVER STRUCTURE FOR IA-PRECODED MC-CDMA SYSTEMS

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ABSTRACT

Interference alignment (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 (Iterative Block Decision Feedback Equalization) can efficiently exploit the inherent space-frequency diversity of the MIMO MC-CDMA systems. In this paper we design a joint iterative IA precoding at the transmitter with IB-DFE successive interference cancellation (SIC) based receiver structure for MC-CDMA systems. The receiver is designed in two steps: first a linear filter is used to mitigate the inter-user aligned interference, and then an iterative frequency-domain receiver is designed to efficiently separate the spatial streams in the presence of residual inter-user aligned interference at the output of the filter. Our scheme achieves the maximum degrees of freedom provided by the IA precoding, while allowing an almost optimum space-diversity gain, with performance close to the matched filter bound (MFB).

***Index Terms*—** interference alignment, interference channels, iterative block equalization, MC-CDMA systems.

1. INTRODUCTION

To achieve the 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 good air-interface candidate, especially for the downlink [1]. This scheme combines efficiently orthogonal frequency division Multiplex (OFDM) and code division multiple access (CDMA). It is well known that the user capacity of MC-CDMA system is essentially limited by interference. This interference can be mitigated by employing precoding techniques [2], iterative block decision feedback equalization (IB-DFE) based receivers [3].

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Conventional frequency domain equalization (FDE) schemes employ a linear FDE optimized under the minimum mean square error (MMSE) criterion. However, the residual interference levels might still be too high, leading to performance that is still several dB from the matched filter bound (MFB) [4]. Nonlinear time-domain equalizers are known to outperform linear equalizers and conventional, time-domain DFEs are known to have good performance-complexity tradeoffs [5]. For this reason, there has been significant interest in the design of nonlinear FDEs in general and frequency-domain FDEs in particular, with the IB-DFE being the most promising nonlinear FDE [4]. IB-DFE was originally proposed in [6] and was extended for a wide range of scenarios in the last 10 years, ranging from diversity scenarios [7], MIMO systems [8] and MC-CDMA systems [9], among many others. Essentially, the IB-DFE can be regarded as a low complexity turbo equalizer implemented in the frequency-domain that does not require the channel decoder output in the feedback loop, although true turbo equalizers based on the IB-DFE concept can also be designed [10].

One interesting scheme to efficiently eliminate the inter-user interference and achieve a linear capacity scaling is interference alignment (IA) [11]. With this strategy more interferers can be completely cancelled than with other interference cancellation methods, thus achieving the maximum degrees of freedom (DoF) [12]. An explicit formulation of the precoding vectors achieving IA for time or frequency selectivity channels has been presented in [11]. As closed-form solution for constant channels is still unknown except with 3 users, iterative algorithms [14][15] have been proposed for an arbitrary number of transmitter-receiver pairs. An MMSE-based iterative IA scheme was proposed in [14]. Several iterative linear precoding designs using alternating minimization were proposed in [15].

In this paper we consider MIMO MC-CDMA systems with iterative IA precoding at transmitter and iterative frequency-domain receivers based on the IB-DFE concept. The main motivation to consider this combination, is that IA based techniques achieve the maximum DoF in MIMO interference channels. However, they cannot by themselves efficiently exploit the space-frequency diversity inherent of the MIMO MC-CDMA systems. On the other hand, IB-DFE based receivers are well known to be one of the most efficient techniques to exploit this space-frequency diversity. Therefore, this combination allows us to design a system that is able to achieve maximum DoF (number of spatial stream per sub-carrier) and exploit the high diversity order inherent to these systems. The proposed receiver structure is designed in two steps: first a linear filter is used to mitigate the inter-user aligned interference, and then an IB-DFE SIC based equalizer is employed to efficiently separate the spatial streams. In the design of the equalizer, we explicitly take into account the residual inter-user interference, allowing not only an efficient separation of the spatial streams but also a reduction the number of iteration in the IA procedure, and thus also reducing the information needed to be exchanged between the different transmitter-receiver pairs. The results are compared with the IB-DFE parallel interference cancellation (PIC) approach, recently proposed for IA-precoded MC-CDMA systems [16].

The remainder of the paper is organized as follows: section II presents the MIMO IA-precoded MC-CDMA system model. Section III, starts by briefly presenting the iterative IA-precoding considered in this paper. Then, the proposed SIC based receiver structure is presented in detail. Section IV presents the performance results. The conclusions will be drawn in section V.

Notation: Throughout this paper, we will use the following notations. tr(**A**) is the trace of matrix **A**, $\nu_{\min}^{P}(\mathbf{A})$ is the matrix whose columns are the eigenvectors corresponding to the P smallest eigenvalues of matrix **A** and $\mathbf{e}_p$ is an appropriate column vector with 0 in all positions except the pth position that is 1.

2. SYSTEM MODEL

We consider a K-user MIMO interference channel with constant coefficients on a per-subcarrier basis. It comprises K transmitter-receiver pairs sharing the same physical channel, where a given transmitter only intends to have its P_k spatial data symbols on each subcarrier decoded by a single receiver. Without loss of generality, we consider a symmetric case where all transmitters and receivers have M antennas, and $P_k = P\ \forall k$, which is denoted by an (M,M,K) interference channel with P data symbols per-subcarrier. For each one of the P L-length data symbols blocks, $\{d_{k,p,l}; k=1,...K,\ p=1,...,P,\ l=0,...,L-1\}$, where the constellation symbol $d_{k,p,l}$ (with $\mathbb{E}\left[\left|d_{k,p,l}\right|^2\right]=\sigma_d^2$) is selected from the data according to given mapping rule, is spread into L chips using orthogonal Walsh-Hadamard codes, leading to the block $\{s_{k,p,l}; p=1,...,P,\ l=0,...,L-1\}$. Then, a set of P chips, one of each block, is weighted by an IA-precoding matrix [16]. 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 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 $\left\|\mathbf{W}_{k,l}\right\|_F^2 \le T_p$ and T_p is the transmit power at the transmitters, with $\mathbf{s}_{k,l} = \begin{bmatrix} s_{k,1,l} & \cdots & s_{k,P,l}\end{bmatrix}^T$.Finally, the precoding signals are mapped into the OFDM symbol and the 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

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

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, $\mathbf{n}_{k,l}$ is the additive white Gaussian noise (AWGN) vector at receiver k on subcarrier l, i.e., $\mathbf{n}_{k,l} \sim \mathcal{CN}\left(0,\sigma_n^2\mathbf{I}_M\right)$.

The detection procedure is done in two steps: first a linear filter is used to mitigate the aligned user's interference. Thus, the signal after the filtering process is given by

$$\bar{\mathbf{y}}_{k,l} = \mathbf{\Phi}_{k,l}^H\mathbf{H}_{k,k,l}\mathbf{W}_{k,l}\mathbf{s}_{k,l} + \mathbf{\Phi}_{k,l}^H\sum_{\substack{j=1\\ j\neq k}}^{K} \mathbf{H}_{k,j,l}\mathbf{W}_{j,l}\mathbf{s}_{j,l} + \mathbf{\Phi}_{k,l}^H\mathbf{n}_{k,l} \tag{3}$$

where $\mathbf{\Phi}_{k,l} \in \mathbb{C}^{MxP}$ denotes the first linear receiving filter. Second, a non-linear equalizer based on IB-DFE principle is designed to efficiently separate the spatial P L-length data block.

3. ITERATIVE EQUALIZER DESIGN

In this section we start by briefly reviewing the iterative minimum interference leakage (IL) IA precoding algorithm [15]. Then, the proposed non-linear iterative space-frequency equalizer is presented in detail.

A) IA Precoders

In constant MIMO interference channel, closed-form solutions have been found in only a few specific cases [15] (e.g. $K \le 3$), thus we consider an iterative approach for a general case. The aim is to precode the signal at transmitter j in such a way that the interference caused by that transmitter in receiver k, with $k \neq j$, is nearly orthogonal to a subspace of its receive space. This subspace, with orthonormal basis $\mathbf{\Phi}_{k,l}$, and precoders are jointly designed to optimize an appropriate cost function. The iterative interference leakage algorithm minimizes the total IL that remains at each receiver after attempting to cancel the aligned interference by left multiplication with $\mathbf{\Phi}_{k,l}^H$ for each user k (see (3)). The global function to optimize is

$$\Im_{IL} = \sum_{k=1}^{K} \sum_{\substack{j=1 \\ j \neq k}}^{K} \left\| \mathbf{\Phi}_{k,l}^H \mathbf{H}_{k,j,l} \mathbf{W}_{j,l} \right\|_F^2 \tag{4}$$

which is usually denoted by IL. The optimization problem can be formulated as

$$\min \Im_{IL}\left(\left\{\mathbf{W}_{j,l}\right\},\left\{\mathbf{\Phi}\right\}\right) \text{ s.t.} \begin{cases} \mathbf{W}_{j,l}^H \mathbf{W}_{j,l} = T_P \mathbf{I}_P, j \in \{1,\ldots,K\} \\ \mathbf{\Phi}_{k,l}^H \mathbf{\Phi}_{k,l} = \mathbf{I}_P, k \in \{1,\ldots,K\}. \end{cases} \tag{5}$$

A simple approach to solve this problem is to use an alternating minimization procedure [15]. This algorithm takes the following iterative form: Define an arbitrary orthogonal basis $\mathbf{\Phi}_{k,l}$ for each receiver subspace on each subcarrier. Find the precoder matrix $\mathbf{W}_{j,l}$ such that each node has maximum squared Euclidean distance between it and the subspace spanned by the columns of each $\mathbf{\Phi}_{k,l}$, by using

$$\mathbf{W}_{j,l} = v_{\min}^{P}\left(\sum_{\substack{k=1 \\ k \neq j}}^{K} \mathbf{H}_{k,j,l}^H \mathbf{\Phi}_{k,l} \mathbf{\Phi}_{k,l}^H \mathbf{H}_{k,j,l} \right) \tag{6}$$

Update the receiver orthonormal subspaces according to

$$\mathbf{\Phi}_{k,l} = v_{\min}^{P}\left(\sum_{\substack{j=1 \\ j \neq k}}^{K} \mathbf{H}_{k,j,l} \mathbf{W}_{j,l} \mathbf{W}_{j,l}^H \mathbf{H}_{k,j,l}^H \right) \tag{7}$$

Repeat (6) and (7) until convergence.

The subspace $\mathbf{\Phi}_{k,l}$ is reserved for each k's signal (i.e., for each user) on each subcarrier, thus the IA at receiver k on subcarrier l is ideally orthogonal to this subspace. Then, each receiver must still separate the desired spatial streams after the interference aligned has been mitigated with left multiplication of $\mathbf{\Phi}_{k,l}^H$. A standard MMSE linear equalizer can be employed for this purpose. Therefore, in the conventional receiver a linear equalizer $\mathbf{G}_{k,l}$ is formed by multiplying $\mathbf{\Phi}_{k,l}^H$ and the linear spatial equalizer $\overline{\mathbf{G}}_{k,l}$, which neglects the inter-user interference aligned and equalizes only the desired signal, so that $\mathbf{G}_{k,l} = \overline{\mathbf{G}}_{k,l} \mathbf{\Phi}_{k,l}^H$. Then the vector $\tilde{\mathbf{s}}_{k,l} = \mathbf{G}_{k,l} \overline{\mathbf{y}}_{k,l}$ is the estimate of the original transmit vector $\mathbf{s}_{k,l}$.

B) Iterative Equalizer Design

It is well known that for MC-CDMA based systems, linear equalization is not efficient to separate the spatial streams due to the residual inter-carrier interference (ICI). In the context of IA based systems, it also neglects the residual inter-user aligned interference. Therefore, a conventional equalizer design may not be the best strategy. As mentioned, the subspace $\mathbf{\Phi}_{k,l}$ should be ideally orthogonal to interference aligned subspace, but full orthogonality may be not possible for some practical scenarios [17] and/or requires a significant number of iterations [15]. Thus, we design a new non-linear receiver structure based on IB-DFE SIC based principles which also takes into account the residual inter-user interference so that perfect alignment constraint can be relaxed. This receiver is quite efficient to separate the P L-length data symbol blocks, even for the case where the system suffers from residual inter-user interference, and thus the full orthogonality requirement can be relaxed.

Fig. 1 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. 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

$$\tilde{s}_{k,p,l}^{(i)} = \mathbf{f}_{k,p,l}^{(i)T} \overline{\mathbf{y}}_{k,l} - \mathbf{b}_{k,p,l}^{(i)T} \overline{\mathbf{s}}_{k,p,l}^{(i-1)}, \tag{8}$$

where $\mathbf{f}_{k,p,l}^{(i)} \in \mathbb{C}^{Px1}$ and $\mathbf{b}_{k,p,l}^{(i)} \in \mathbb{C}^{Px1}$ denoting the feedforward and feedback vectors coefficients of the pth data block applied on the lth subcarrier at the kth user, and $\overline{\mathbf{s}}_{k,p,l}^{(i-1)} = \left[\overline{s}_{k,1,l}^{(i)},\ldots,\overline{s}_{k,p-1,l}^{(i)},\overline{s}_{k,p,l}^{(i-1)},\ldots,\overline{s}_{k,P,l}^{(i-1)} \right]^T$.

Setting $\mathbf{H}_{k,j,l}^{eq} = \mathbf{\Phi}_{k,l}^H \mathbf{H}_{k,j,l} \mathbf{W}_{j,l}$, (3) can be rewritten as

$$\overline{\mathbf{y}}_{k,l} = \underbrace{\mathbf{H}_{k,k,l}^{eq} \mathbf{s}_{k,l}}_{\text{desired signal}} + \underbrace{\sum_{\substack{j=1 \\ j \neq k}}^{K} \mathbf{H}_{k,j,l}^{eq} \mathbf{s}_{j,l}}_{\text{res. inter-user aligned interf.}} + \underbrace{\mathbf{\Phi}_{k,l}^H \mathbf{n}_{k,l}}_{\text{noise}}, \tag{9}$$

where $\mathbf{H}_{k,j,l}^{eq} \in \mathbb{C}^{PxP}$ represents the equivalent channels after the IA procedure. The block $\left\{ \overline{s}_{k,p,l}^{(i-1)}; l = 0,\ldots,L-1 \right\}$ is the spreading of the pth de-spreading block average values

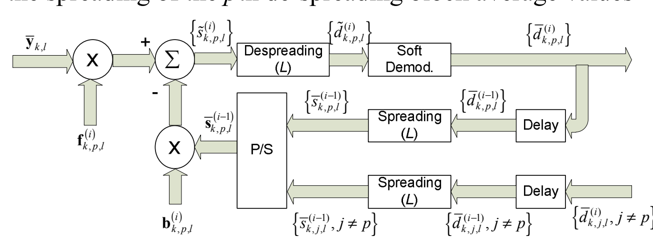

Fig. 2: Details of the IB-DFE SIC block.

conditioned to the detector output $\left\{\overline{d}_{k,p,l}^{(i)}; l=0,\ldots,L-1\right\}$ for each iteration i.

It can be shown that $\overline{\mathbf{s}}_{k,l}^{(i-1)} \approx \mathbf{\Psi}_k^{(i-1)^2}\mathbf{s}_{k,l} + \mathbf{\Psi}_k^{(i-1)}\mathbf{\Delta}_{k,l}$ [10], where the error $\mathbf{\Delta}_{k,l} = \left[\Delta_{k,1,l} \ \cdots \ \Delta_{k,P,l}\right]^T$ has zero mean and uncorrelated with $\mathbf{\Psi}_k^{(i-1)}$, $\mathbf{\Psi}_k^{(i)} = diag\left(\psi_{k,1}^{(i)},\ldots,\psi_{k,P}^{(i)}\right)$ with correlation coefficients defined as

$$\psi_{k,p}^{(i)} = \frac{\mathbb{E}\left[\hat{d}_{k,p,l}^{(i)} d_{k,p,l}^*\right]}{\mathbb{E}\left[\left|d_{k,p,l}\right|^2\right]}, \qquad (10)$$

being a measure of the pth L-length block estimates reliability associated to the ith iteration, that can be obtained as described in [9] for QSPK modulations and in [18] for larger constellations. Considering QPSK modulations, the hard decision associated to the symbol $d_{k,p,l}$ is $\hat{d}_{k,p,l}^{(i)} = sign\left(\mathrm{Re}\left\{\tilde{d}_{k,p,l}^{(i)}\right\}\right) + jsign\left(\mathrm{Im}\left\{\tilde{d}_{k,p,l}^{(i)}\right\}\right)$.

For a given iteration and at each receiver, the iterative non-linear equalizer is characterized by the coefficients $\mathbf{f}_{k,p,l}^{(i)}$ and $\mathbf{b}_{k,p,l}^{(i)}$. These vectores are computed by minimizing the MSE of the spreading samples associated to the pth L-length data block on lth subcarrier $\mathrm{MSE}_{k,p,l}^{(i)} = \mathbb{E}\left[\left|\tilde{s}_{k,p,l}^{(i)} - s_{k,p,l}\right|^2\right]$. The optimization problem for this approach can be formulated as

$$\min_{\mathbf{f}_{k,p,l},\mathbf{b}_{k,p,l}} \mathrm{MSE}_{k,p,l}^{(i)} \ \text{s.t} \ \frac{1}{L}\sum_{l=0}^{L-1}\mathbf{f}_{k,p,l}^{(i)T}\mathbf{h}_{k,k,p,l}^{eqT} = 1 \qquad (11)$$

where $\mathbf{h}_{k,k,p,l}^{eqT} \in \mathbb{C}^{P\mathrm{x}1}$ is the equivalent channel associated to the pth data block. After some mathematical manipulations, the feedforward and feedback vectors can be obtained as

$$\mathbf{f}_{k,p,l}^{(i)} = \begin{pmatrix} \mathbf{H}_{k,k,l}^{eqH}\left(\mathbf{I}_P - \mathbf{\Psi}_k^{(i-1)^2}\right)\mathbf{H}_{k,k,l}^{eq} \\ + \sum_{j=1,j\neq k}^{K}\mathbf{H}_{k,j,l}^{eqH}\mathbf{R}_d\mathbf{H}_{k,j,l}^{eq} + \frac{\sigma_n^2}{\sigma_d^2}\mathbf{I}_P \end{pmatrix}^{-1} \mathbf{H}_{k,k,l}^{eqH}\mathbf{\Omega}_{k,p}^{(i)}, \qquad (12)$$

and

$$\mathbf{b}_{k,p,l}^{(i)} = \mathbf{H}_{k,k,l}^{eq}\mathbf{f}_{k,p,l}^{(i)} - \mathbf{e}_p, \qquad (13)$$

with

$$\mathbf{\Omega}_{k,p}^{(i)} = \left(\mathbf{I}_P - \mathbf{\Psi}_k^{(i-1)^2}\right)\mathbf{e}_p - \frac{\mu_{k,p}^{(i)}}{\sigma_d^2 L}\mathbf{e}_p. \qquad (14)$$

where $\mathbf{R}_d = \sigma_d^2\mathbf{I}_P$. The Lagrangian multiplier is selected, at each iteration i, to ensure that the constraint $\frac{1}{L}\sum_{l=0}^{L-1}\mathbf{f}_{k,p,l}^{(i)T}\mathbf{h}_{k,k,p,l}^{eqT} = 1$ 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 $\overline{\mathbf{s}}_{k,p,l}^{(0)}$, p = 1 is a null vector.

4. PERFORMANCE RESULTS

In this section we present a set of performances results for the proposed receiver structure (IA equalizer with IB-DFE based equalizer). We consider 3 (K=3) transmitter-receiver pairs, each equipped with 4 (M=4) and transmitting simultaneously 2 (P=2) L-length data block, referred as (4, 4, 3) interference channel with P=2.The FFT size is set to 128 and a QPSK constellation under Gray mapping rule is considered. We assume a Lp = 32-path frequency-selective block Rayleigh fading channel with uniform power delay profile (i.e., each path with average power of $1/Lp$). The same conclusions could be drawn for other multipath fading channels, provided that the number of separable multipath components is high. Also, we 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. For the sake of comparisons we include the matched filter bound (MFB). For Fig. 3, the number of iteration for the IA procedure was set to 100, this number was found enough to have free inter-user aligned interference.

We present results for 1, 2 and 4 iterations of the IB-DFE based equalizer. Also, the results are compared with the IB-DFE PIC approach recently proposed in [16].

Starting by analysing the results presented in Fig. 2, as expected, the BER performance improves with the iterations for both IB-DFE SIC and PIC approaches, and it can be observed that for the 4th iteration the performance is close the one obtained by the MF. Therefore, the proposed scheme is quite efficient to separate the spatial streams and achieve the high diversity order inherent to this scenario, wi-

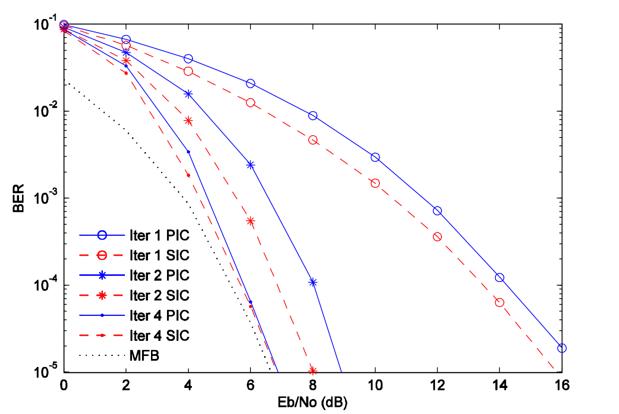

Fig. 2: Performance of the proposed receiver structure for (5, 5, 4) with P=2 and 100 IA iterations.

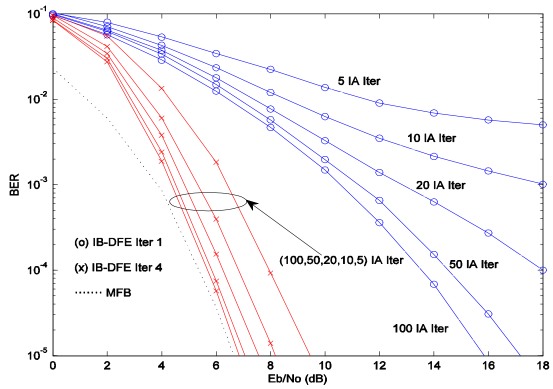

Fig. 3: Performance of the proposed receiver structure for (5, 5, 4) with P=2 and several IA iterations.

th only a few iterations. Comparing the SIC and the PIC approach, it is clear that for the first and second iterations the SIC approach outperforms the PIC one. This is because the SIC- based structure to detect a given user takes into account the previous detected ones, with the exception for the first user. However, when the number of iteration increases, the performance of the PIC approach tends to the one given by the SIC approach. We can observe that the BER performance of both approaches is basically the same for four 4 iterations.

Fig. 3 presents results for (5, 5, 4) with P=2, 1 and 4 iterations for the IB-DFE based equalizer and several iterations for the IA procedure, 5, 10, 20, 50 and 100. When we have a single equalizer iteration there is a significant performance degradation from 100 to 5 IA iterations, since the residual inter-user interference increases as the number of IA iterations decreases. However, for the fourth equalizer iteration we can see that the performance degradation from 100 to 5 IA iterations is much lower. In fact, we have almost the same performance with 100 and 10 to 20 IA iterations, which means we can reduce significantly the number of IA iterations. Thus also reducing the information needed to be exchanged between the different transmitter-receiver pairs.

5. CONCLUSIONS

In this paper we considered IA precoding, on a per-subcarrier basis, at the transmitter with IB-DFE SIC based processing at the receiver for MIMO MC-CDMA systems. The proposed receiver structure was designed in two steps: first a linear filter was considered to mitigate the inter-user aligned interference, and then an iterative frequency-domain SIC based equalizer was designed to efficiently separate the spatial streams.

The results have shown that the proposed receiver structure is robust to the residual inter-user aligned interference 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 iterations of the equalizer. The performance of both IB-DFE PIC and IB-DFE SIC receiver structures is basically the same after three or four equalizer iterations. The proposed robust residual inter-user aligned interference allows significant reduction in the number of iterations of the IA process, and thus reducing the information needed to be exchanged between the different transmitter-receiver pairs.

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