Downlink Strategies for MISO TDD MC-CDMA Systems

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***Abstract*—This paper propose two advanced space-frequency schemes with a pre-filtering technique for downlink time division duplex (TDD) MC-CDMA system. We consider the use of antenna arrays at the base station (BS) and a single antenna at the mobile terminal (MT). We derive a multi-user pre-filtering technique that allows to format the transmitted signals so that the multiple access interference at mobile terminals is completely removed and eliminating the channel deep fades, without enhancing the noise power and keeping the MT at low complexity. This pre-filtering technique is combined with the MC-CDMA and 2x1 SFBC MC-CDMA systems. Simulations results are carried out in scenarios with spectral efficiency equal to 3 bps/Hz and using the UMTS channel turbo code, to demonstrate the effectiveness of the proposed advanced schemes.**

I. Introduction

As demand for wireless services increases, more capacity is needed, and since the spectrum is a very scarce resource, the development of efficient techniques regarding the usage of this resource are mandatory. The beyond 3G broadband component of wireless system must be able to offer bit rates more than 100Mbps in indoor and picocell. MC-CDMA is one the most promising multiple access scheme, especially in downlink, for achieving such high data rates, in order to meet the quality of service requirements of the future multimedia application [1]. 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 multi-path propagation, while CDMA allows a flexible multiple access with good interference properties for cellular environments [2]. However, the user capacity of MC-CDMA system is essentially limited by the Multiple Access Interference (MAI). MC-CDMA is for example studied within the European IST-4MORE project [3].

It is consensual that provision of the broadband wireless component will probably rely on the use of multiple antennas at transmitter/receiver side. For the DL using multiple antennas at the BS is more feasible than at the MT. Thus, transmit diversity schemes relying on multiple transmit antennas are very attractive. Space-time coding schemes, such as space-time block coding (STBC), relying on multiple antenna at transmitter side and appropriate signal processing at the receiver were proposed [4][5]. STBC is a simple diversity scheme which can improve the performance, and for the particular case of two transmit antennas this can be achieved without any rate loss. An MC-CDMA system can be designed in a way that two or more adjacent narrowband subchannels are affected by nearly the same channel coefficients and the combined symbols can be sent on different subcarriers. Unlike the STBC, in this case the chips are encoded in space and frequency domains and this diversity scheme is known by space-frequency block coding (SFBC). In [6] an efficient realization of SFBC for OFDM system was introduced and its combination with single user pre-equalizers techniques in UL of MC-CDMA systems. Considering TDD, another solution consists in performing pre-filtering at transmitter side using the TDD channel reciprocity between alternative uplink and downlink transmission periods [7]. Normally this principle is valid for indoor or pedestrian environments, i.e., in low mobility scenarios. The aim of this solution is to allow the use of simple low-cost, low-consuming MT.

Unlike the BS, where complexity requirements are not so tight, low complexity is required at MTs, thus only simple and single user equalizers techniques can be implemented, limiting the MAI cancellation capability. Motivated by this constraint at the MT, this paper proposes two advanced space-frequency schemes with a pre-filtering technique for the downlink of MC-CDMA systems. For the first scheme a space-frequency pre-filtering algorithm is derived and for second one this algorithm is combined with 2x1 SFBC MC-CDMA system. The multi-user pre-filtering technique allows formatting the transmitted signal so that the MAI at mobile is completely removed and eliminating the channel deep fades, without enhancing the noise power, keeping MT at very low complexity. The pre-filtering processing is designed using as criterion the minimization of the transmitted power at the BS.

The paper is organized as follows: In section II we present the proposed advanced downlink MC-CDMA schemes. In section III, we analytically derive the multi-user pre-filtering algorithm which we call multi-user constrained zero forcing (MCZF) for first scheme and *J*-MCZF for second one. In section IV, we present some simulation results obtained with the proposed advance schemes for a spectral efficiency of 3bps/Hz in order to assess the pre-filtering algorithm in a high data rate context. We also compare the propose schemes against conventional 2x1 SFBC schemes using single user equalizer techniques such as, Zero Forcing Combining (ZFC) and Minimum Mean Square Error Combining (MMSEC). Finally the main conclusions are pointed out in section V.

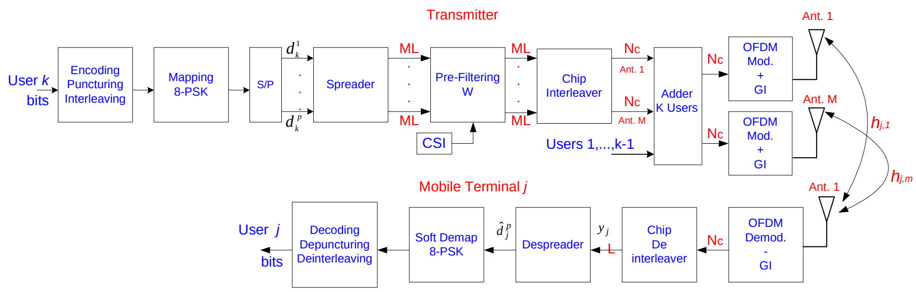

Figure 1: Downlink Space-Frequency Pre-Filtering MC-CDMA Scheme – a).

II. SYSTEM MODELS

In this section we present two different downlink schemes: MC-CDMA with a space-frequency pre-filtering algorithm and 2X1 SFBC MC-CDMA combined with the same pre-filtering technique.

A. MC-CDMA with a Space-Frequency Pre-Filtering Algorithm

Figure 1 shows the first proposed downlink MC-CDMA transmitter for user k and the receiver of the j^{th} MT. After channel encoding, puncturing and interleaving operations, the bit streams are mapped to an 8-PSK constellation. Each user k transmits $P=N_c/L$, (where N_c is the number of carriers and L the length of the spreading code) data symbols per OFDM symbol. Then the data symbols are spread into L chips using the orthogonal Walsh-Hadamard code set. We denote the code vector of user k as $\boldsymbol{c}_k = \left[c_{k,1},...,c_{k,q},...,c_{k,L}\right]^T$, where $c_{k,q}$ is the q^{th} chip and $(.)^T$ denotes the transpose operator. We propose to perform pre-filtering jointly in space and frequency before OFDM modulation. Therefore, the chips of the data symbols are copied M times in order to obtain $M{\cdot}L$ versions of the original symbols which are weighted and transmitted over the M antenna branches. Thus, each user's symbol is affected by a specific weight on each subcarrier and antenna branch. Mathematically, we represent the spreading and copying process by a single vector containing M times the code vector of the corresponding user, *i.e.* $\bar{\boldsymbol{c}}_k = \left[c_k^T,...,c_k^T\right]^T$. The pre-filtering weights of user k and a generic symbol p are also gathered in a vector of length $M{\cdot}L$ denoted by $\boldsymbol{w}_k^p = \left[\boldsymbol{w}_{k,1}^{pT} \quad \boldsymbol{w}_{k,2}^{pT} \quad \cdots \quad \boldsymbol{w}_{k,M-1}^{pT}\right]^T$ where $\boldsymbol{w}_{k,m}^{pT}$ is a vector of size L that contains the set of coefficients that weight the chips that go to antenna m. These weights are calculated on the basis of channel state information (CSI) according to the criteria presented in section III. Then the chips are interleaved in order to transmit them on distant positions in the OFDM frame. We consider frequency non-selective Rayleigh fading per subcarrier. However, due to the frequency interleaving operation, each data symbol experiences L uncorrelated frequency complex channel fading coefficients increasing the frequency diversity gain. After that, the signals of all users on each sub-carrier and antenna branch are added to form the muti-user transmitted signal. Finally, a guard interval (GI) is inserted to avoid ISI interference.

At the MT, the single antenna implicitly recombines the signals transmitted from the M branches at the BS. Contrary, to the conventional MC-CDMA receiver where single user or multi-user equalizers are performed, in the proposed scheme we can reduce this equalization operation to a simple despreading operation. Thereby, we can avoid channel estimation and equalization on each subcarrier at the MT. This results in a very low complex receiver design. Thus, the decision variable at the input of the 8-PSK demodulator, is for the desired user j and symbol p given by,

$$\hat{d}_j^p = \underbrace{\bar{\boldsymbol{c}}_j^H.(\boldsymbol{h}_j^p \circ \boldsymbol{w}_j^p \circ \bar{\boldsymbol{c}}_j)d_j^p}_{Desired\ Signal} + \underbrace{\sum_{k=1,k\neq j}^{K} \bar{\boldsymbol{c}}_j^H(\boldsymbol{h}_j^p \circ \boldsymbol{w}_k^p \circ \bar{\boldsymbol{c}}_k)d_k^p}_{MAI} + \underbrace{\boldsymbol{c}_j^H n_j}_{Noise} \qquad (1)$$

where $\boldsymbol{h}_k^p = \left[\boldsymbol{h}_{k,1}^{pT} \quad \boldsymbol{h}_{k,2}^{pT} \quad \cdots \quad \boldsymbol{h}_{k,M}^{pT}\right]^T$ of size $M.L$, where $\boldsymbol{h}_{k,m}^{pT}$ is the channel frequency response between antenna m and mobile terminal j for user k and symbol p, (o) means an element wise vector product. The signal of (1) involves the three main terms: the desired signal, the MAI caused by the loss of code orthogonality among the users, and the residual noise after despreading.

B. 2x1 SFBC MC-CDMA with a Pre-filtering Algorithm

Figure 2 shows the second proposed downlink MC-CDMA transmitter for user k and the receiver of the j^{th} MT. As can be seen up to spreading operation we perform the same operations as in the first scheme. After that each spreaded data symbol is weighted by a vector $\boldsymbol{g}_k^p = \left[g_{k,1}^p,...,g_{k,q}^p,...,g_{k,L}^p\right]^T$ of size L, where $g_{k,q}^p$ is the q^{th} weight chip of the k^{th} user and symbol p. We compute a vector weight for each data symbol of size L, instead $M.L$ as in the first scheme. After the pre-filtering operation the signal for a generic user k is given by,

$$\boldsymbol{s}_k = \left[\boldsymbol{s}_k^{1T},...,\boldsymbol{s}_k^{pT}\right] = \left[s_{k,1},...,s_{k,n},s_{k,n+1},...,s_{k,N_C}\right] \qquad (2)$$

of size N_c , where $\boldsymbol{s}_k^p$ is of size L and is given by,

$$\boldsymbol{s}_k^p = d_k^p \boldsymbol{c}_k \circ \boldsymbol{g}_k^p = \left[s_{k,1}^p,...,s_{k,q}^p,...,s_{k,L}^p\right] \qquad (3)$$

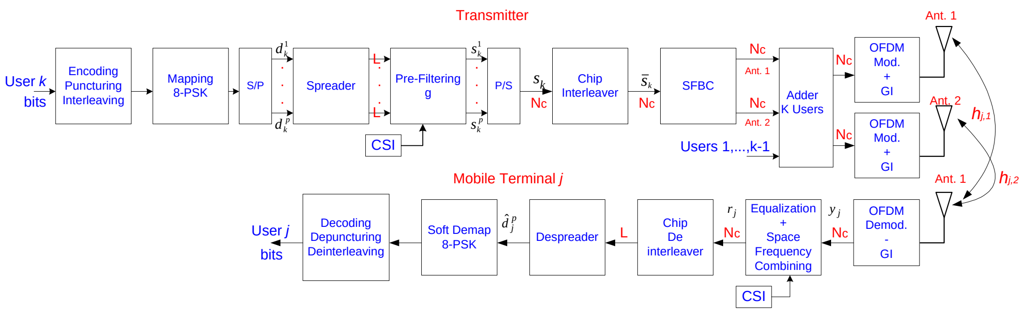

Figure 2: Downlink 2x1 SFBC MC-CDMA with a pre-filtering Scheme – b).

Then the chips of the sequence $\boldsymbol{s_k}$ are interleaved in frequency domain to produce the sequence $\bar{\boldsymbol{s}}_{\boldsymbol{k}}$. In this new sequence the chips of each data symbol are separated by L positions in the OFDM symbol. The mapping scheme of the chips of the sequence $\bar{\boldsymbol{s}}_{\boldsymbol{k}}$ for SFBC with two transmit antennas is shown in Table 1.

Table 1: SFBC mapping [6] for 2 transmit antennas.

	Antenna 1	Antenna 2
Subcarrier n	$\bar{s}_{\boldsymbol{k,n}}$	$-\bar{s}^*_{\boldsymbol{k,n+1}}$
Subcarrier n+1	$\bar{s}_{\boldsymbol{k,n+1}}$	$\bar{s}^*_{\boldsymbol{k,n}}$

After that, the signals of all users on each sub-carrier and antenna branch are added to form the multi-user transmitted signal. Finally, a guard interval (GI) is inserted to avoid ISI interference. The received signals on sub-channels n and $n+1$ after OFDM demodulation and GI removal are given by,

$$\begin{cases} y_{j,n} = \sum_{k=1}^{K} \bar{s}_{k,n} h_{j,1,n} - \sum_{k=1}^{K} \bar{s}^*_{k,n+1} h_{j,2,n} + n_n \\ y_{j,n+1} = \sum_{k=1}^{K} \bar{s}_{k,n+1} h_{j,1,n+1} + \sum_{k=1}^{K} \bar{s}^*_{k,n} h_{j,2,n+1} + n_{n+1} \end{cases} \tag{4}$$

where $h_{j,m,n}$ represents the frequency complex fading channel for mobile j, antenna m and subchannel n; $\boldsymbol{n_n}$ is the additive white Gaussian noise on subcarrier n. OFDM systems are usually designed so that the subcarrier separation is significantly lower than the coherence bandwidth of the channel, and therefore, the fading in two adjacent subcarriers can be considered flat i.e. we can consider $h_{j,m,n}$ to be equal to $h_{j,m,n+1}$.

At the MT we propose a very simple single user equalizer, with the coefficients given by:

$$z_{j,m,n} = h_{j,m,n}^{\ *} / \sum_{m=1}^{2} |h_{j,m,n}| \tag{5}$$

These equalization coefficients are slightly different from the ones used by an EGC equalizer. The reason to use the equalization coefficients of (5) instead of the ones given by EGC, is due to the fact that with the coefficients of (5) we do not get intersymbol interference in the SFBC decoding process contrarily to what would arise using EGC coefficients..

The signal for on arbitrary pair of adjacent sub-carriers n and $n+1$, after using the SF combining scheme and the equalization coefficients defined in (5), can be written by,

$$\begin{cases} r_{j,n} = z_{j,1,n} y_n + z^*_{j,2,n+1} y^*_{n+1} \\ r_{j,n+1} = -z^*_{j,2,n} y^*_n + z_{j,1,n+1} y_{n+1} \end{cases} \tag{6}$$

After some mathematical manipulations and considering that $z_{j,m,n} = z_{j,m,n+1}$, we can write,

$$\begin{cases} r_{j,n} = \dfrac{|h_{j,1,n}|^2 + |h_{j,2,n}|^2}{|h_{j1,n}| + |h_{j,2,n}|} \bar{s}_{j,n} + MAI_n + z_{j,1,n} n_n + z^*_{j,2,n} n^*_{n+1} \\ r_{j,n+1} = \dfrac{|h_{j,1,n}|^2 + |h_{j,2,n}|^2}{|h_{j,1,n}| + |h_{j,2,n}|} \bar{s}_{j,n+1} + MAI_{n+1} - z^*_{j,2,n} n^*_n + z_{j,1,n} n_{n+1} \end{cases} \tag{7}$$

After chip de-interleaving and despreading operations, we obtain the decision variable for a generic symbol p and MT j,

$$\hat{d}^p_j = \underbrace{\boldsymbol{c}^H_j (\boldsymbol{f}^p_j \circ \boldsymbol{g}^p_j \circ \boldsymbol{c}_j) d^p_j}_{DesiredSignal} + \underbrace{\sum_{k=1,k\neq j}^{K} \boldsymbol{c}^H_j (\boldsymbol{f}^p_j \circ \boldsymbol{g}^p_k \circ \boldsymbol{c}_k) d^p_k}_{MAI} + \underbrace{\boldsymbol{z}^T_{j,1} n_j + \boldsymbol{z}^H_{j,2} n^*_j}_{Noise} \tag{8}$$

where $\boldsymbol{f}^p_j$ is a vector of size L, and each elements given by,

$$\boldsymbol{f}^p_j = \left[\frac{\sum_{m=}^{2} |h^p_{j,m,1}|^2}{\sum_{m=}^{2} |h^p_{j,m,1}|}, \ldots, \frac{\sum_{m=}^{2} |h^p_{j,m,q}|^2}{\sum_{m=}^{2} |h^p_{j,m,q}|}, \ldots, \frac{\sum_{m=}^{2} |h^p_{j,m,L}|^2}{\sum_{m=}^{2} |h^p_{j,m,L}|} \right]^T \tag{9}$$

where $h^p_{g,m,n}$ represents the frequency complex fading channel for a generic symbol p, mobile g, antenna m and sub-channel n.

As in the first scheme the signal of (8) involves the same three well known terms. The pre-filtering algorithm proposed in the next section completely removes the MAI component and deep fades of (1) in the first scheme and of (8) for second one, respectively.

III. PRE-FILTERING ALGORITHM

The aim of pre-filtering at BS is to eliminate the MAI and deep fades at MTs, while allowing the use of a very low complex receiver at MT. Here, we derive the pre-filtering algorithm used in the above proposed MC-CDMA schemes. The algorithm is based on zero forcing criteria and is designed in space-frequency domain for the first scheme and only in frequency domain for second one in order to remove the MAI term of (1) and of (8) at all MTs, respectively. Furthermore, it takes into account the transmitted power at BS, reason why we call this algorithm the multi-user constrained zero-forcing (MCZF). In the second scheme we combine the 2x1 SFBC MC-CDMA with this pre-filtering algorithm, in this case to distinguish from first scheme we call joint MCZF (*J*-MCZF).

The interference that the signal of a given user j produces at an other MT k is from (1) given by

$$MAI(j \to k) = \bar{\boldsymbol{c}}_k^H (\boldsymbol{h}_k^p \circ \boldsymbol{w}_j^p \circ \bar{\boldsymbol{c}}_j) = \boldsymbol{v}_{k,j}^T \boldsymbol{w}_j^p \tag{10}$$

with $\boldsymbol{v}_{k,j} = \bar{\boldsymbol{c}}_k^* \circ \boldsymbol{h}_k^p \circ \bar{\boldsymbol{c}}_j$.

and from (8) given by,

$$MAI(j \to k) = \boldsymbol{c}_k^H (\boldsymbol{f}_k^p \circ \boldsymbol{g}_j^p \circ \boldsymbol{c}_j) = \boldsymbol{l}_{k,j}^T \boldsymbol{g}_j^p \tag{11}$$

with $\boldsymbol{l}_{k,j} = \boldsymbol{c}_k^* \circ \boldsymbol{f}_k^p \circ \boldsymbol{c}_j$.

The weight vector for MT j is then obtained by constraining the desired signal part of its own decision variable to a constant while cancelling its MAI contribution at all other mobile terminals at same time. This leads to the following set of conditions for first scheme,

$$\begin{cases} \bar{\boldsymbol{c}}_j^H \left(\boldsymbol{h}_j^p \circ \boldsymbol{w}_j^p \circ \bar{\boldsymbol{c}}_j\right) = \alpha \\ \bar{\boldsymbol{c}}_k^H \left(\boldsymbol{h}_k^p \circ \boldsymbol{w}_j^p \circ \bar{\boldsymbol{c}}_j\right) = 0 \qquad \forall k \neq j \end{cases} \tag{12}$$

and for second scheme,

$$\begin{cases} \boldsymbol{c}_j^H \left(\boldsymbol{f}_j^p \circ \boldsymbol{g}_j^p \circ \boldsymbol{c}_j\right) = \alpha \\ \boldsymbol{c}_k^H \left(\boldsymbol{f}_k^p \circ \boldsymbol{g}_j^p \circ \boldsymbol{c}_j\right) = 0 \qquad \forall k \neq j \end{cases} \tag{13}$$

Hence, to compute the weights for user j we have to solve a linear system of K equations, for both schemes given by,

$$A(\boldsymbol{g}_j^p \ or \ \boldsymbol{w}_j^p) = \boldsymbol{b} \tag{14}$$

where A is a matrix of size *KxML* for first scheme and of size *KxL* for second one and $\boldsymbol{b}$ a vector of size K, given by,

$$A = \begin{bmatrix} h_j^{pT} \\ v_{0,j}^T \\ \vdots \\ v_{j-1,j}^T \\ v_{j+1,j}^T \\ \vdots \\ v_{K-1,j}^T \end{bmatrix} \ or \ \begin{bmatrix} f_j^{pT} \\ l_{0,j}^T \\ \vdots \\ l_{g-1,j}^T \\ l_{j+1,j}^T \\ \vdots \\ l_{K-1,j}^T \end{bmatrix} \qquad \boldsymbol{b} = \begin{bmatrix} \alpha \\ 0 \\ \vdots \\ 0 \end{bmatrix} \tag{15}$$

As stated above the pre-filtering algorithms should take into account the minimization of the transmitted power. Here, the transmitted power must be minimized under the K constraints given by (12) or (13) depending of the scheme. This problem can be solved using the method of Lagrange multipliers [8].

After same mathematical manipulations, we obtain the MCZF and *J*-MCZF based pre-filtering vectors given by,

$$\boldsymbol{w}_j^p = \alpha A^H (A \ A^H)^{-1} \boldsymbol{b} = \alpha A^H \psi^{-1} \boldsymbol{b} \tag{16}$$

$$\boldsymbol{g}_j^p = \alpha A^H (A \ A^H)^{-1} \boldsymbol{b} = \alpha A^H \psi^{-1} \boldsymbol{b} \tag{17}$$

where $\psi = A \ A^H$ is a complex square and Hermitian matrix of size *KxK* for scheme one and a real square and Hermitian matrix of size *KxK*, α is a constant used to normalize the weight vectors according,

$$\left|\boldsymbol{w}_k^p\right|^2 = \boldsymbol{w}_k^{pH} \boldsymbol{w}_k^p = 1 \quad \forall k = 1...K \tag{18}$$

$$\left|\boldsymbol{g}_k^p\right|^2 = \boldsymbol{g}_k^{pH} \boldsymbol{g}_k^p = 1 \quad \forall k = 1...K \tag{19}$$

ensuring that the pre-filtering weight vectors always have unit norm in order to compare with systems without pre-equalization. It can be seen from (13) that for second scheme we only need to know the modulus of the channel frequency at the BS. Basically, in this scheme we just perform a pre-amplitude equalization in the frequency domain, while in the first scheme a phase and amplitude pre-filtering is performed in space and frequency domains.

IV. NUMERICAL RESULTS

To evaluate the performance of the proposed advanced MC-CDMA schemes with pre-filtering algorithm, we used an indoor Rayleigh fading channel, whose system parameters are derived from the European BRAN Hiperlan/2 standardization project [9]. We extended this time model to a space-time, assuming that the distance between antenna elements is large enough, to consider for each user M independents channels, i.e, we assume independent fading processes. The main parameters used in the simulations are presented in Table 2.

Table 2: Main simulation parameters.

Number of Carriers	1024
Spreading factor	32
Guard period samples/time	256 / 3.2μs
Number of users	32 (full load)
Total OFDM symbol duration (T_s)	16μs
Sampling frequency	80Mz
Frame duration	32*Ts=512μs
Modulation	8-PSK
Channel profile	BRAN A
Maximum delay/Number of Taps	390ns / 18

It is assumed that the receiver and transmitter have perfect knowledge of the channel. The transmitter power is normalized to one in all presented schemes. We compare the two advanced schemes proposed against Alamouti's SFBC using conventional single user equalizers (MMSEC and ZFC) at the MT. We present results for a spectral efficiency of η=3 bps/Hz in order to assess the proposed scheme in high data rate context scenario. The channel coding scheme is the turbo-code defined for UMTS, combined with a puncturing process and an interleaver to get an overall coding rate of 1/2. Figure 3 shows the performance results for the average bit error rate (BER) as function of Et/No. i.e, the transmitted energy (assuming a normalized channel) per bit over the noise spectral density, while Figure 4 shows the performance for the frame error rate (FER) as function of Et/No in order to assess the proposed algorithm in data and multimedia applications context.

Figure 3 shows the performance of the proposed MCZF and *J*-MCZF algorithms against 2x1 SFBC schemes employing two antennas and using single user ZFC and MMSEC equalization. As can be observed from this figure the performance of the 4x1 MCZF and 2x1 *J*-MCZF outperforms the 2x1 SFBC schemes using MMSE or ZFC. We also can observe that the performance of the 4X1 MCZF and of the 2x1 J-MCZF is similar. This means that for the first scheme we need more antennas to obtain the same performance as the second one. For a BER=1E-5, we get approximately 0.9dB and 2.5dB gain for 4x1 MCZF and 0.7dB and 2.1dB gain for 2x1 *J*-MCZF, when compared with 2x1 SFBC and using MMSEC and ZFC equalization, respectively. However it should be noted that when the MMSEC is used the noise variance must be estimated at MT therefore increasing the complexity. The two proposed advanced schemes outperform the 2x1 SFBC with MMSEC without needing to estimate the noise variance at MT and for the first scheme we also do not need estimate the frequency response at MT.

Figure 4 shows the performance of the proposed advanced schemes against 2x1 SFBC schemes for the same scenario of the Figure 3, using the average frame error rate (FER) as metric. In this case we also can see that the performance of the 4x1 MCZF and 2x1 *J*-MCZF outperforms all 2x1 SFBC schemes. The results presented in the above figures show that the advanced schemes are suitable for data and multimedia applications.

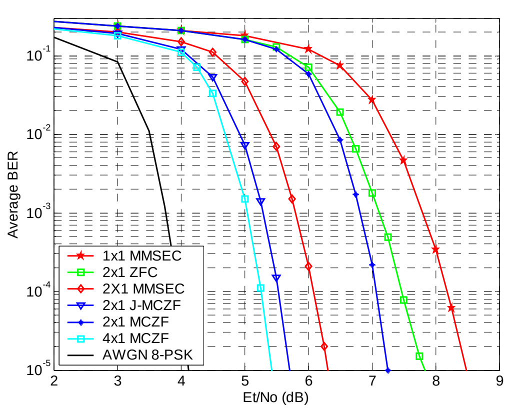

Figure 3**:** Schemes comparison for η=3bps/Hz, BER.

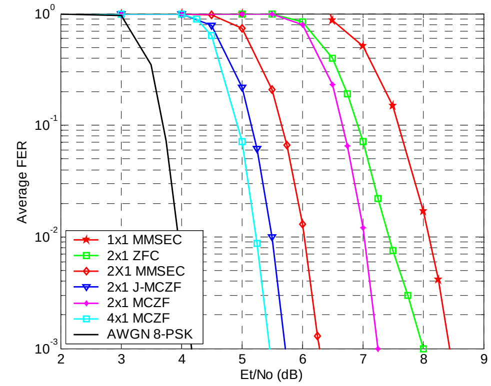

Figure 4: Schemes comparison for η=3bps/Hz, FER.

V. CONCLUSION

We proposed two advanced schemes with a pre-filtering algorithm for the downlink of TDD MC-CDMA systems, using antenna arrays at BS and a single antenna at mobile terminal. We analytically derived the proposed multi-user pre-filtering algorithms, based on a constrained zero-forcing criterion. The performance was assessed for scenarios with spectral efficiency equal to 3 bps/Hz considering the use of the turbo codes specified for UMTS. We compared the performance of the proposed MCZF and *J*-MCZF algorithms against 2x1 SFBC using MMSEC and ZFC single user equalization. The numerical results demonstrated that the proposed schemes outperform the 2x1 SFBC scheme using single user MMSEC without being necessary to estimate the noise power at the MT for the *J*-MCZF and either noise power and the channel response for the MCZF, therefore keeping it at a very low complexity.

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REFERENCES

[1] K. Fazel, S. Kaiser, *Multi-Carrier and Spread Spectrum Systems* Wiley, 2003.

[2] S.Hara, R. Prasad, "Overview of multicarrier CDMA", *IEEE Communications Magazine*, vol.35, December 1997.

[3] IST 4MORE project, web site *http://www.ist-4more.org*

[4] S.M Alamouti, " A simple transmit diversity technique for wireless communications", *IEEE Journal on Selected Areas in Communications*, pp.1451-1458, vol. 16, Oct. 1998.

[5] Vahid Tarokh and A. Robert Calderbank, "Space-Time block coding for wireless communications: performance results", *IEEE Journal on Selected Areas in Comunications*, Vol. 17, No. 3, March 1999.

[6] Stefan Kaiser, "Space Frequency Block Coding in the Uplink of Broadband MC-CDMA Mobile Radio Systems with Pre-Equalization", IEEE VTC Fall 2003.

[7] B. R. Vojcic, W.M. Jang, 'Transmitter Precoding in Synchronous Multiuser Communications,' *IEEE Transact. on Communications*, vol. 46, no. 10, pp. 1346-11355, Oct. 1998.

[8] S. Haykin, *Adaptive Filter Theory 3rd Ed.*, Prentice Hall, 1996

[9] J. Medbo, 'Channel Models for Hiperlan/2 in different Indoor Scenarios,' *ETSI BRAN doc*. 3ERI085b, 1998