SFBC with Pre-Filtering Technique for DL TDD
MC-CDMA Systems in High Data Rate Context

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Abstract—This paper deals with downlink space-frequency block coding with a pre-filtering technique for time division duplex MC-CDMA. We consider the use of antenna arrays at the base station (BS) and a single antenna at the mobile terminal (MT). The space-frequency block codes allow that the spatial diversity be efficiently exploited and the linear multi-user pre-filtering technique allows to format the transmitted signals so that the multiple access interference (MAI) and channel distortions at mobile terminals are completely removed. The aim of this technique is to transfer the most computational complexity from the receiver were proposed [3][4]. 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. Their effectiveness relies however on the assumption that the channel coefficients remain constant for the duration of two consecutive symbols in order to guarantee the diversity gain. This requirement can be critical in OFDM systems. To avoid the necessity to use two successive OFDM symbols for coding, the combined symbols can be sent on different subcarriers in multi-carrier system, since it can be designed in a way that two adjacent narrowband subchannels are affected by nearly the same channel coefficients. Thus space frequency block coding (SFBC) requires only the reception of one OFDM symbol for detection thus avoiding the problems related to with coherence time restrictions and also reduces the delay in the detection process. In [5] an efficient realization of SFBC for OFDM system was introduced.

Unlike the BS, where these requirements are not so tight, low complexity is required at MTs, thus only simple and single user (SU) equalizers techniques can be implemented, limiting the MAI cancellation capability. Considering TDD, another approach consists in performing pre-filtering at transmitter side using the TDD channel reciprocity between alternative UL and DL transmission periods. This solution allows to move the MAI mitigation task from the MT to the BS. The general concept of linear multi-user pre-filtering can be found in [6], and recently some linear approaches have been proposed for DL TDD MC-CDMA [7][8].

This paper proposes a space-frequency coding with pre-filtering technique for the downlink of MC-CDMA systems. We analytically derive the pre-filtering approach for 2x1 SFBC MC-CDMA, using the same principles of the approach presented in [7]. Here, SFBC allows that the spatial diversity to be efficiently exploited and the multi-user pre-filtering technique allows formatting the transmitted signal so that the MAI and channel distortions at MT are completely removed, keeping MT at very low complexity. The pre-filtering processing is performed in the frequency domain. The algorithm is based on zero forcing criterion and minimization of the transmitted power. This issue is very important because it avoids an excessive transmit power allocation.
The paper is organized as follows: In section II we present the proposed advanced downlink MC-CDMA system. In section III, we analytically derive the pre-filtering algorithm with SFBC which we call multi-user constrained zero forcing (Pre-MCZF). In section IV, we present some simulation results obtained with the Pre-MCZF for two modulations schemes: 8-PSK and 16-QAM in order to assess the pre-filtering algorithm in a high data rate context. We also compare the Pre-MCZF with 2x1 SFBC schemes using SU equalizer techniques such as, Zero Forcing Combining (ZFC) and Minimum Mean Square Error Combining (MMSEC). Finally, the main conclusions are presented in section V.

II. SYSTEM MODEL

Figure 1 shows the proposed downlink MC-CDMA transmitter with K active users and the receiver of the $g^t$ MT. After channel encoding, puncturing and interleaving operations, the bit streams are mapped to an 8-PSK or 16-QAM 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 $c_k = [c_{k,1}, ..., c_{k,q}, ..., c_{k,L}]^T$, where $c_{k,q}$ is the $q^{th}$ chip code and $(.)^T$ denotes the transpose operator. Then each spreaded data symbol are weighted by a vector $w_k^p = [w_{k,1}^p, ..., w_{k,q}^p, ..., w_{k,L}^p]$ of size $L$, where $w_{k,q}^p$ is the $q^{th}$ weight chip of the $k^{th}$ user and of the $p^{th}$ symbol. We compute a vector weight for each data symbol, which are calculated using the CSI according the criterion presented in next section. After the pre-filtering operation the signal for a generic user $k$ is given by,

$$s_k = [s_{k,1}^T, ..., s_{k,L}^T] = [s_{k,1}, ..., s_{k,n}, s_{k,n+1}, ..., s_{k,N_c}]^T$$

(1)

of size $N_c$, where $s_{k,p}^p$ corresponding to the specific symbol $p$ is of size $L$ and given by,

$$s_{k,p}^p = a_k^p c_k \cdot w_k^p = [s_{k,1}^{p,q}, ..., s_{k,q}^{p,q}, ..., s_{k,L}^{p,q}]^T$$

(2)

with $(.)$ denoting the element wise product vector, and therefore the index relation between the components of vectors $s_{k,p}^p$ and $s_k$ is such that,

$$s_{k,n}^p = s_{k,(p-1)L+n}$$

(3)

Then the chips of the sequence $s_k$ are interleaved in frequency domain to produce sequence $\bar{s}_k$. In this new sequence the chips of each data symbol are separated by $L$ positions in the OFDM symbol. We consider frequency non-selective Rayleigh fading per subcarrier. However, due to the frequency interleaving operation, each data symbol experiment $L$ uncorrelated frequency complex channel fading coefficients increasing the frequency diversity gain. The mapping scheme of the sequence $\bar{s}_k$ for SFBC with two transmit antennas is shown in Table 1.

Table 1: SFBC mapping [5] with 2 transmitter antennas.

<table>
<thead>
<tr>
<th>Subcarrier $n$</th>
<th>Antenna 1</th>
<th>Antenna 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\bar{s}_{k,n}$</td>
<td>$\bar{s}_{k,n}^*$</td>
<td>$\bar{s}_{k,n+1}^*$</td>
</tr>
<tr>
<td>$\bar{s}_{k,n+1}$</td>
<td>$\bar{s}_{k,n+1}^*$</td>
<td>$\bar{s}_{k,n}^*$</td>
</tr>
</tbody>
</table>

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 subchannels $n$ and $n+1$ after OFDM demodulation and GI removal are given by,

$$y_{g,n} = \sum_{k=1}^{K} \bar{s}_{k,n} h_{g,1,n} + \sum_{k=1}^{K} \bar{s}_{k,n+1} h_{g,2,n} + n_n$$

(4)

$$y_{g,n+1} = \sum_{k=1}^{K} \bar{s}_{k,n+1} h_{g,1,n+1} + \sum_{k=1}^{K} \bar{s}_{k,n} h_{g,2,n+1} + n_{n+1}$$

where $h_{g,n,m}$ represents the frequency complex fading channel of the $g^t$ MT, of the $m^{th}$ antenna branch and of the $n^{th}$ sub-channel; $n_n$ is the additive white Gaussian noise on sub-carrier $n$. OFDM systems are usually designed so that the sub-carrier separation is significantly lower than the coherence bandwidth.

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of the channel, and therefore, the fading in two adjacent subcarriers can be considered flat i.e. we can consider that \( h_{g,m,n} = h_{g,m,n+1} \).

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

\[
z_{g,m,n} = h_{g,m,n}^* / \left( \sum_{m=1}^{D} h_{g,m,n} \right)
\]  

These equalization coefficients are slightly different from the ones used by an EGC equalizer, which are given by,

\[
z_{g,m,n}^{\text{egc}} = h_{g,m,n}^* / \left| h_{g,m,n} \right|
\]  

The reason to use the equalization coefficients of (5) instead of (6), 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 (6).

The signal for one arbitrary pair of adjacent subcarriers \( n \) and \( n+1 \), after using the SF combining scheme and the equalization coefficients defined in (5), can be written as,

\[
\begin{align*}
    r_{g,n} &= z_{g,1:n,y,n} + z_{g,2:n+1,y,n}^* \\
    r_{g,n+1} &= -z_{g,2:n,y,n}^* + z_{g,1:n+1,y,n+1}^*
\end{align*}
\]  

After some mathematical manipulations and considering that \( z_{g,m,n} = z_{g,m,n+1} \), we can write,

\[
\begin{align*}
    r_{g,n} &= \frac{h_{g,1,n}^2 + h_{g,2,n}^2}{h_{g,1,n}^2 + h_{g,2,n}^2} \left( z_{g,n} + \text{MAI}_n + z_{g,1:n,y,n}^* + z_{g,2:n+1,y,n+1}^* \right) \\
    r_{g,n+1} &= \frac{h_{g,1,n}^2 + h_{g,2,n}^2}{h_{g,1,n}^2 + h_{g,2,n}^2} \left( z_{g,n+1} + \text{MAI}_{n+1} - z_{g,2:n,y,n}^* + z_{g,1:n+1,y,n+1}^* \right)
\end{align*}
\]  

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

\[
\hat{d}_g^p = \frac{\text{DesiredSignal}}{\sum_{k=1,k \neq g}^{K} c_k^H \left( f_g^p \ast w_g^p \ast c_k \right) d_k^p + n_g Noise}
\]  

where \( f_g^p \) is a vector of size \( L \), and each elements given by,

\[
f_g^p = \left[ \begin{array}{c} \sum_{m=1}^{D} h_{g,m,1}^p \\
                        \sum_{m=1}^{D} h_{g,m,2}^p \\
                        \vdots \\
                        \sum_{m=1}^{D} h_{g,m,L}^p \end{array} \right]
\]  

where \( h_{g,m,n}^p \) represents the frequency complex fading channel for a generic symbol \( p \). The signal of (9) involves the three well known terms: the desired signal, the MAI caused by loss of code orthogonality among the users, and the residual noise after despreading.

III. PRE-FILTERING ALGORITHM FOR 2X1 SFBC MC-CDMA

Due to complexity restrictions, only SU detectors can be employed at MT, limiting the MAI reduction capability. The aim of the pre-filtering algorithm is to transfer the multiuser detector from MT to BS, where the complexity problem is less restrictive. The multi-user pre-filtering completely removes the MAI and channel distortions at MTs by pre-formating the signal so that the received signal at the decision point is free from interferences without enhancing the noise power at receiver side. The algorithm is based on zero forcing criteria and is designed in frequency domain in order to remove the MAI term at all MTs. Furthermore, it is designed taking into account the transmitted power at BS, reason why we call this algorithm the multi-user constrained zero-forcing (Pre-MCZF).

The interference that the signal of a given user \( g \) produces at an other \( k \) MT is from (9) given by,

\[
\text{MAI}(g \rightarrow k) = c_k^H (f_g^p \ast w_g^p \ast c_k) = v_{k,g}^T w_g^p
\]  

with \( v_{k,g} = c_k \ast f_g^p \ast c_k \).

The weight vector of the \( g^{th} \) MT 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,

\[
\begin{align*}
    c_k^H (f_g^p \ast w_g^p \ast c_k) &= \alpha_g \\
    c_k^H (f_g^p \ast w_g^p \ast c_k) &= 0 \quad \forall k \neq g
\end{align*}
\]  

The aim of the proposed approach is remove the channel distortions of the \( g^{th} \) user data symbol at the \( g^{th} \) MT and at the same time completely remove the MAI that this data symbol causes in other \( K-1 \) MTs. Hence, to compute the weights for user \( g \) we have to solve a linear system of \( K \) equations given by,

\[
A_g w_g^p = \alpha_g b_g
\]  

where \( A_g \) is a \( K \times L \) matrix and \( b_g \) a vector of size \( K \), given by,

\[
A_g = \left[ \begin{array}{c} v_{1,g}^T \\
                        v_{1+1,g}^T \\
                        \vdots \\
                        v_{K-1,g}^T \\
                        v_{K,g}^T \end{array} \right]
\]  

\[
b_g = \left[ \begin{array}{c} 0 \\
                        0 \\
                        \vdots \\
                        1 \end{array} \right]
\]  

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). This problem can be solved using the method of Lagrange multipliers [9]. After same mathematical manipulations [7], we obtain the Pre-MCZF based pre-filtering vector,

\[
w_g = \alpha_g A_g^p (A_g^p A_g^p)^{-1} b_g = \alpha_g A_g^p \psi^{-1} b_g
\]  

(15)

where \(\psi = A_g^p A_g^p\) is a real square and Hermitian matrix of size \(KxK\) and \(\alpha_g\) is a constant used to normalize the vector according,

\[
|w_g|^2 = w_k^p w_k^p = 1 \quad \forall k = 1...K
\]  

(16)

ensuring that the pre-filtering weight vector always has unit norm in order to compare with systems without pre-equalization. It can be seen from (10) that we only need to know the modulus of the channel frequency at the BS. Basically, we just perform a pre-amplitude equalization. Observing equation (15) it easy to see that the most computational intensive task to obtain the weights, comes from matrix \(\psi^{-1}\) inversion. However, the size of this real matrix is just \(KxK\) independently of the spreading factor and the number of antennas, which makes this algorithm very attractive for real time applications. Note that in approach discussed in [7], a complex matrix should be inverted, increasing the BS complexity.

IV. NUMERICAL RESULTS

To evaluate the performance of the proposed advanced SFBC MC-CDMA system with pre-filtering algorithm, we use an indoor uncorrelated Rayleigh fading channel, whose system parameters are derived from the European BRAN Hiperlan/2 standardization project [10]. 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.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of Carriers</td>
<td>1024</td>
</tr>
<tr>
<td>Spreading factor</td>
<td>32</td>
</tr>
<tr>
<td>Guard period samples/time</td>
<td>256 / 3.2(\mu)s</td>
</tr>
<tr>
<td>Number of users</td>
<td>32 (full load)</td>
</tr>
<tr>
<td>Total OFDM symbol duration (T_s)</td>
<td>16(\mu)s</td>
</tr>
<tr>
<td>Sampling frequency/Bandwidth</td>
<td>80MHz / 64MHz</td>
</tr>
<tr>
<td>Frame duration</td>
<td>32(\times)(T_s) = 512(\mu)s</td>
</tr>
<tr>
<td>Channel Coding</td>
<td>without / UMTS turbo code (TC)</td>
</tr>
<tr>
<td>Modulation</td>
<td>8-PSK / 16-QAM</td>
</tr>
<tr>
<td>Overall data rate</td>
<td>128 Mbps</td>
</tr>
<tr>
<td>(1/2 TC and 16-QAM)</td>
<td></td>
</tr>
<tr>
<td>Channel profile</td>
<td>BRAN A</td>
</tr>
<tr>
<td>Maximum delay/ Number of paths</td>
<td>390ns / 18</td>
</tr>
</tbody>
</table>

It is assumed that the receiver and transmitter have perfect knowledge of the channel. We considered a DL synchronized transmission using Walsh-Hadamard spreading sequences. The channel is considered to be flat at least between two sub-carriers. The transmitted power is normalized to one in all presented schemes.

We compare the multi-user pre-filtering with Alamouti’s SFBC using conventional SU equalizers at the MT, such as MMSEC and ZFC. We present results for high order modulations 8-PSK and 16-QAM in order to assess the proposed scheme in high data rate context scenarios. The channel coding scheme is the turbo-code defined in current UMTS with a rate of 1/3, combined with a puncturing process and an interleaver to have an overall coding rate of 1/2.

Figure 2, 3 show the performance results for 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 without channel coding, while Figure 4, 5 show the performance for BER and frame error rate (FER) as function of Et/No with the referred turbo code.

Figure 2 shows the performance of the proposed Pre-MCZF algorithm against SFBC schemes employing two antennas and using SU ZFC and MMSEC equalization. To have the same spectral efficiency of 3 bps/Hz in all cases, the modulation scheme is 8-PSK. As can be observed from this figure the performance of the Pre-MCZF outperforms the two SU 2x1 SFBC schemes. For a BER=1E-2, we got approximately 1.9dB when compared with 2x1 SFBC using MMSEC and ZFC equalization. However it should be noted that when MMSE is applied the noise variance must be estimated at MT increasing the complexity. While Pre-MCZF outperforms SFBC using MMSEC equalizer without being necessary estimate the noise variance at MT. Figure 3 shows the performance of the proposed Pre-MCZF algorithm against 2x1 SFBC schemes for spectral efficiency of 4 bps/Hz, the modulation scheme in this case is 16-QAM. In this scenario we also can observe that the performance of the Pre-MCZF outperforms all SU 2x1 SFBC schemes. For a BER=1E-2, we got approximately 2.0dB gain, as compared with 2x1 SFBC schemes using SU MMSEC and ZFC equalization.

Figure 4 shows the performance of the proposed Pre-MCZF algorithm against 2x1 SFBC schemes for the same scenario of the Figure 3, but now with the referred turbo code, decreasing the spectral efficiency for 2bps/Hz. In this case we also can see that the performance of the Pre-MCZF outperforms all 2x1 SFBC schemes. For BER=1E-5, we get approximately 1.0dB and 2.0 dB gain, when compared with 2x1 SFBC schemes using MMSEC and ZFC equalization, respectively. In Figure 5 we use the same scenario of the last figure, but now using the average frame error rate (FER) as metric. In this case we also can see that the performance of the Pre-MCZF outperforms all 2x1 SFBC schemes. For BER=1E-3, we get approximately 1.0dB and 2.2 dB gain, when compared with 2x1 SFBC schemes using MMSEC and ZFC equalization, respectively. The results presented in the last two figures show that the Pre-MCZF algorithm is suitable for data and multimedia applications.

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We proposed a 2x1 SFBC with pre-filtering algorithm for the DL of TDD MC-CDMA systems, using two antennas at BS and single antenna at mobile terminal. We analytically derived the proposed linear multi-user pre-filtering algorithm, based on a constrained zero-forcing criterion and minimization of the transmitted power. The performance was assessed for scenarios without channel coding and considering the use of the turbo codes specified for UMTS. We compared the performance of proposed Pre-MCZF against 2x1 SFBC using MMSEC and ZFC SU equalizers. We demonstrated that Pre-MCZF algorithm allows a significant improvement of the user capacity either for low and high data rate applications, outperforming the 2x1 SFBC schemes while using SU equalizers, without being necessary to estimate the noise variance at MT, keeping it at low complexity.

ACKNOWLEDGMENT

The work presented in this paper was supported by the European project IST-2001-507039 4MORE and Portuguese Foundation for Science and Technology (FCT) through project POSI/CPS/46701/2002 and grant to the first author.

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