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Performance analysis of closed-loop pre-equalisation for multiuser multiple-input multiple-output with multicarrier code division multiple access systems

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Abstract: The use of multiple transmit and receive antennas is widely recognised as an effective technology to boost the capacity of wireless communication systems. Moreover, the combination of multiple-input multiple-output (MIMO) systems with multicarrier code division multiple access (MC-CDMA) offers a strong alternative to satisfy the demand for high data rates with rigorous quality-of-service (QoS) restrictions. In this study, this paper applies a closed-loop pre-equalisation methodology under a unified framework for MIMO and MC-CDMA systems that satisfies the QoS target with a single-user-based detector while minimising the power of the pre-equalisation factors. It is of particular interest to investigate the impact and limitations of combining the robustness of the feedback scheme with the degrees of freedom available in the system, given in terms of the number of subcarriers and multiple antennas. The contribution of this work includes the derivation of the distributed and centralised optimal closed-loop pre-equalisation solutions under the MIMO–MC-CDMA structure. The results and analysis illustrate important gains in the form of power savings, enabled by the spatial diversity of the MIMO scheme.

1 Introduction

A multicarrier code division multiple access (MC-CDMA) is an important wireless technology that combines two communication strategies: orthogonal frequency division multiplexing and code division multiple access (CDMA) [1–6]. Since the performance of MC-CDMA is affected by the multiple-access interference (MAI) in the communication channels, there have been important contributions attempting to overcome this limitation, such as multi-user detection at the receiver side [7, 8] or power control [9]. However, in downlink transmissions [1], where the complexity of the detection scheme is a critical issue, multi-user detection is unattractive since it requires knowledge of the channel and also the signalling waveforms of all users, which may be impractical. On the other hand, power control has been extensively investigated for CDMA technology [10–15], where the main limitations come from time-varying channels, measurement and processing delays, measurement uncertainty and noise effects [16, 17]. This strategy was also introduced for MC-DS-CDMA systems, but from a game-theory perspective in [9].

A related problem to power control for quality-of-service (QoS) assignation is pre-equalisation. In fact, pre-equalisation is a more general problem, compared to power control, since it adapts both magnitude and phase of the transmitted symbols and not just the magnitude. Downlink pre-equalisation is an attractive strategy in a multiple-access environment that simplifies the receiver complexity by transferring the signal processing to the base station (BS). Some of the first research works to address the pre-equalisation problem for MC-CDMA networks can be found in [18–20]. In these works, they were looking to maximise the signal-to-interference-plus-noise ratio (SINR) subject to energy constraints at the transmitter or to eliminate completely the MAI at the receiver side. Other research efforts in this topic include [21–23]. In [21], a different methodology is adopted, where a grouping strategy is proposed for MC-CDMA systems and space-time block coding MC-CDMA, based on the minimisation of the total transmission power under a target SINR restriction. Recently, in [22], pre- and post-equalisations were applied to MC-CDMA indoor optical wireless communications to improve the bit error rate performance. Finally, an iterative closed-loop pre-equalisation was first proposed for downlink and uplink transmissions in [23].

Multiple-input multiple-output (MIMO) systems have become a key enabling technology for wireless communications that provide significant advantages by exploiting spatial diversity [24]. There has been extensive research on different aspects of MIMO systems, theoretical as well as experimental for the last 20 years or so. Traditionally, the concept of MIMO presupposed spatial multiplexing or coding diversity, but today it may indicate more degrees of freedom that can be exploited in different forms such as multiuser diversity, interference reduction, spatial modulation and hybrid approaches. Thus, different to these previous MIMO schemes, in this paper, the spatial diversity provided by the MIMO scheme is used to increase the degrees of freedom in the MC-CDMA system.

Therefore, and following our previous works in [23, 25], this paper applies the same methodology to derive and analyse a closed-loop pre-equalisation solution in a unified framework for MC-CDMA and MIMO systems under a multiuser scenario. While the feedback pre-equalisation structure of the analysed multiuser MIMO–MC-CDMA system can be applied for both downlink and uplink, our study focuses only on downlink pre-equalisation. As a first stage, the resulting SINR is analytically derived by assuming a linear detection strategy at the receiver. The aim of the downlink pre-equalisation concept presented here is to adjust the transmission power and phase of the data symbols on each subcarrier per antenna within an iterative closed-loop scheme, so that, interference and channel variations are compensated. In this fashion, there is no need of pilot symbols and equalisation for detection, allowing the use of a single user detector [matched filter (MF)] at each mobile user (MU). Our evaluations demonstrate that closed-loop pre-equalisation for downlink MIMO–MC-CDMA creates additional degrees of freedom that can be exploited to provide transmit diversity in the form of power savings.

The rest of the paper is organised as follows. First, the mathematical model for the received SINR in the downlink transmissions of a multiuser MIMO–MC-CDMA system is derived in Section 2. Then, Section 3 introduces the optimal solution of the closed-loop pre-equalisation scheme for the MIMO and MC-CDMA unified system using both distributed and centralised strategies. Section 4 describes an iterative algorithm to implement the optimal solution given in Section 3. The overall analysis of the closed-loop pre-equalisation scheme is presented in Section 5 using numerical evaluations. The performance analysis is focused on the QoS requirements and power consumption for different system configurations. Finally, the conclusions are given in Section 6.

The notation used in this paper is described as follows. $\mathbb{R}$ and $\mathbb{C}$ denote real and complex numbers, respectively; and $\mathbb{R}^N$ and $\mathbb{C}^N$ represent real and complex N-dimensional vectors, respectively. Scalars are represented by lower-case italic letters, and vector and matrices by boldface letters. $(\cdot)^\top$ and $(\cdot)^*$ describe the transpose and complex conjugate-transpose, respectively. $\boldsymbol{I}$ and $\mathbf{1}$ denote the identity matrix and the vector with unit entries, respectively, and $\text{diag}(\boldsymbol{X}_1, \ldots, \boldsymbol{X}_n)$ represents a block diagonal matrix with diagonal terms $\boldsymbol{X}_i$. For complex vectors $\boldsymbol{x} = [x_1, \ldots, x_N]^\top$ and $\boldsymbol{y} = [y_1, \ldots, y_N]^\top$, and the inner product is defined as $\langle \boldsymbol{x}, \boldsymbol{y}\rangle = \boldsymbol{y}^*\boldsymbol{x}$. The operator $\circ$ indicates the Hadamard product. Finally, $\mathcal{E}\{\cdot\}$ is used to denote the expectation operator.

2 Multiuser MIMO–MC-CDMA system model

We consider the downlink of a MIMO–MC-CDMA wireless system with U users and N_s subcarriers, the system employs N_T transmit antennas at the BS and N_{R_j} receive antennas on each MU $\forall\, j \in [1, U]$. Notice that we use the subscript j in N_{R_j} to allow the general case when the number of receiving antennas on each MU receiver is different. Fig. 1 illustrates in a block diagram the system model under analysis. After modulating the data bits, the data symbol of user j at k-time instant, $b_j[k] \in \mathbb{C}$ is spread using a Walsh–Hadamard sequence $\boldsymbol{c}_j = [c_{j,1} \ldots c_{j,N}]^\top \in \mathbb{R}^N$ where $c_{j,i} \in \{-1/\sqrt{N}, 1/\sqrt{N}\}$ $\forall i = 1, \ldots, N$, given N as the processing gain. The N chips of the spread data symbol are simultaneously compensated by N pre-equalisation factors, per transmit antenna, before modulating a set of N_s orthogonal subcarriers using the inverse fast Fourier transform (IFFT) [1]. After IFFT and before signal transmission, a serial to parallel (S/P) converter is applied and a guard band (cyclic prefix) of length at least equal to the channel delay spread is added to prevent inter-symbol interference. It is assumed that the number of subcarriers must be a multiple of the processing gain, i.e. $N_\text{s} = \alpha N$, where α is a positive integer. For simplicity and without loss of generality, α is set to 1 for the rest of this paper.

At the jth MU receiver, after cyclic prefix removal and serial to parallel conversion, the output signal of the fast Fourier transform at instant k, $\boldsymbol{y}_r^j[k]$, and receive antenna r can be described as

$$\boldsymbol{y}_r^j[k] = \sum_{t=1}^{N_\text{T}} \sum_{l=1}^{U} \left(\boldsymbol{h}_j^{r,t}[k] \circ \boldsymbol{c}_l \circ \boldsymbol{p}_l^t[k]\right) b_l[k] + \boldsymbol{\eta}_r[k] \tag{1}$$

where $\boldsymbol{p}_l^t[k] = \left[p_{l,1}^t[k] \ \ldots \ p_{l,N}^t[k]\right]^\top \in \mathbb{C}^N$ denotes the vector of complex pre-equalisation factors (power and phase compensations) for the lth user transmitting from the tth antenna through all N subcarriers, $\forall l \in [1, U]$ and $t \in [1, N_\text{T}]$, and vector $\boldsymbol{h}_j^{r,t}[k] = \left[h_{j,1}^{r,t}[k], \ldots, h_{j,N}^{r,t}[k]\right]^\top \in \mathbb{C}^N$ contains the channel gains linking the tth transmit antenna and the rth receive antenna of the jth MU over the N subcarriers $\forall r \in [1, N_{\text{R}_j}]$. Moreover, it is assumed a frequency selective Rayleigh fading profile for each user channel and for practical implications, the magnitudes of the pre-equalisation factors are restricted to a feasible interval at any time instant k, i.e. $p_{\min} \le |p_{l,i}^t[k]|^2 \le p_{\max}$, where $0 < p_{\min} < p_{\max}$. Finally, $\boldsymbol{\eta}_r[k] \in \mathbb{C}^N$ symbolises a zero-mean complex Gaussian noise vector with covariance matrix $\mathcal{E}\{\boldsymbol{\eta}_r[k]^*\boldsymbol{\eta}_r[k]\} = \sigma^2\boldsymbol{I}$ for any time instant k, where $0 < \sigma < \infty$.

In order to formulate and examine the performance of the proposed closed-loop pre-equalisation method under the unifying MIMO–MC-CDMA framework, we consider the application, in general, of a linear detector to retrieve the data symbols on each antenna at the MU receiver [7]. The coefficient vector for the rth detector is defined as $\boldsymbol{\chi}_j^r = [\chi_{j,1}^r \ldots \chi_{j,N}^r]^\text{T} \in \mathbb{C}^N$. The general

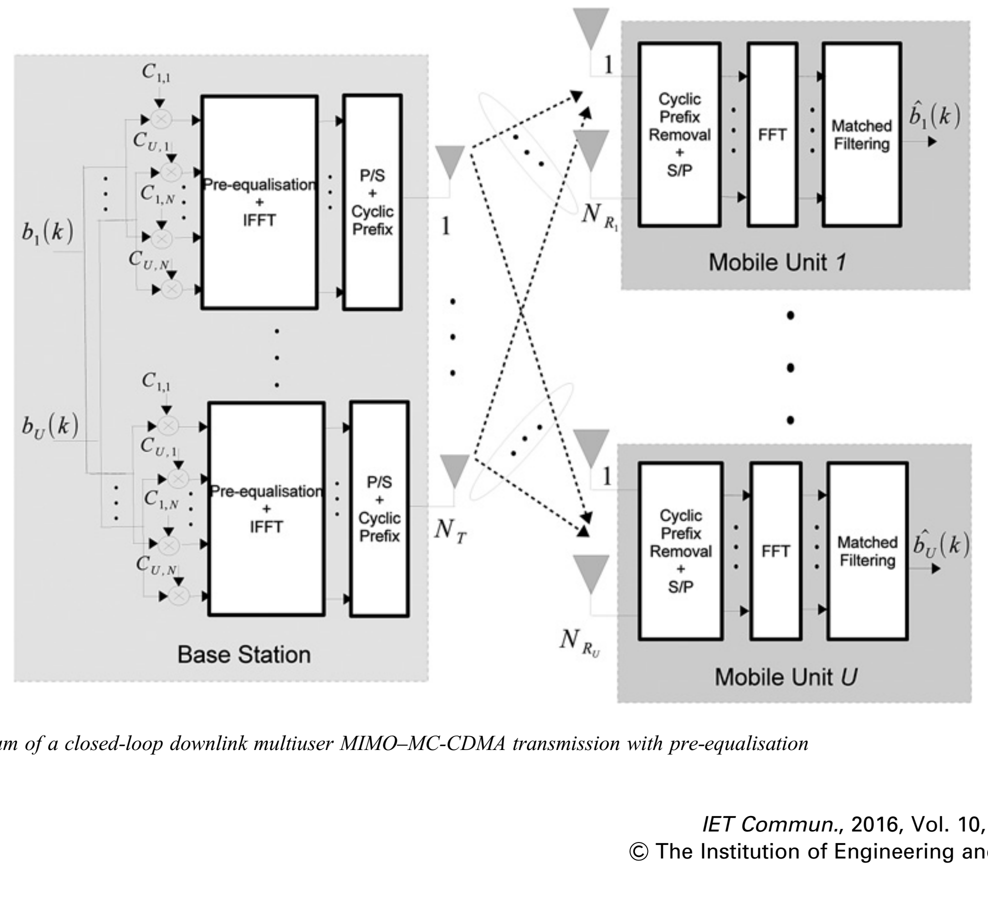

Fig. 1 *Block diagram of a closed-loop downlink multiuser MIMO–MC-CDMA transmission with pre-equalisation*

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structure for the detection process of the jth MU data symbol is then formed by a bank of linear detectors that can be expressed by

$$\hat{b}_j[k] = \sum_{r=1}^{N_{\mathrm{R}_j}} (\boldsymbol{\chi}_j^r)^* \boldsymbol{y}_r^j[k], \tag{2}$$

where the decision statistic $\hat{b}_j[k]$ for the data symbol at time k of user j is obtained by adding together the N_{R_j} detectors' outputs, one for each receive antenna. Therefore, substituting the received signal model in (1) into (2), we obtain the following representation for the decision statistic of the kth data symbol of user j

$$\begin{aligned}\hat{b}_j[k] = {} & \underbrace{\sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{t=1}^{N_{\mathrm{T}}} \sum_{i=1}^{N} (\chi_{j,i}^r)^* h_{j,i}^{r,t}[k] c_{j,i} p_{j,i}^t[k] b_j[k]}_{j\text{th user information}} \\ & + \underbrace{\sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{t=1}^{N_{\mathrm{T}}} \sum_{\substack{l=1,\\ l\neq j}}^{U} \sum_{i=1}^{N} (\chi_{j,i}^r)^* h_{j,i}^{r,t}[k] c_{l,i} p_{l,i}^t[k] b_l[k]}_{\text{Interference}} \\ & + \underbrace{\sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{i=1}^{N} (\chi_{j,i}^r)^* \eta_{r,j}[k].}_{\text{Noise component}}\end{aligned} \tag{3}$$

Note that the above equation reflects the contribution of the pre-equalisation factors on the kth data symbol at the receiver side, where a triple summation expression is obtained. These terms come come from all N subcarriers, and each of the transmit and receive antennas used in the link. In the same manner, the pre-equalisation factors have an impact in the multiple interference term that may be added up from other users' transmissions. Observe that the jth mobile terminal sees a block of $N \cdot N_{\mathrm{T}}$ interference terms on each receive antenna from each undesired user.

Let us now analyse the estimate of the data symbol energy at the receiver output using (3), (see (4)). It has been assumed in (4) that the data symbols have zero mean and normalised energy, i.e. $\mathcal{E}\{b_j[k]\} = 0$ and $\mathcal{E}\{|b_j[k]|^2\} = 1$. Moreover, it is also assumed that the data symbols among active users and noise samples are uncorrelated. To highlight the joint presence of the data symbol, interference and noise energy at the receiver, $\mathcal{E}\{|\hat{b}_j[k]|^2\}$ in (4) can be rewritten in vector notation as (see equation at bottom of the page)

where $\boldsymbol{h}_{j,j}^{r,t}[k] = \boldsymbol{\chi}_j^r \circ (\boldsymbol{h}_j^{r,t}[k])^* \circ \boldsymbol{c}_l \in \mathbb{C}^N$ is the vector of correlated channel gains per receive antenna. Finally, the measured SINR for the jth user at instant k is given by the following expression:

$$\begin{aligned}\gamma_j[k] &= \frac{\sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{t=1}^{N_{\mathrm{T}}} (\boldsymbol{p}_j^t[k])^* \boldsymbol{h}_{j,j}^{r,t}[k] (\boldsymbol{h}_{j,j}^{r,t}[k])^* \boldsymbol{p}_j^t[k]}{\sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{t=1}^{N_T} \sum_{l=1, l\neq j}^{U} (\boldsymbol{p}_l^t[k])^* \boldsymbol{h}_{j,l}^{r,t}[k] (\boldsymbol{h}_{j,l}^{r,t}[k])^* \boldsymbol{p}_l^t[k] + \omega_j}, \\ &= \frac{\sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{t=1}^{N_{\mathrm{T}}} |\langle \boldsymbol{p}_j^t[k], \boldsymbol{h}_{j,j}^{r,t}[k] \rangle|^2}{\sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{t=1}^{N_{\mathrm{T}}} \sum_{l=1, l\neq j}^{U} |\langle \boldsymbol{p}_l^t[k], \boldsymbol{h}_{j,l}^{r,t}[k] \rangle|^2 + \omega_j},\end{aligned} \tag{5}$$

$\forall j \in [1, U]$, with

$$\omega_j = \sigma^2 \sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{i=1}^{N} |\chi_{j,i}^r|^2 \in \mathbb{R} \tag{6}$$

as the resulting noise components after the detection stage. In order to evaluate the capabilities and limitations of the closed-loop pre-equalisation scheme, we make use of a low complexity detector which has the benefit of reducing cost, size and weight at the receiver. It is for this reason that the single-user MF ($\boldsymbol{\chi}_j^r = \boldsymbol{c}_j$, for $j \in [1, U]$) is applied for detection on each antenna at the receiver of the MUs, see Fig. 1. We emphasise that in the context of pre-equalisation, a MF has the advantage of not requiring knowledge of the channel or other users information for detection, only the user's signature spreading sequence. It is also important to remark that the assessment of the closed-loop pre-equalisation scheme using a MF at receiver will determine the upper-bound of power consumption from the users viewpoint. Since MF does not combat interferences, then the closed-loop pre-equalisation scheme will be the only structure in the system dealing with the interference factor. Other advanced signal processing techniques for signal detection (multiuser detectors) are possible, at the expense of increasing the receiver complexity, but more importantly, it is our interest to investigate the performance of the closed-loop pre-equalisation in the worst-case scenario. Finally, for the downlink transmission, there is practically no restriction in the computational complexity at the transmitter (BS).

3 Optimal closed-loop pre-equalisation for a MIMO–MC-CDMA system

In this section, we derive the optimal closed-loop pre-equalisation for the MIMO–MC-CDMA system following the same methodology proposed in [23, 25]. The objective of this pre-equalisation scheme is to minimise the users' transmission power under the constraints of the QoS requirement of each user, measured in terms of the SINR. In order to support a system with U active users, also U QoS constraints are generated. Note that any user can have multiple QoS requirements over time, but only one at a particular time instant k. For a desired bit-error probability performance for the jth user, the QoS requirement can be

$$\begin{aligned}&\mathcal{E}\{|\hat{b}_j[k]|^2\} \\ &= \sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{t=1}^{N_{\mathrm{T}}} \left\{ \sum_{i=1}^{N} \chi_{j,i}^r (h_{j,i}^{r,t}[k])^* c_{j,i} (p_{j,i}^t[k])^* \times \sum_{i=1}^{N} (\chi_{j,i}^r)^* h_{j,i}^{r,t}[k] c_{j,i} p_{j,i}^t[k] \right\} \\ &\quad + \sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{t=1}^{N_{\mathrm{T}}} \sum_{\substack{l=1,\\ l\neq j}}^{U} \left\{ \sum_{i=1}^{N} \chi_{j,i}^r (h_{j,i}^{r,t}[k])^* c_{l,i} (p_{l,i}^t[k])^* \times \sum_{i=1}^{N} (\chi_{j,i}^r)^* h_{j,i}^{r,t}[k] c_{l,i} p_{l,i}^t[k] \right\} + \sigma^2 \sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{i=1}^{N} |\chi_{j,i}^r|^2.\end{aligned} \tag{4}$$

$$\mathcal{E}\{|\hat{b}_j[k]|^2\} = \sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{t=1}^{N_{\mathrm{T}}} (\boldsymbol{p}_j^t[k])^* \boldsymbol{h}_{j,j}^{r,t}[k] (\boldsymbol{h}_{j,j}^{r,t}[k])^* \boldsymbol{p}_j^t[k] + \sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{t=1}^{N_{\mathrm{T}}} \sum_{\substack{l=1,\\ l\neq j}}^{U} (\boldsymbol{p}_l^t[k])^* \boldsymbol{h}_{j,l}^{r,t}[k] (\boldsymbol{h}_{j,l}^{r,t}[k])^* \boldsymbol{p}_l^t[k] + \sigma^2 \sum_{r=1}^{N_{\mathrm{R}_j}} \sum_{i=1}^{N} |\chi_{j,i}^r|^2$$

translated into an objective SINR after signal detection per MU, i.e.

$$\gamma_j[k] = \gamma_j^{\text{obj}} \quad j \in [1, U]. \tag{7}$$

The appropriate selection of the pre-equalisation factors is indeed a way to meet the QoS for the multiuser MIMO–MC-CDMA system. This objective can be attained by controlling the transmission signals (power and phase) to reach the given thresholds of SINRs. It is important to highlight that phase and power compensations are needed to achieve this goal under the assumption of a single-user detection strategy at the mobile unit [23, 25]. By contrast, when power adaptation is solely utilised (without phase compensation), the power control strategy fails to reach the target QoS. From the SINR expression in (5), and by an extension of the single carrier results for the minimum-norm solution in [25, 26], the pre-equalisation factors for the jth user and the tth transmitting antenna have to satisfy simultaneously the following linear restrictions:

$$\langle \boldsymbol{p}_j^t[k], \boldsymbol{h}_{j,l}^{r,t}[k] \rangle = \begin{cases} \sqrt{\dfrac{\gamma_j^{\text{obj}} \omega_j}{N_{\text{T}} N_{\text{R}_j}}} & l = j \\ 0 & l \neq j \end{cases} \quad l \in [1, U], \quad r \in [1, N_{\text{R}_j}]. \tag{8}$$

These conditions can be interpreted as a zero-forcing property where the MAI is cancelled at each receive antenna, and the data energy of the targeted MU is divided equally among the transmit antennas, so that after detection, the data energy associated to the desired SINR is achieved.

We highlight that the pre-equalisation factors, $p_{j,i}^{r,t}[k]$, linked to each t antenna and subcarrier at the transmitter are adopted to induce diversity in the system. Also, an important property is that the weighted channel gains $\{\boldsymbol{h}_{l,j}^{r,t}[k]\}$ define a linear independent set for all j [26]. In this context, the following remarks can be made about the global optimisation problem:

- In a single-input single-output (SISO) MC-CDMA system with U MUs [25], there are $U \cdot N$ pre-equalisation terms and U QoS restrictions. Since $N > 0$, the global problem for this scenario is overdetermined and then a solution is always guaranteed.
- In the overall MIMO–MC-CDMA system, by considering all MUs, there are $N_{\text{T}} \cdot N \cdot U$ diversity variables (pre-equalisation factors) in the transmitter, but from (8), there are $N_{\text{T}} \cdot U \cdot \sum_{j=1}^{U} N_{\text{R}_j}$ linear restrictions. In this case, if the condition $N \geq \sum_{j=1}^{U} N_{\text{R}_j}$ is met the minimum-norm solution is assured. Otherwise, the optimal solution cannot be obtained.

Observe that when $N < \sum_{j=1}^{U} N_{\text{R}_j}$ in a MIMO–MC-CDMA system, the QoS condition for each MU could not be guaranteed since it is not feasible to cancel completely the MAI term at each receive antenna (there are more restrictions that free variables) and provide the desired data energy, simultaneously. While indeed this becomes a limiting condition, to provide a feasible and practical solution to the problem of pre-equalisation, with the minimum possible power consumption, still constitutes an important contribution. If fact, such condition will always be satisfied for the (multiple-input single-output) MISO–MC-CDMA system case (since the number of MUs U does not exceed the processing gain N), where the use of a single antenna may be more appealing to reduce further the complexity at the mobile receiver. Nonetheless, some channel gains could have better responses than others, therefore, adding more receive antennas may provide some additional improvements regarding the total transmission energy.

3.1 Distributed perspective

By carefully inspecting the restrictions in (8), we can observe that the pre-equalisation factors $\boldsymbol{p}_j^t[k]$ could be computed independently per each transmitting antenna t and MU j in a distributed perspective. Assuming that $N \geq \sum_{j=1}^{U} N_{\text{R}_j}$, and recalling the classical result in [27], the optimal solution for the jth user and the tth transmitting antenna that satisfy the restrictions in (8) is a linear combination of the vectors $\{\boldsymbol{h}_{j,l}^{r,t}[k]\}_{l\in[1,U],\, r\in[1,N_{\text{R}}]}$:

$$\begin{aligned} \boldsymbol{p}_j^t[k] &= \sum_{r=1}^{N_{\text{R}_j}} \sum_{l=1}^{U} \beta_{l,r}^{j,t} \boldsymbol{h}_{j,l}^{r,t}[k] \quad j \in [1, U], \ t \in [1, N_{\text{T}}] \\ &= \mathcal{H}_{j,t}[k] \boldsymbol{\beta}_{j,t} \end{aligned} \tag{9}$$

where

$$\mathcal{H}_{j,t}[k] = \left[\boldsymbol{h}_{j,1}^{1,t}[k] \quad \ldots \quad \boldsymbol{h}_{j,U}^{1,t}[k] \quad \ldots \quad \boldsymbol{h}_{j,1}^{N_{\text{R}_j},t}[k] \quad \ldots \quad \boldsymbol{h}_{j,U}^{N_{\text{R}_j},t}[k] \right] \in \mathbb{C}^{N \times \hat{N}_{\text{R}_j}}, \tag{10}$$

$$\boldsymbol{\beta}_{j,t} = \left[\beta_{1,1}^{j,t} \quad \ldots \quad \beta_{U,1}^{j,t} \quad \ldots \quad \beta_{1,N_{\text{R}_j}}^{j,t} \quad \ldots \quad \beta_{U,N_{\text{R}_j}}^{j,t} \right]^{\top} \in \mathbb{C}^{\hat{N}_{\text{R}_j}}, \tag{11}$$

with $\hat{N}_{\text{R}_j} = \sum_{j=1}^{U} N_{\text{R}_j}$. In the optimal solution structure in (9), the coefficients $\{\beta_{l,r}^{j,t}\}$ represent the scaling variables for the vectors and $\{\boldsymbol{h}_{j,l}^{r,t}[k]\}$ that denote a base for the solution space [27]. Therefore, by a direct substitution of (9) into (8), the coefficients $\beta_{l,r}^{j,t}$ can be computed by solving the following system of $\hat{N}_{\text{R}_j}$ linear equations:

$$\sum_{\hat{r}=1}^{N_{\text{R}_j}} \sum_{\hat{l}=1}^{U} \beta_{\hat{l},\hat{r}}^{j,t} \langle \boldsymbol{h}_{j,\hat{l}}^{\hat{r},t}[k], \boldsymbol{h}_{j,l}^{r,t}[k] \rangle = \begin{cases} \sqrt{\dfrac{\gamma_j^{\text{obj}} \omega_j}{N_{\text{T}} N_{\text{R}_j}}} & l = j \\ 0 & l \neq j \end{cases} \quad l \in [1, U], \ r \in [1, N_{\text{R}_j}], \tag{12}$$

which is equivalent to

$$(\mathcal{H}_{j,t}[k])^* \mathcal{H}_{j,t}[k] \boldsymbol{\beta}_{j,t} = \boldsymbol{e}_{j,t}, \tag{13}$$

where

$$\boldsymbol{e}_{j,t} = \left[\sqrt{\frac{\gamma_j^{\text{obj}} \omega_j}{N_{\text{T}} N_{\text{R}_j}}} \quad \ldots \quad 0 \quad \ldots \quad 0 \quad \ldots \quad \sqrt{\frac{\gamma_j^{\text{obj}} \omega_j}{N_{\text{T}} N_{\text{R}_j}}} \right]^{\top} \in \mathbb{R}^{\hat{N}_{\text{R}_j}}, \tag{14}$$

i.e. the coefficients $\{\beta_{l,r}^{j,t}\}$ or equivalently vector $\boldsymbol{\beta}_{j,t}$ depends on the pseudo-inverse of matrix $\mathcal{H}_{j,t}[k]$.

In this way, the weighted minimum norm problem for the jth MU and t-transmitter antenna is formulated as

$$\begin{aligned} &\min \frac{1}{2} (\boldsymbol{p}_j^t[k])^* \boldsymbol{\Theta}_{j,t}^{-1} \boldsymbol{p}_j^t[k] \\ &\text{such that} \quad \gamma_j[k] = \gamma_j^{\text{obj}} \end{aligned} \tag{15}$$

where $\boldsymbol{\Theta}_{j,t} \in \mathbb{C}^{N \times N}$ and $\boldsymbol{\Theta}_{j,t} > 0$ is a free weight matrix in the quadratic cost. The optimal pre-equalisation vector that solves the previous problem is

$$\tilde{\boldsymbol{p}}_j^t[k] = \boldsymbol{\Theta}_{j,t} \mathcal{H}_{j,t}[k] \left\{ (\mathcal{H}_{j,t}[k])^* \boldsymbol{\Theta}_{j,t} \ \mathcal{H}_{j,t}[k] \right\}^{-1} \boldsymbol{e}_{j,t}. \tag{16}$$

This optimal solution will minimise the power of the pre-equalisation factors of the jth user at the tth transmitting antenna for a given weight matrix $\boldsymbol{\Theta}_{j,t}$.

3.2 Centralised perspective

Although a distributed methodology could be appealing for implementation purposes, the global transmitting power in the downlink could not be efficiently assigned without sharing knowledge of the remaining MUs and their pre-equalisation factors in the transmitting antennas. Therefore, a centralised strategy could bring a best distribution of the power without sacrificing QoS at the price of an increased complexity. Nonetheless, the BS could manage efficiently this raise in complexity. For this purpose, the pre-equalisation factors can be computed with a centralised approach by stacking the pre-equalisation elements of all MUs and transmitting antennas into a global vector, and arranging the linear restrictions into a block diagonal matrix. Departing from the matrix notation for $\mathcal{H}_{j,t}$ in (10) and $\boldsymbol{e}_{j,t}$ in (14), the linear restrictions in (8) can be written as

$$(\mathcal{H}_{j,t}[k])^* \boldsymbol{p}_j^t[k] = \boldsymbol{e}_{j,t} \quad \forall j, t. \tag{17}$$

Therefore, by rewriting all the matrices and vectors as

$$\mathcal{H}[k] = \operatorname{diag}\Big(\mathcal{H}_{1,1}[k], \dots, \mathcal{H}_{U,1}[k], \dots, \mathcal{H}_{1,N_\mathrm{T}}[k], \dots, \mathcal{H}_{U,N_\mathrm{T}}[k]\Big) \in \mathbb{C}^{N \cdot N_\mathrm{T} \cdot U \times N_\mathrm{T} \cdot U \cdot \hat{N}_{\mathrm{R}_j}}, \tag{18}$$

$$\boldsymbol{e} = \left[\boldsymbol{e}_{1,1}^\top, \dots, \boldsymbol{e}_{U,1}^\top, \dots, \boldsymbol{e}_{1,N_\mathrm{T}}^\top, \dots, \boldsymbol{e}_{U,N_\mathrm{T}}^\top\right]^\top \in \mathbb{R}^{N_\mathrm{T} \cdot U \cdot \hat{N}_{\mathrm{R}_j}}, \tag{19}$$

$$\boldsymbol{p}[k] = \left[(\boldsymbol{p}_1^1[k])^\top, \dots, (\boldsymbol{p}_U^1[k])^\top, \dots, (\boldsymbol{p}_1^{N_\mathrm{T}}[k])^\top, \dots, (\boldsymbol{p}_U^{N_\mathrm{T}}[k])^\top\right]^\top \in \mathbb{C}^{N \cdot N_\mathrm{T} \cdot U} \tag{20}$$

the overall linear restrictions are described by the system of $N_\mathrm{T} \cdot U \cdot \hat{N}_{\mathrm{R}_j} \times N \cdot N_\mathrm{T} \cdot U$ equations

$$(\mathcal{H}[k])^* \, \boldsymbol{p}[k] = \boldsymbol{e}. \tag{21}$$

In this framework, the weighted minimum norm problem for the overall MIMO system is

$$\begin{aligned} &\min \frac{1}{2} (\boldsymbol{p}[k])^* \boldsymbol{\Theta}^{-1} \boldsymbol{p}[k] \\ &\text{such that} \quad \gamma_j[k] = \gamma_j^{\text{obj}} \quad \forall j \in [1, U] \end{aligned} \tag{22}$$

where $\boldsymbol{\Theta} \in \mathbb{C}^{N \cdot N_\mathrm{T} \cdot U \times N \cdot N_\mathrm{T} \cdot U}$ and $\boldsymbol{\Theta} > 0$ is a weight matrix for the total transmission power in the quadratic cost. As a result, the optimal MIMO pre-equalisation vector is

$$\tilde{\boldsymbol{p}}[k] = \boldsymbol{\Theta} \mathcal{H}[k] \big\{ (\mathcal{H}[k])^* \boldsymbol{\Theta} \, \mathcal{H}[k] \big\}^{-1} \boldsymbol{e}. \tag{23}$$

As pointed out earlier, although both solutions in (16) and (23) cancel the MAI after signal detection at the receiver and achieve the desired SINR, the resulting pre-equalisation factors and power distribution are different and depend on the weight matrices $\boldsymbol{\Theta}_{j,t}$ and $\boldsymbol{\Theta}$. An important advantage of these solutions is that under a specific condition they can be computed recursively [23]. Nonetheless, the recursive approach assumes that the channel variations are slowly time-varying or piece-wise constant over certain time windows, in order to reach convergence. However, since the pre-equalisation updating is generally much faster than the channel dynamics, these assumptions are practical [28]. For this reason, the time index (k) will be dropped from $\mathcal{H}_{j,t}$ or $\mathcal{H}$, and the scalars and matrices that depend on the channel information.

4 Optimal approximate solution

The optimal solutions in (16) and (23) for $N \geq \sum_{j=1}^{U} N_{\mathrm{R}_j}$ can be approached iteratively [23], and for this purpose, we define

$$\widehat{\mathcal{H}} = \begin{cases} \mathcal{H}_{j,t} & \text{Distributed} \\ \mathcal{H} & \text{Centralised} \end{cases}, \quad \widehat{\boldsymbol{\Theta}} = \begin{cases} \boldsymbol{\Theta}_{j,t} & \text{Distributed} \\ \boldsymbol{\Theta} & \text{Centralised} \end{cases} \tag{24}$$

$$\widehat{\boldsymbol{p}}[k] = \begin{cases} \boldsymbol{p}_{j,t}[k] & \text{Distributed} \\ \boldsymbol{p}[k] & \text{Centralised} \end{cases}, \quad \widehat{\boldsymbol{e}}[k] = \begin{cases} \boldsymbol{e}_{j,t}[k] & \text{Distributed} \\ \boldsymbol{e}[k] & \text{Centralised.} \end{cases} \tag{25}$$

In this way, the resulting dynamic structure is

$$\begin{aligned} \boldsymbol{x}[k+1] &= \Big\{ \boldsymbol{I} - (\widehat{\mathcal{H}})^* \widehat{\boldsymbol{\Theta}} \widehat{\mathcal{H}} \Big\} \boldsymbol{x}[k] + \widehat{\boldsymbol{e}} \cdot 1[k], \\ \widehat{\boldsymbol{p}}[k] &= \widehat{\boldsymbol{\Theta}} \widehat{\mathcal{H}} \boldsymbol{x}[k], \end{aligned} \tag{26}$$

where $\boldsymbol{x}[k] \in \mathbb{C}^N$ identifies the case of the distributed scheme and $\boldsymbol{x}[k] \in \mathbb{C}^{N \cdot N_\mathrm{T} \cdot U}$ of the centralised one, and $1[k]$ represents the step function. If the spectral radius $\rho(\cdot)$ satisfies the condition

$$\rho \Big\{ \boldsymbol{I} - (\widehat{\mathcal{H}})^* \widehat{\boldsymbol{\Theta}} \widehat{\mathcal{H}} \Big\} < 1, \tag{27}$$

then $\widehat{\boldsymbol{p}}[k] \to \tilde{\boldsymbol{p}}_j^t$ (distributed) or $\widehat{\boldsymbol{p}}[k] \to \tilde{\boldsymbol{p}}$ (centralised). In fact, by a direct extension of the single-carrier result in [23], a positive definite weight matrix $\widehat{\boldsymbol{\Theta}}$ that satisfies (27) is given by

$$\widehat{\boldsymbol{\Theta}} = \alpha \Big(\boldsymbol{U} \boldsymbol{\Sigma}^{-2} \boldsymbol{U}^* \Big) > 0, \tag{28}$$

where $\boldsymbol{U}$, $\boldsymbol{V}$ and $\boldsymbol{\Sigma}$ are obtained from the reduced singular value decomposition of $\widehat{\mathcal{H}} = \boldsymbol{U} \boldsymbol{\Sigma} \boldsymbol{V}^*$. The scalar $\alpha \in (0, 1)$ defines a convergence rate parameter such that

$$\rho \Big\{ \boldsymbol{I} - (\widehat{\mathcal{H}})^* \widehat{\boldsymbol{\Theta}} \widehat{\mathcal{H}} \Big\} = 1 - \alpha. \tag{29}$$

The previous result clearly shows that if $\alpha \to 1$, then the convergence speed to the optimal solution is increased. Observe that iterative methods for solving linear equations, such as (16), can be considered as a discrete-time control system. Using this perspective, the iterative approach in (26) can be rewritten by defining an error vector $\boldsymbol{a}[k]$, as follows:

$$\begin{aligned} \boldsymbol{x}[k+1] &= \boldsymbol{x}[k] + \boldsymbol{a}[k] \\ \boldsymbol{a}[k] &= \widehat{\boldsymbol{e}} - (\widehat{\mathcal{H}})^* \widehat{\boldsymbol{p}}[k] \quad \forall k \geq 0 \end{aligned} \tag{30}$$

with $\widehat{\boldsymbol{p}}[k]$ as given in (26). Hence, if (27) is satisfied, then $\boldsymbol{a}[k] \to 0$ as the iterative solution converges to (16) or (23) when $k \to \infty$. Based on the representation of the iterative approach by (30), a pre-equalisation technique can be implemented using a feedback control strategy, with respect to the convergence error $\boldsymbol{a}[k]$, for the multiuser MIMO–MC-CDMA system. This strategy requires that the MUs feedback information to the BS, in a closed-loop fashion, to compute $\widehat{\mathcal{H}}$ and $\widehat{\boldsymbol{e}}$ in order to evaluate $\boldsymbol{a}[k]$, and next, the BS must adjust the pre-equalisation factors for the multiuser MIMO–MC-CDMA system using an integral control action, as described in (30). Observe that the closed-loop implementation requires to update the product $\widehat{\boldsymbol{\Theta}}\widehat{\mathcal{H}}$ at the BS (needed to compute $\widehat{\boldsymbol{p}}[k]$), but since $\widehat{\boldsymbol{\Theta}}$ is a constant matrix, just $\widehat{\mathcal{H}}$ needs to be eventually updated. We remark also that $\widehat{\mathcal{H}}$ in (24) depends on the channel gains (constant for long periods of time in most common scenarios), and spreading codes of all MUs. Therefore, the information of $\widehat{\boldsymbol{\Theta}}\widehat{\mathcal{H}}$ does not need to be updated in each iteration algorithm. Furthermore, an interesting feature of the iterative algorithm is its distributed property with respect to the channel

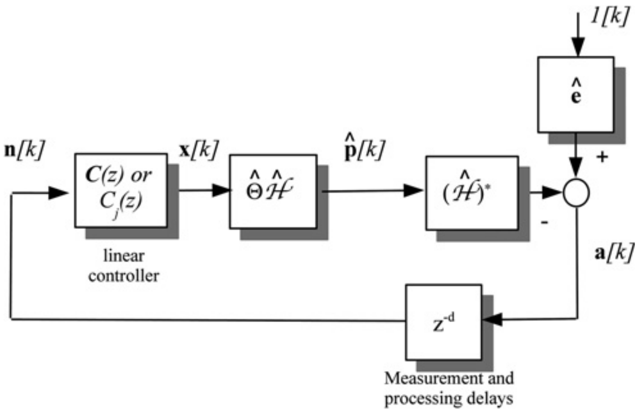

Fig. 2 *General iterative pre-equalisation scheme in the downlink*

gains of the remaining MUs. Hence, the assignment strategy in (30) is feasible for a practical MIMO–MC-CDMA scenario.

For robustness and transient performance, a general distributed controller with respect to each MU is suggested

$$\boldsymbol{C}(z) = \begin{bmatrix} C_1(z) & & \\ & \ddots & \\ & & C_U(z) \end{bmatrix} \tag{31}$$

where $C_j(z)\ \forall\, j \in [1, U]$ represents the transfer function of the linear controller for the jth user in $\mathcal{Z}$-domain, which is common to all the transmitting antennas. Hence, departing from the closed-loop structure in (30), replacing the integral control action by (31) for any transmitter antenna, introducing d delays after the error quantification, and representing the error signal in the BS as $\boldsymbol{n}[k]$ after the quantification delays, a new closed-loop interaction is obtained and shown in Fig. 2. For $C_j(z)$, we suggest without loss of generality the optimal LQ distributed controller proposed in [15] that takes into account the effect of measurement and processing delays

$$C_j^{\mathrm{LQ}}(z) = \frac{\Omega z^{-1}}{1-(1-\Omega)z^{-1} - \Omega z^{-d-1}} \quad j \in [1, U], \tag{32}$$

where $0 < \Omega < 1$ is a free tuning parameter.

5 Simulation results

The performance of the closed-loop pre-equalisation scheme is now presented and analysed in terms of the QoS requirements and transmission power under the framework of the MIMO–MC-CDMA system. For the sake of brevity, we only show the centralised case. In all evaluations, binary phase-shift keying is assumed as modulation and Walsh–Hadamard codes are used for spreading the users' information through all subcarriers and transmit antennas. It is also assumed that each user goes through a different multipath channel, modelled as a 4-tap impulse response with coefficients that follow a Rayleigh distribution. Without loss of generality, the same target SINR will be considered for all active users in the evaluations, i.e. $\gamma_j^{\mathrm{obj}} = \gamma^{\mathrm{obj}}$ for $j \in [1, U]$. The overall channel and simulation parameters employed through all simulations are summarised in Table 1. The effect of the convergence parameter α in (28) was independently analysed, if α is close to 1.0, the convergence to the objective SINR is faster, but also the transient response to channel variations is increased. Hence the value of α was set to 0.4 as a good balance between fast convergence and robustness which satisfies the global stability condition. Recall that a single-user MF is applied for detection at each MU. We have also restricted our simulation results solely to the condition where all users' devices use the same number of receive antennas, then for convenience the subscript j will be dropped from N_{R_j} to N_{R}.

Table 1 Parameters of the MIMO–MC-CDMA system

Physical parameter	Variable	Value
number of subcarriers	N_s	32
processing gain	N	32
noise variance	σ^2	–25 dBm
minimum transmission power	$p_{\min}$	1 pW
maximum transmission power	$p_{\max}$	500 mW
control gain	Ω	0.4
roundtrip delay	d	2

As a performance benchmark, we first evaluate the averaged transmission power required by the proposed closed-loop pre-equalisation scheme for a target SINR of 8 dB as a function of the number of active users. An MC-CDMA system with two transmit antennas and one receive antenna is considered to isolate the diversity gains in the transmitter for the proposed pre-equalisation scheme. Fig. 3 shows the simulation results and compared with a relevant scheme, based on channel inversion, called pre-equalisation with orthogonality restoring combining (PRE-ORC) [20]

$$p_{j,i}^{t}[k] = \frac{1}{h_{j,i}^{1,t}[k]} \quad \text{PRE}-\text{ORC}. \tag{33}$$

In order to provide with a fair comparison, both schemes assume perfect CSI knowledge at the transmitter (BS), and that an MF detector is applied for signal detection at each MU. It is important to observe that both schemes require complex pre-equalisation factors, i.e. they need the estimation of power and signal phase. The results show that the proposed scheme outperforms the PRE-ORC scheme by reducing significantly the power required for the pre-equalisation factors. While the distinctive strength of the proposed solution lies in the fact that it minimises the power of the pre-equalisation factors to cancel the interference on the system, PRE-ORC increases the power consumption to compensate for deep fades. We remark that in order to perform the comparison in a common scenario and understand the differences between both solutions, we have not placed a power constraint on the pre-equalisation factors for PRE-ORC so that both solutions can guarantee to eliminate completely the system's distortions. Adding a transmission power constraint to PRE-ORC will represent to lose the orthogonality of the MC-CDMA system and therefore resulting in performance degradation.

We now evaluate the performance of the closed-loop pre-equalisation scheme in the SISO antenna configuration. This case can be seen as a reference scenario to illustrate the impact of the proposed pre-equalisation scheme by adding more antennas both at the transmitter and receiver. Fig. 4 shows the achieved average SINR and transmission power consumption for different system loads as a function of the iterations in the closed-loop scheme. The plots present the evaluation of four cases: 25, 50, 75 and 100% of loading factor. To illustrate the effect of channel variations, the channel impulse gains for each user are updated every 100 iterations of the closed-loop structure. Also, two different target SINRs at 8 and 10 dB are considered. Fig. 4*a* shows that every time the channel is updated, the target SINR is always reached within very few iterations in a similar fashion for all different system loads. Meanwhile, Fig. 4*b* depicts the average transmission power demand for the loading cases under consideration. As expected, the results indicate that the closed-loop pre-equalisation requires to increase the average transmission power levels in order to overcome the system's distortions, given in the form of multiple access interference, multipath channel propagation and noise. We observe, however, that the average power required in the 100% loading system case is smaller for the 300–400 iterations interval than during the first 100 iterations,

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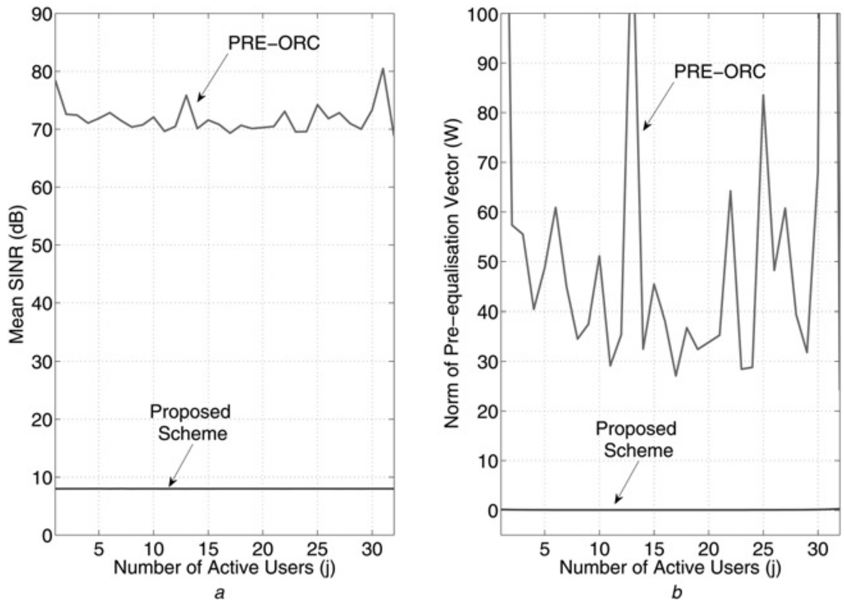

Fig. 3 *Comparison of PRE-ORC with the proposed closed-loop pre-equalisation scheme under a common scenario with two transmit antennas and one receive antenna*

a Mean SINR
b Norm of pre-equalisation vector $||\boldsymbol{p}[k]||$

where the target SINR is lower. Since the system's conditions are the same for both scenarios (8 and 10 dB SINR targets), we should expect initially to necessitate higher power transmissions to reach the 10 dB SINR target as compared with the 8 dB target. However, this particular behaviour can be explained from the fact that the level of interference in the system depends significantly on the instantaneous channel impulse response. The results in Fig. 4 include four different random channel realisations (one for each interval of 100 iterations), then the overall instantaneous interferences for each 100 iterations interval are not the same. This situation can also be seen during the last 200 iterations of the same full loaded system case, note that two different levels of transmission powers (average) are required for the same SINR target. Still, we can remark that the closed-loop pre-equalisation scheme needs to increase the transmission power in a controlled manner to compensate properly the system's perturbations.

Next, we assess the effect of adding more degrees of freedom to the closed-loop equalisation scheme solution in (16) through the use of multiple antennas at the BS. An MISO configuration is now considered in order to separate the influence of the receive antennas. The simulation results are given in Fig. 5 for a MISO–MC-CDMA system using $N_T = 1$, 2, 4 and 8 and only one receive antenna, $N_R = 1$, for a 75% loading system. Fig. 5*a* shows that the users' SINR converge to its objective value in the same fashion to

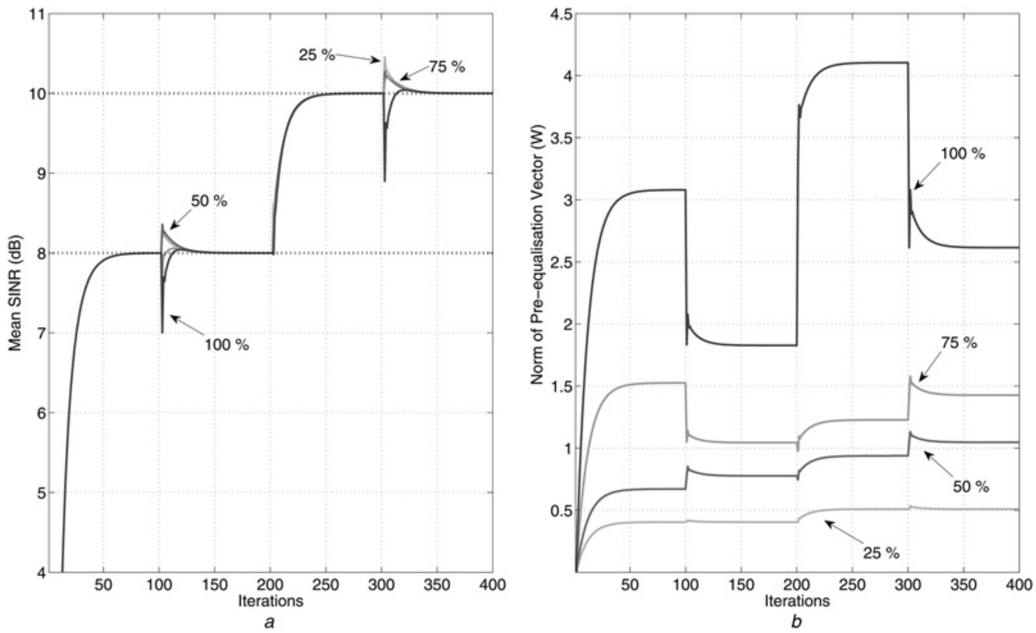

Fig. 4 *Downlink closed-loop pre-equalisation performance for a SISO MC-CDMA system under a 25, 50, 75 and 100% loaded conditions*

a Mean SINR
b Norm of pre-equalisation vector $||\boldsymbol{p}[k]||$

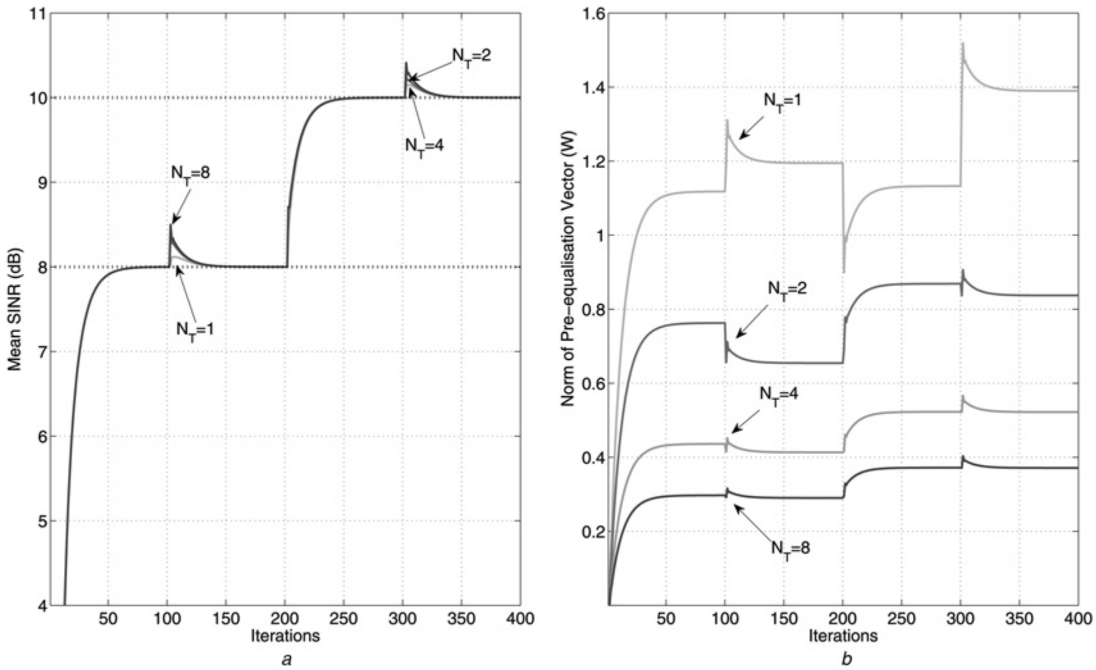

Fig. 5 *Downlink closed-loop pre-equalisation performance for a 75% loaded MISO–MC-CDMA system*

a Mean SINR
b Norm of pre-equalisation vector $||\boldsymbol{p}[k]||$

those results presented in Fig. 4, for all scenarios. More important, Fig. 5*b* in Fig. 4 illustrates how the closed-loop pre-equalisation scheme converts the spatial diversity gain, induced by the multiple antennas at the BS, into a notable power reduction. It is clear that the pre-equalisation factors contribute to the compensation of the system's distortions by reducing the minimum power consumption while maintaining the objective QoS. In other words, the distinct pre-equalisation factors per subcarrier on each antenna act as additional degrees of freedom for the system, which is consistent with the conjecture presented in Section 3.

The closed-loop pre-equalisation performance is now analysed in a fully MIMO configuration. Fig. 6 displays the performance for 1, 2 and 4 receive antennas using $N_T = 4$ transmit antennas, i.e. a system with $N_T \cdot N$ pre-equalisation factors or degrees of freedom. It is seen

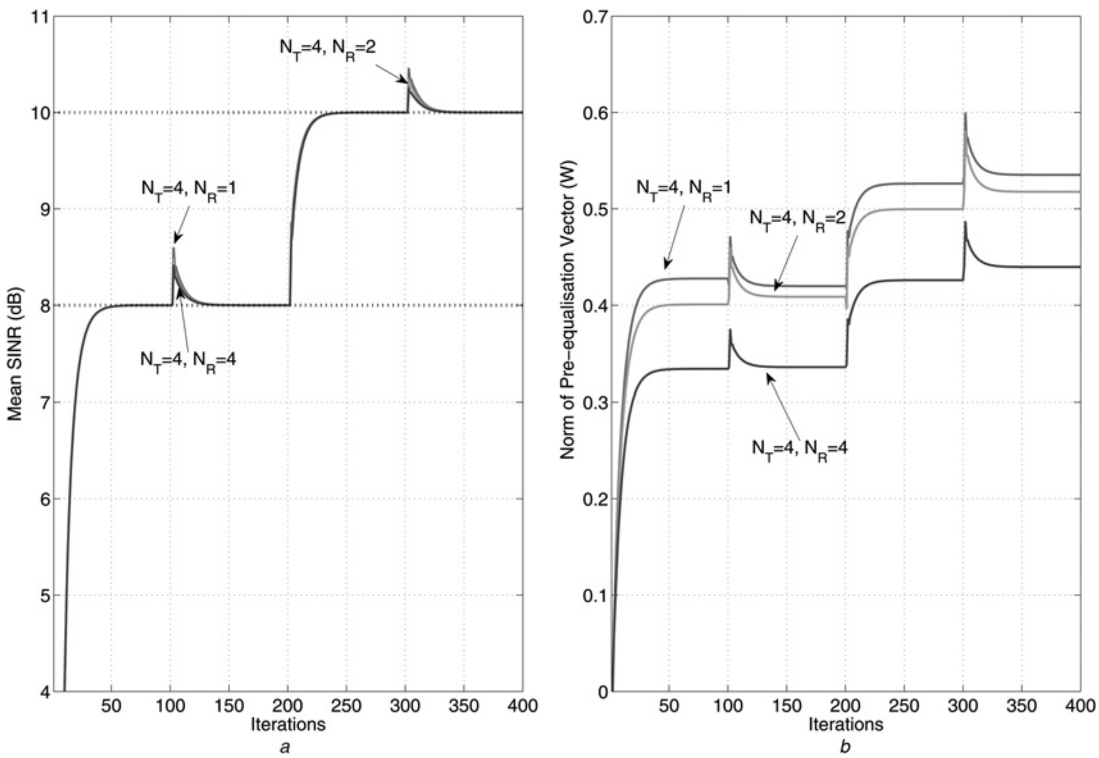

Fig. 6 *Downlink closed-loop pre-equalisation performance for a 75% loaded MIMO–MC-CDMA system*

a Mean SINR
b Norm of pre-equalisation vector $||\boldsymbol{p}[k]||$

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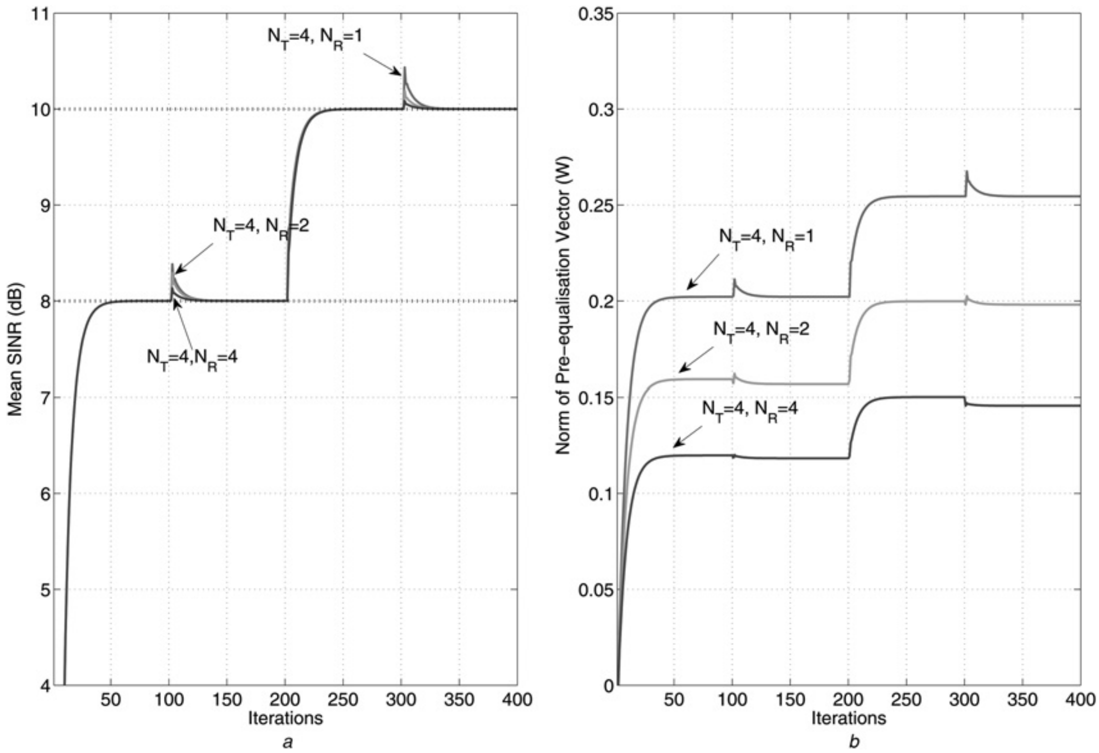

Fig. 7 *Downlink closed-loop pre-equalisation performance for a 25% loaded MIMO–MC-CDMA system*

a Mean SINR
b Norm of pre-equalisation vector $||\boldsymbol{p}[k]||$

that only a small power reduction can be achieved by increasing the number of receive antennas. This behaviour can be explained by the fact that each user's receiver has to deal with much higher levels of interference that grows with the factor N_R. Moreover, the condition $N \geq \sum_{j=1}^{U} N_{R_j} = N_R \cdot U$ is not satisfied, therefore the optimal solution for the closed-loop pre-equalisation is not guaranteed. Despite having this limiting condition, the closed-loop pre-equalisation scheme manages to reach the target SINR but the benefit of using spatial diversity from the receive antennas is negligible. For the comparison, Fig. 7 shows the same evaluations but in order to fulfil the condition $N \geq N_R \cdot U$, the number of user is set to $U = 8$ (25% load). It is clear that the benefit of using multiple receive antennas comes with a penalty in the number of users that can be allocated in the system. This limitation could overcome by adding multiuser detection or redefining the linear restrictions in (8) according to the channel gains. In some conditions, some channels could have better responses than others, and equally dividing the data energy according to (8) could result in an increment of the total transmission energy. However, these approaches are not straightforward and future studies needs to be conducted in order to exploit spatial diversity at the receiver of the closed-loop pre-equalisation scheme.

6 Conclusions

In this paper, the performance of closed-loop pre-equalisation under a multiuser MIMO–MC-CDMA-based system was investigated. As a first step, we formulated and derived the optimal closed-loop pre-equalisation solution for the combined MIMO–MC-CDMA system from a distributed and centralised viewpoint. The analysis demonstrates that multiple-access interference and channel variations are completely cancelled by the proposed pre-equalisation scheme. More importantly, this compound system reduces power consumption by capitalising on the number of subcarriers and multiple antennas in the system while meeting individual QoS targets. Simulation results confirm the above benefits and show the robustness of the closed-loop pre-equalisation concept in a highly loaded system with severe channel conditions. However, the use of multiple antennas at the receiver is restricted to the condition that $N \geq \sum_{j=1}^{U} N_{R_j}$. Therefore, further research is suggested to identify a more effective way to exploit spatial diversity at the receiver side of the proposed scheme.

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