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Downlink Space-Frequency Preequalization Techniques for TDD MC-CDMA Mobile Radio Systems

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The paper considers downlink space-frequency preequalizations techniques for time division duplex (TDD) MC-CDMA. We consider the use of antenna arrays at the base station (BS) and analytically derive different preequalization schemes for two different receiver configurations at the mobile terminal: a simple despread receiver without channel equalization and an equal-gain combiner (EGC) conventional receiver. We show that the space-frequency preequalization approach proposed allows to format the transmitted signals so that the multiple access interference at mobile terminals is reduced allowing to transfer the most computational complexity from mobile terminal to the BS. Simulation results are carried out to demonstrate the effectiveness of the proposed preequalization schemes.

Keywords and phrases: MC-CDMA, preequalization, antenna array, TDD, downlink.

1. INTRODUCTION

The beyond 3G broadband component of wireless system must be able to offer bit rates of more than 2 Mbps in a vehicular environment and at least 10–20 Mbps in indoor and pedestrian environments [1].

It is consensual that MC-CDMA is one of the most promising multiple-access schemes for achieving such high data rates [2]. This scheme combines efficiently orthogonal frequency division multiplex (OFDM) and CDMA. Therefore, MC-CDMA benefits from OFDM characteristics such as high spectral efficiency and robustness against multipath propagation, while CDMA allows a flexible multiple access with good interference properties for cellular environments [3].

Recent publications have shown that MC-CDMA is particularly advantageous for the downlink, that is, from BS to mobile terminal (MT) [4]. However, the user capacity of MC-CDMA system is essentially limited by the multiple-access interference (MAI) provoked by the loss of code orthogonality among the users in multipath propagation. Usually, in conventional MC-CDMA downlink, the MAI is mitigated by frequency domain equalization techniques at receiver side. Since low complexity is required at MTs, only simple detection techniques can be implemented, limiting the MAI cancellation capability. Considering time division duplex (TDD), another solution consists in performing preequalization at the transmitter side using the TDD channel reciprocity between alternative uplink and downlink transmission period [5, 6, 7]. The knowledge of the channel state information (CSI) from uplink can be used to improve the performance in downlink. The crucial assumption is that channel dynamics are sufficiently slow so that the multipath profile remains essentially constant over the block of transmitted symbols. Normally, this principle is valid for indoor and pedestrian environments, that is, in low-mobility scenarios. The aim of this solution is to allow the use of simple low-cost, low-consuming MT.

It is well known that the use of antenna arrays increases the system capacity by reducing the effect of frequency selective fading and improving the spectral efficiency, without additional frequency spectrum [8]. It has been show [9] that the combination of antenna array with MC-CDMA system is very advantageous in cellular communications.

This paper proposes a downlink TDD downlink MC-CDMA space-frequency preequalization algorithm designed

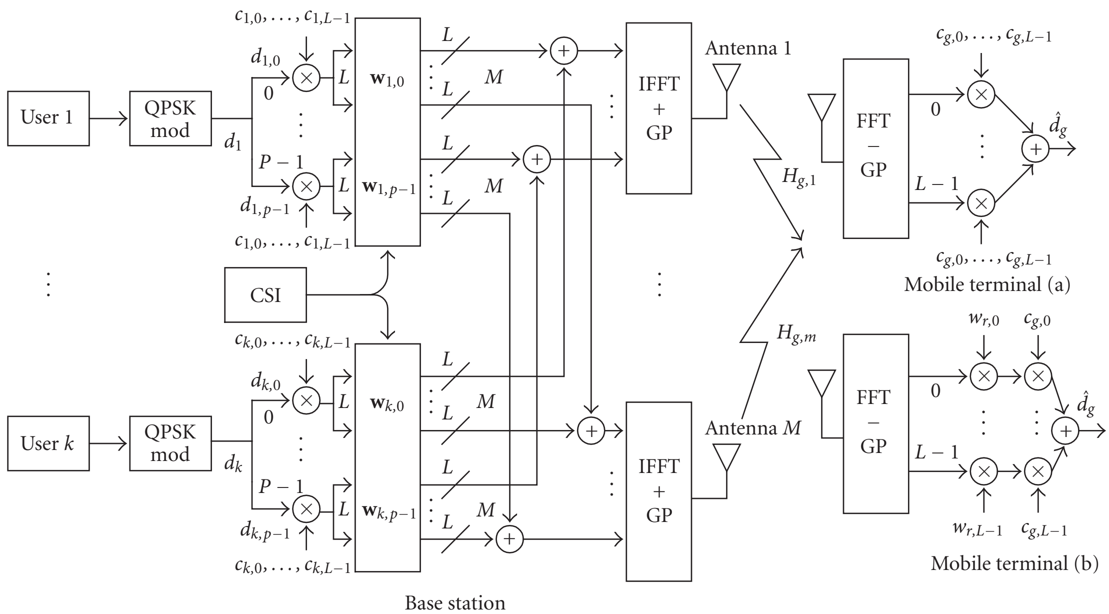

FIGURE 1: Transmitters and receivers schemes for downlink MC-CDMA.

for two different types of receivers: equal-gain combiner (EGC) conventional equalizer and a very simple receiver without channel equalization. Both algorithms operate in the frequency domain and optimization is done in frequency for the single-antenna case, and jointly in space and frequency when considering an antenna array. Moreover, these algorithms are designed using as criterion the minimization of the transmitted power at the BS, which is a very important issue for most preequalization algorithms.

The paper is organized as follows. In Section 2, we present the proposed downlink MC-CDMA system. In Section 3, we analytically derive the preequalization algorithms for the two receiver configurations which we call constrained zero forcing (CZF) and CZF-EGC, respectively. In Section 4, we present some simulation results obtained with the CZF and CZF-EGC preequalization techniques in two different scenarios, beamforming and diversity, and compare both preequalization schemes against conventional equalizer techniques such as MRC, EGC, and MMSE. Finally, the main conclusions are pointed out in Section 5.

2. SYSTEM MODEL

Figure 1 shows the proposed downlink MC-CDMA transmitters and receivers. As presented in Figure 1, for each user k, a complex QPSK data symbol d_k ($k = 1, \ldots, K$) is converted from serial to parallel to produce p symbols, $d_{k,p}$ ($k = 1, \ldots, K$ and $p = 0, \ldots, P-1$), where P denotes the number of data symbols transmitted per OFDM symbol. The data symbols are spread into L chips using the orthogonal Walsh-Hadamard code set and scrambled by a pseudorandom code. We denote the code vector of user k as $\mathbf{c}_k = [c_{k,0}, \ldots, c_{k,L-1}]^T$, where $(\cdot)^T$ is the transpose operator. Then, the chips of the data symbols are copied M times in order to obtain $L \cdot M$ versions of the original symbols which are weighted and transmitted over M antenna branches. The LM chips for user k and symbol p are weighted by a vector $\mathbf{w}_{k,p} = [\mathbf{w}_{k,p,1}^T \ \mathbf{w}_{k,p,1}^T \ \cdots \ \mathbf{w}_{k,p,M-1}^T]^T$ where $\mathbf{w}_{k,p,m}$ of size L contains the set of coefficients that weight the chips that go to antenna m, and thus $\mathbf{w}_{k,p}$ is of size LM. These weights are calculated using the CSI according to the criteria presented in Sections 3.1 and 3.2. After that, the signals of all users on each subcarrier and antenna branch are added to form the multiuser transmitted signal. Finally, a guard period (GP) longer than the channel multipath spread is inserted in the transmitted signal, on each antenna, to avoid intersymbol interference (ISI).

The transmitted signal, in frequency domain, for a generic data symbol p is given by

$$\mathbf{y}_p = \sum_{k=1}^{K} d_{k,p} \widehat{\mathbf{c}}_k \circ \mathbf{w}_{k,p}, \tag{1}$$

where $\widehat{\mathbf{c}}_k = [\mathbf{c}_k^T, \ldots, \mathbf{c}_k^T]^T$ is a column vector of size of $L \cdot M$ that represents the spreading operation and consists of M repetitions of the code vector for user k since the same code is used for all antenna branches, and $(\circ)$ means an elementwise vector product. The vector signal $\mathbf{y}_p$ of length $L \cdot M$ is mapped to the antenna branch so that the first L elements are transmitted over the L subcarriers of the OFDM modulation on the first antenna branch, the second L elements to the second branch, and so on.

The input signal at the generic mobile g, for symbol p, is obtained multiplying (1) for the channel frequency response from the BS to the MT of the desired user and adding AWGN noise:

$$\mathbf{x}_{g,p} = \sum_{m=1}^{M} \sum_{k=1}^{K} d_{k,p} \mathbf{c}_k \circ \mathbf{w}_{k,p,m} \circ \mathbf{h}_{g,p,m} + \mathbf{n}_g, \tag{2}$$

where $\mathbf{h}_{g,p,m}$ of size $L \times 1$ is the channel frequency response between antenna m and MT.

At the receiver side we propose two different receivers: receiver (a) which is composed just by a single antenna, an FFT, despreading and descrambling operations, that is, we do not perform channel equalization, and a conventional EGC single user receiver (b). For this latter case the weights are given by

$$\mathbf{w}_r = \frac{\mathbf{h}^*}{|\mathbf{h}|}, \tag{3}$$

that is, we just perform phase equalization.

For receiver (a), the decision variable at the input of the QPSK demodulator is, for the desired user g and symbol p, given by

$$\begin{aligned} \hat{d}_{g,p} &= \underbrace{d_{g,p} \cdot \mathbf{c}_g \left(\sum_{m=1}^{M} \mathbf{w}_{g,p,m} \circ \mathbf{h}_{g,p,m} \right) \mathbf{c}_g^H}_{\text{desired signal}} \\ &+ \underbrace{\sum_{k=1, k \neq g}^{K} d_{k,p} \cdot \mathbf{c}_k \sum_{m=1}^{M} \left(\mathbf{w}_{k,p,m} \circ \mathbf{h}_{g,p,m} \right) \mathbf{c}_g^H}_{\text{MAI}} + \underbrace{\mathbf{n}_r}_{\text{noise}} . \end{aligned} \tag{4}$$

For receiver (b), the decision variable expression is very similar to (4), but in this case the vector $\mathbf{h}_{g,p,m}$ is replaced by

$$\mathbf{z}_{g,p,m} = \frac{\left(\left| \mathbf{h}_{g,p,m} \right|^2 + \mathbf{h}_{g,p,m} \sum_{i=1, i \neq m}^{M} \mathbf{h}_{g,p,l}^* \right)}{\sum_{m=1}^{M} \left| \mathbf{h}_{g,p,m} \right|}, \tag{5}$$

where $(\cdot)^*$ denotes the complex conjugate.

The vector $\mathbf{n}_r$ represents the residual noise samples of ML subcarriers. The signal of (4) involves the three terms: the desired signal, the MAI caused by the loss of code orthogonality among the users, and the residual noise after despreading.

3. TRANSMIT PREEQUALIZATION SCHEMES

In this section we analytically derive a space-frequency preequalization algorithm for the two receiver configurations: (a) and (b). In the latter case, the weights are computed taking into account that at the receiver we have the EGC combiner. However, we use the same criterion, zero forcing, in both preequalization schemes.

The use of preequalization algorithms has two main advantages: reducing the MAI at MTs by preformatting the signal so that the received signal at the decision point is free from interferences and allowing moving the most computational burden from MT to BS, keeping the MT at a low complexity level. When we use an antenna array at the BS, the preequalization can be done in both dimensions, space and frequency. We propose to jointly optimize the user separation in space and frequency by the use of criteria based on the decision variable after despreading at the MT. This optimization task is performed taking into account the power minimization at the transmitter side.

3.1. CZF preequalization algorithm

The CZF preequalization algorithm is based on a zero-forcing criterion, since we put a zero in the MAI term. This algorithm is designed in order to remove the MAI term of (4) at all MTs at same time. Furthermore, it takes into account the transmitted power at BS, the reason we call this algorithm the CZF.

Applying the zero-forcing criterion to (4), we obtain the following conditions:

$$\begin{aligned} \mathbf{c}_g \circ \left(\sum_{m=1}^{M} \mathbf{w}_{g,p,m} \circ \mathbf{h}_{g,p,m} \right) \mathbf{c}_g^H &= 1, \\ \sum_{k=1, k \neq g}^{K} \mathbf{c}_k \circ \left(\sum_{m=1}^{M} \mathbf{w}_{k,p,m} \circ \mathbf{h}_{g,p,m} \right) \mathbf{c}_g^H &= 0, \end{aligned} \tag{6}$$

ensuring that each user receives a signal that after despreading is free of MAI. The first term of the right-hand side of (4) is the desired signal, and has been made, for normalization purposes, equal to 1, while the second term represents the interference caused by other $K-1$ users, and according to the criterion used should be equal to 0.

The interference that the signal of a given user g produces at another MT k is obtained for a generic data symbol according to (4):

$$\begin{aligned} \text{MAI}(g \longrightarrow k) &= \mathbf{c}_g \left(\sum_{m=1}^{M} \mathbf{w}_{g,p,m} \circ \mathbf{h}_{k,p,m} \right) \mathbf{c}_k^H \\ &= \mathbf{v}_{k,g} \circ \mathbf{w}_{g,p} \end{aligned} \tag{7}$$

with $\mathbf{v}_{k,g} = \widehat{\mathbf{c}}_g \circ [\mathbf{h}_{k,p,1} \ \ \mathbf{h}_{k,p,2} \ \ \cdots \ \ \mathbf{h}_{k,p,M}] \circ \widehat{\mathbf{c}}_k^H$.

The weight vector for user g is then obtained by constraining the desired signal part of its own decision variable to 1, while cancelling its MAI contribution all other MTs at same time. This leads to the following set of conditions:

$$\begin{aligned} \mathbf{c}_g \circ \left(\sum_{m=1}^{M} \mathbf{w}_{g,p,m} \circ \mathbf{h}_{g,p,m} \right) \mathbf{c}_g^H &= 1, \\ \mathbf{c}_g \circ \left(\sum_{m=1}^{M} \mathbf{w}_{g,p,m} \circ \mathbf{h}_{k,p,m} \right) \mathbf{c}_k^H &= 0 \quad \forall k \neq g. \end{aligned} \tag{8}$$

Therefore, to compute the weights for user g, we have to solve a linear system of K EQUATIONS (constraints) and

$L \cdot M$ variables (degrees of freedom) given by

$$A_{g,p}\mathbf{w}_{g,p} = \mathbf{b}, \tag{9}$$

where $A_{g,p}$ is a matrix of size $K \times ML$, given by

$$A_{g,p} = \begin{bmatrix} \begin{bmatrix} \mathbf{h}_{g,p,1} & \mathbf{h}_{g,p,2} & \cdots & \mathbf{h}_{g,p,M} \end{bmatrix} \\ \mathbf{v}_{0,g}^H \\ \vdots \\ \mathbf{v}_{g-1,g}^H \\ \mathbf{v}_{g+1,g}^H \\ \vdots \\ \mathbf{v}_{K-1,g}^H \end{bmatrix}, \qquad \mathbf{b} = \begin{bmatrix} 1 \\ 0 \\ \vdots \\ 0 \end{bmatrix}. \tag{10}$$

As pointed above the prefiltering algorithms should take into account the minimization of the transmitted power. Therefore, the transmitted power must be minimized under K above constraints. When the number of constraints equals the number of degrees of freedom, a single solution exists provided there are no singularities. If however we have more degrees of freedom than constraints ($ML > K$), then signal design can be done to optimize some cost function, normally, the total transmitted power. The optimization will be more effective the higher $ML - K$ is. This optimization can be solved with the Lagrange multipliers method [10].

The transmitter power $\mathbf{p}_t$ is given by,

$$\mathbf{p}_t = \mathbf{w}_{g,p}^H \mathbf{w}_{g,p} \tag{11}$$

and consequently we need to minimize the following cost function with Lagrange multiplier α,

$$\mathbf{j} = \mathbf{w}_{g,p}^H \mathbf{w}_{g,p} - \alpha A_{g,p}\mathbf{w}_{g,p}, \tag{12}$$

computing the gradient vector of j and equating to zero, we get

$$\nabla_w \mathbf{j} = 2 \cdot \mathbf{w}_{g,p}^H - \alpha A_{g,p} = 0 \tag{13}$$

thus the vector weight is given by

$$\mathbf{w}_{g,p} = \frac{1}{2} A_{g,p}^H \alpha^H, \tag{14}$$

and then using this result in (9), we get

$$\alpha^H = 2 \cdot \left(A_{g,p}A_{g,p}^H\right)^{-1} \cdot \mathbf{b}. \tag{15}$$

Finally, replacing α^H in (14), we obtain the CZF-based prefiltering vector given by

$$\mathbf{w}_{g,p} = A_{g,p}^H \left(A_{g,p}A_{g,p}^H\right)^{-1} \mathbf{b} = A_{g,p}^H \psi_{g,p}^{-1} \mathbf{b}, \tag{16}$$

where $\psi_{g,p} = [A_{g,p}A_{g,p}^H]$ is a square and Hermitian matrix of size $K \times K$.

Observing the last equation, it easy to see that the most computational intensive task, to calculate the weights, comes from matrix $\Psi_{g,p}$ inversion. However, the size of this matrix is just $K \times K$ independently of the spreading factor and the number of antennas, which makes this algorithm very attractive for practical implementations.

3.2. *CZF-EGC preequalization algorithm*

As referred, for this scheme we also use the zero-forcing criterion, but now to compute the weights, we should take into account that we have the EGC combiner at receiver side, the reason we call this algorithm CZF-EGC. When we use the EGC at receiver side, we increase the complexity as compared with receiver (a). However, the complexity level is still perfectly tolerable for practical implementations. EGC requires only knowledge about the channel phase.

The interference that the signal of a given user g produces at another MT k is obtained for a generic data symbol according to (4) and (5) by

$$\begin{aligned} &\mathrm{MAI}(g \longrightarrow k) \\ &\quad = \mathbf{c}_g \left(\sum_{m=1}^{M} \mathbf{w}_{g,p,m} \circ \frac{\left|\mathbf{h}_{k,p,m}\right|^2 + \mathbf{h}_{k,p,m} \circ \sum_{i=1, i\neq m}^{M} \mathbf{h}_{k,p,l}^*}{\sum_{m=1}^{M} \left|\mathbf{h}_{k,p,m}\right|} \right) \mathbf{c}_k^H \\ &\quad = \mathbf{s}_{k,g} \circ \mathbf{w}_{g,p}, \end{aligned} \tag{17}$$

with

$$\begin{aligned} &\mathbf{s}_{k,g} \\ &\quad = \hat{\mathbf{c}}_g \circ \left[\left(\frac{\left|\mathbf{h}_{k,p,1}\right|^2 + \mathbf{h}_{k,p,1} \circ \left(\mathbf{h}_{k,p,2}^* + \cdots + \mathbf{h}_{k,p,M}^*\right)}{\left|\mathbf{h}_{k,p,1}\right| + \cdots + \left|\mathbf{h}_{k,p,M}\right|} \right) \cdots \right. \\ &\qquad \left. \left(\frac{\left|\mathbf{h}_{k,p,M}\right|^2 + \mathbf{h}_{k,p,M} \circ \left(\mathbf{h}_{k,p,1}^* + \cdots + \mathbf{h}_{k,p,M-1}^*\right)}{\left|\mathbf{h}_{k,p,1}\right| + \cdots + \left|\mathbf{h}_{k,p,M}\right|} \right) \right] \\ &\quad \circ \hat{\mathbf{c}}_k^H. \end{aligned} \tag{18}$$

The weight vector for user g is also obtained by constraining the desired signal part of its own decision variable to one while cancelling its MAI contribution to all other MTs at same time. This leads to the following set of conditions:

$$\begin{aligned} &\mathbf{c}_g \left(\sum_{m=1}^{M} \mathbf{w}_{g,pm} \circ \frac{\left(\left|\mathbf{h}_{g,p,m}\right|^2 + \mathbf{h}_{g,p,m} \circ \sum_{i=1, i\neq m}^{M} \mathbf{h}_{g,p,i}^*\right)}{\sum_{m=1}^{M} \left|\mathbf{h}_{g,p,m}\right|} \right) \mathbf{c}_g^H = 1, \\ &\mathbf{c}_g \left(\sum_{m=1}^{M} \mathbf{w}_{g,p,m} \circ \frac{\left(\left|\mathbf{h}_{k,p,m}\right|^2 + \mathbf{h}_{k,p,m} \cdot \sum_{i=1, i\neq m}^{M} \mathbf{h}_{k,p,i}^*\right)}{\sum_{m=1}^{M} \left|\mathbf{h}_{k,p,m}\right|} \right) \mathbf{c}_k^H \\ &\qquad = 0 \quad \forall g \neq k. \end{aligned} \tag{19}$$

As the CZF algorithm, the CZF-EGC-based pre-filtering vector is given by (16). However, the matrix $A_{g,p}$ is now given by

$$A_{g,p}=\begin{bmatrix} \left[\left(\dfrac{|\mathbf{h}_{g,p,1}|^2+\mathbf{h}_{g,p,1}\circ(\mathbf{h}^*_{g,p,2}+\cdots+\mathbf{h}^*_{g,p,M})}{|\mathbf{h}_{g,p,1}|+\cdots+|\mathbf{h}_{g,p,M}|}\right)\cdots \right. \\ \left. \left(\dfrac{|\mathbf{h}_{g,p,M}|^2+\mathbf{h}_{g,p,M}\circ(\mathbf{h}^*_{g,p,1}+\cdots+\mathbf{h}^*_{g,p,M-1})}{|\mathbf{h}_{g,p,1}|+\cdots+|\mathbf{h}_{g,p,M}|}\right)\right] \\ s^H_{0,g} \\ \vdots \\ s^H_{g-1,g} \\ s^H_{g+1,g} \\ \vdots \\ s^H_{K-1,g} \end{bmatrix}. \tag{20}$$

From (19) we can see that for the case $M = 1$ (single antenna), we obtain an expression very similar to (8) for a single-antenna case, given by

$$\begin{gathered} \mathbf{c}_g \cdot \mathbf{w}_{g,p} \circ |\mathbf{h}_{g,p}| \cdot \mathbf{c}^H_g = 1, \\ \mathbf{c}_g \cdot \mathbf{w}_{g,p} \circ |\mathbf{h}_{k,p}| \cdot \mathbf{c}^H_k = 0 \quad \forall k \neq g. \end{gathered} \tag{21}$$

In this case, the weights are real, because we use the modulus of the channel frequency response; thus we just equalize the amplitude at the transmitter whereas the phase is equalized at the receiver side.

4. NUMERICAL RESULTS

To evaluate the performance of the proposed preequalization algorithms, we used a pedestrian Rayleigh fading channel, whose system parameters are derived from the European BRAN Hiperlan/2 standardization project [11]. This channel model has 18 taps, multipath spread of 1.76 μs and coherence bandwidth approximately equal to 637 KHz.

We extended this time model to a space model in two different ways: for the diversity case, we assumed that the distance between antenna elements is large enough to consider for each user M independent channels, that is, we assume independent fading processes; for the beamforming case, we allocated a direction of arrival (DOA) to each path (beamforming), with the DOAs randomly chosen within a 120° sector. In this latter case, the BS is equipped with a half-wavelength-spaced uniform linear array. We considered a DL synchronized transmission using Walsh-Hadamard spreading sequences of length 32 scrambled by a pseudorandom code. We used 1024 carriers, a bandwidth equal to 100 MHz, and a carrier frequency equal 5.0 GHz. The duration of the GP is 20% of the total OFDM symbol duration. The channel is considered to be constant during an OFDM symbol.

The simulations were carried out to assess the performance of the CZF and CZF-EGC algorithms in the two different scenarios presented above, and to compare against the performance achieved with conventional frequency equalization receivers, such as MRC, EGC, and MMSE. For a better

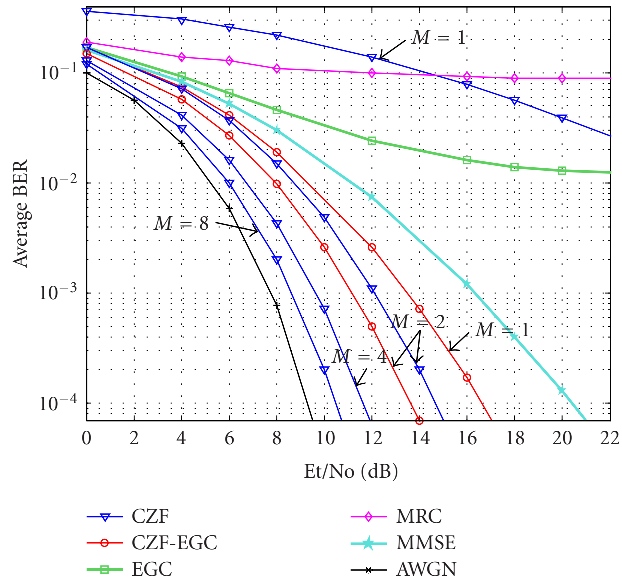

Figure 2: Performance comparison between the CZF, CZF-EGC, and conventional receivers as function of Et/No, for diversity case.

comparison with a variable number of antennas, the results have been normalized, that is, for the case of multiple transmitting antennas, the figures do not take into account the array gain which is $10\,\mathrm{Log}(M)$ in dB.

The simulation results for the diversity case are shown in Figures 2 and 3. The simulations of Figure 2 were run for a number of users $K = 32$, that is, a full-load system, and the metric used is the average bit error rate (BER) as function of Et/No, the transmitted energy (assuming a normalized channel) per bit over the noise spectral density. The performance of the CZF and CZF-EGC algorithms is illustrated for the cases of $M = 1, 2, 4$, and 8 transmit antennas. With a single antenna at the BS, there is no spatial separation and the preequalization operation is done only in the frequency dimension. As it can be seen from Figure 2, the performance of the CZF algorithm for a single antenna is modest. At low values of Et/No, the performance is even worse than with all single-user conventional detectors, and only for high values of Et/No the CZF outperforms the conventional MRC equalizer. This occurs because for a single antenna and full load system, we do not have enough degrees of freedom to minimize the transmitted power. The number of degrees of freedom is equal to $M \cdot L$ and the number of constraints is K. Thus, for $M = 1$ and $K = L$ (full-load system), the number of degrees is equal to the number of constraints. For multiple antennas at BS, it is possible to optimize the preequalization algorithm in both dimensions, space and frequency. When we use an array of 2, 4, and 8 antennas, the performance of the CZF algorithm is much better than all single-user conventional equalizers for any Et/No value. We can see that with 4 and 8 antennas, the performance is very close to the one obtained with the Gaussian channel. As it can be seen from Figure 2, the performance of the CZF-EGC algorithm for single antenna outperforms the MRC, EGC, MMSE

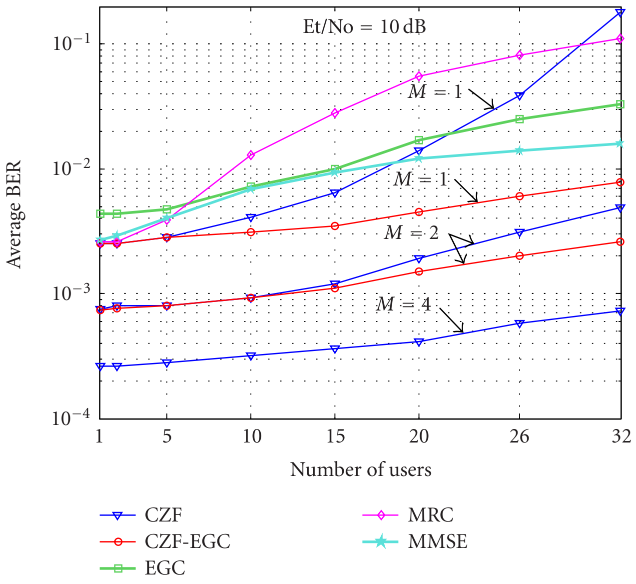

FIGURE 3: Performance comparison between the CZF, CZF-EGC and conventional receivers as function of number of users, for diversity case.

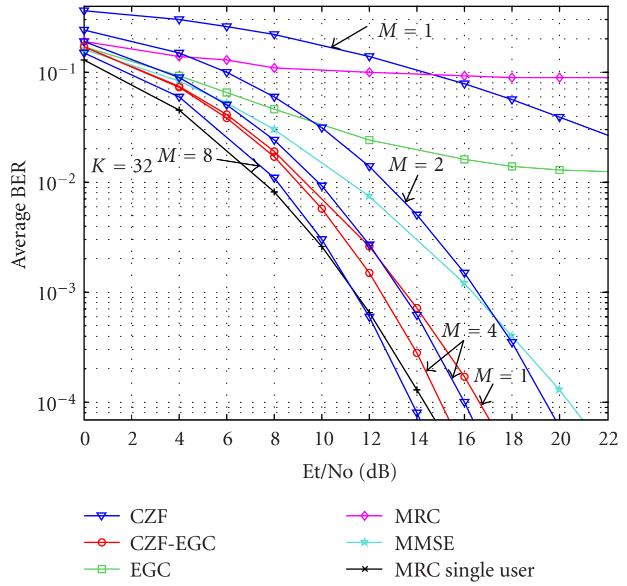

FIGURE 4: Performance comparison between the CZF, CZF-EGC and conventional receivers as function of Et/No, for beamforming case.

conventional equalizers, and the CZF. This occurs because, as can be seen from (21), with single antennas the CZF-EGC weights are real. Thus, we just perform a preequalization amplitude operation at the transmission side while the phase equalization is done at the receiver. In the case of CZF, we perform the amplitude and phase equalization at transmission side. For two antennas, the performance of the CZF-EGC is slightly better than the one of the CZF algorithm, while for a number of antenna elements greater than four, the performance of both algorithms is nearly identical.

The simulations leading to Figure 3 were run for Et/No = 10 dB, and the metric used is the average BER as a function of number of the users and considers the cases of a single-, two-, and four-antenna elements. From this figure we can see that for a single antenna, the CZF algorithm performance outperforms the one obtained with conventional receiver equalizers for number of users up to 16 (half load).

Beyond this point the performance degradation rapidly increases with the number of users. This arises because for a single antenna the difference between the number of degrees of freedom and constraints tends to zero as the number of users increases. When the number of antennas increases, the performance of the CZF improves, because we increase the number of degrees of freedom keeping the number of constraints. From Figure 3 we can see that for $M = 2$ and 4, and the CZF gives better results than all conventional equalizers, even for full load system. Concerning CZF-EGC algorithm, we can see that performance is much better that all conventional receiver equalizers even for single antenna. When we increase the number of antenna elements, the performance of the CZF-EGC tends to the CZF performance, even for full-load system.

The simulation results for the beamforming case are shown in Figure 4, where the simulation metrics and parameters other than the channels correlation are identical to the ones considered for Figure 2. The results show that the CZF algorithm outperforms all the conventional equalizers, except for the single-antenna case (with loads close to full) as happened with the diversity scenario. From this figure we can see that the performance of the CZF with two antennas is worse than the conventional MMSE combiner. With eight-antenna elements, the CZF performance is very close to the performance of the MRC single user. The performance of the CZF-EGC with single antenna is very good as compared with all conventional equalizers and CZF. For the single-antenna case we have the same number of degrees of freedom for both algorithms, but for CZF-EGC case we just perform the amplitude equalization at transmitter side, while for CZF we perform amplitude and phase equalization. We can even see that the performance is similar to the one obtained with CZF algorithm for four antennas. The performance of the CZF-EGC for four and eight antennas is very close to the one obtained with CZF.

5. CONCLUSIONS

We proposed space-frequency preequalization techniques for downlink TDD MC-CDMA, using antenna arrays at BS, for two different receivers: the conventional EGC and a simple despread receiver without channel equalization. We analytically derived the proposed preequalization algorithms, based on a constrained zero-forcing criterion. The performance was assessed for either of the diversity and beamforming scenarios and compared against one of conventional receivers.

The results have shown that a considerable MAI reduction is obtained with the CZF-EGC technique, and with CZF when an antenna array is used at BS. For a single antenna, the performance of the CZF-EGC outperforms the CZF for a full-load case, while with multiple antennas the performances are very similar. Both techniques allow a significant improvement of the user capacity and move the most-demanded processing task from BS to MT, keeping this one as simple as possible.

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Preliminary call for papers

The 2011 European Signal Processing Conference (EUSIPCO-2011) is the nineteenth in a series of conferences promoted by the European Association for Signal Processing (EURASIP, www.eurasip.org). This year edition will take place in Barcelona, capital city of Catalonia (Spain), and will be jointly organized by the Centre Tecnològic de Telecomunicacions de Catalunya (CTTC) and the Universitat Politècnica de Catalunya (UPC).

EUSIPCO-2011 will focus on key aspects of signal processing theory and applications as listed below. Acceptance of submissions will be based on quality, relevance and originality. Accepted papers will be published in the EUSIPCO proceedings and presented during the conference. Paper submissions, proposals for tutorials and proposals for special sessions are invited in, but not limited to, the following areas of interest.

Areas of Interest

- Audio and electro-acoustics.
- Design, implementation, and applications of signal processing systems.
- Multimedia signal processing and coding.
- Image and multidimensional signal processing.
- Signal detection and estimation.
- Sensor array and multi-channel signal processing.
- Sensor fusion in networked systems.
- Signal processing for communications.
- Medical imaging and image analysis.
- Non-stationary, non-linear and non-Gaussian signal processing.

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Procedures to submit a paper and proposals for special sessions and tutorials will be detailed at www.eusipco2011.org. Submitted papers must be camera-ready, no more than 5 pages long, and conforming to the standard specified on the EUSIPCO 2011 web site. First authors who are registered students can participate in the best student paper competition.

Important Deadlines:

Proposals for special sessions	15 Dec 2010
Proposals for tutorials	18 Feb 2011
Electronic submission of full papers	**21 Feb 2011**
Notification of acceptance	23 May 2011
Submission of camera-ready papers	6 Jun 2011

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