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DOWNLINK STRATEGIES USING ANTENNA ARRAYS FOR INTERFERENCE MITIGATION IN MULTI-CARRIER CDMA

Abstract. The paper considers transmitter strategies for interference mitigation in the downlink of a multi-carrier CDMA system using antenna arrays. We describe typical indoor and outdoor scenarios and derive the corresponding transmitter strategy according to the channel knowledge available at the base station for each case. We show the effectiveness of multi-user space-frequency transmit filtering for the indoor and single-user beamforming for the outdoor scenario. Both strategies considerably reduce the multiple access interference and thus allow a low-complexity receiver design for the mobile terminal.

1. INTRODUCTION

Virtues such as robustness to multipath propagation and flexibility of the multi-user access make Multi-Carrier CDMA (MC-CDMA) a transmission scheme that is now considered by a growing community of researchers striving for new efficient air interfaces [1][2]. Recent publications show that this scheme is particularly advantageous for the Down-Link (DL), i.e. from a Base-Station (BS) to Mobile Terminals (MT) [3]. However, like all CDMA-based systems, MC-CDMA suffers from Multiple Access Interference (MAI), which is caused by the loss of orthogonality among the users' signals in multipath propagation. MAI mitigation has, therefore, been a challenging research topic since the very beginning of studies on MC-CDMA [1]. Yet, the frequently considered approach of performing Multi-User Detection (MUD) at the receiver is quite unattractive for the DL, because it entails an increase of complexity and power consumption at the MTs [1][4]. Here, we propose an alternative approach using an antenna array at the BS.

Antenna arrays endow a wireless system with the spatial dimension, which can be exploited in various ways. Striving for a light design of the MT, we only consider an array for transmission at the BS. Here, the strategy to adopt essentially depends on the knowledge about the propagation channel that is available at the BS prior to transmission. Three cases may be distinguished: If the BS has no knowledge at all, we may opt for a transmit diversity or space time coding scheme, e.g. [5]. These systems transmit redundant information over the different antenna branches to benefit from spatial diversity by combining or decoding at the receiver side. In contrast, if the BS has perfect instantaneous knowledge of the channel fading, it is preferable to adapt the transmitted signal to the channel conditions. Such techniques are generally referred to as pre-filtering or pre-coding. In a multi-user context they not only aim at pre-compensating the channel fading but can also reduce the MAI. The general concept of multi-user prefiltering can be found in [6], and we already proposed approaches for prefiltering in space and frequency applied to MC-CDMA in [7] and [8]. When only partial knowledge, e.g. the Directions of Departure (DODs) of the multipath components or any other related long-term spatial channel statistics, is available, Beamforming (BF) may be a good choice, especially when the channels at different antennas are correlated. BF exploits the spatial separation of MTs by adapting the antenna pattern so as to illuminate only desired directions and avoid interference at other MTs positions [9]. BF applied to MC-CDMA was studied in [10] for reception in the Up-Link (UL).

In this paper, we focus on the latter two cases where we have either instantaneous knowledge, which may be available at the BS from estimation of the UL in a Time Division Duplex (TDD) system with low mobility, or long-term spatial statistics, which can still be obtained in scenarios with high mobility and arbitrary duplex mode [11]. In both cases, we can use the channel knowledge to transfer at least some of the demanding signal processing tasks for MAI reduction from the MT to the BS.

The paper is organised as follows. We present the system model of a DL MC-CDMA transmission with an antenna array and the corresponding transmit filter at the BS in section 2. The scenarios and system considerations are exposed in section 3. Section 4 is dedicated to the optimisation of the transmit filter in the two mentioned scenarios leading to Space-Frequency Transmit Filtering (SFTF) and Beamforming (BF), respectively. Numerical results for both scenarios including an assessment of the impact of Doppler variations are presented in section 5. Finally, we give concluding remarks in section 6.

2. THE SYSTEM MODEL

We consider the DL MC-CDMA system depicted in figure 1. Like in the conventional system, the data symbols, e.g. QPSK symbols, of users 1 to K are spread into L chips using orthogonal codes $\mathbf{c}_k$ (k=1,..,K), for instance Walsh-Hadamard codes. These chips are then copied M times for the M antenna branches. Both operations are mathematically represented by vector $\tilde{\mathbf{c}}_k=[\mathbf{c}_k^T,\ldots,\mathbf{c}_k^T]^T$ of length ML, which is a repetition of the code vector $\mathbf{c}_k$ of length L. $(.)^T$ denotes vector transposition. These chips are weighted by the transmit filter, whose components are represented by vector $\tilde{\mathbf{w}}_k$ of length ML. The transmit filters are calculated using available channel knowledge, c.f. section 3. Finally, the contributions of all users are summed chip-by-chip, and the result is mapped on L subcarriers of the Orthogonal Frequency Division Multiplex (OFDM) system on each antenna branch.

The underlying OFDM transmission uses a guard interval Δ and is assumed to be ideal in the sense that the channel can be represented in the frequency domain by a single flat fading coefficient on each subcarrier. Hence, we can represent the channel between the M antennas of the BS and the single antenna of MT g by ML complex fading coefficients gathered in vector $\mathbf{h}_g$.

To keep the complexity at the MT as low as possible, we only use a single antenna and Single User Detection (SUD) techniques. Thus, the receiver antenna implicitly recombines the M signals from the transmit array in space. Then, after OFDM demodulation and a potential channel estimation, the receiver equalises and despreads the signals of the L subcarriers using vector $\mathbf{q}_g=[q_g(1),\ldots,q_g(L)]^T$. In analogy to the transmitter description, we use an expanded vector $\tilde{\mathbf{q}}_g=[\mathbf{q}_g^T,\ldots,\mathbf{q}_g^T]^T$

of size ML to mathematically represent the signal recombination in space and frequency. The resulting decision variable for MT g is given by

$$\hat{d}_g = \underbrace{\tilde{\mathbf{q}}_g^H \cdot \left(\mathbf{h}_g \circ \tilde{\mathbf{w}}_g^* \circ \tilde{\mathbf{c}}_g\right) \; d_g}_{\text{Desired Signal}} + \underbrace{\tilde{\mathbf{q}}_g^H \cdot \left(\mathbf{h}_g \circ \sum_{k=1,k\neq g}^{K} \left(\tilde{\mathbf{w}}_k^* \circ \tilde{\mathbf{c}}_k\right) \; d_k\right)}_{\text{MAI}} + \underbrace{\mathbf{q}_g^H \cdot \mathbf{n}_g}_{\text{Noise}} \qquad (1)$$

where vector $\mathbf{n}_g$ gathers the noise samples on the L subcarriers, $(.)^*$ is the complex conjugate operator, $(.)^H$ denotes the conjugate transposition, and $\circ$ represents the element-wise vector multiplication. The decision variable is the sum of the desired signal, the MAI and the noise after subcarrier combining. We notice that the desired signal and the MAI are functions of the transmit filters. In contrast to a conventional MC-CDMA DL, the MAI, i.e. the loss of the signal orthogonality installed by the spreading codes, arises here not only from the channel fading but from the combination of channel fading and transmit filtering. As this combination is specific to each user, a subcarrier combining aiming at restoring the orthogonality among users' signals is not advantageous. Therefore, we use Equal Gain Combining (EGC), i.e. phase equalisation on each subcarrier and despreading, as low complexity SUD scheme. The corresponding weight vector $\mathbf{q}_g$ depends on the transmit filter, and we derive it in section 3.

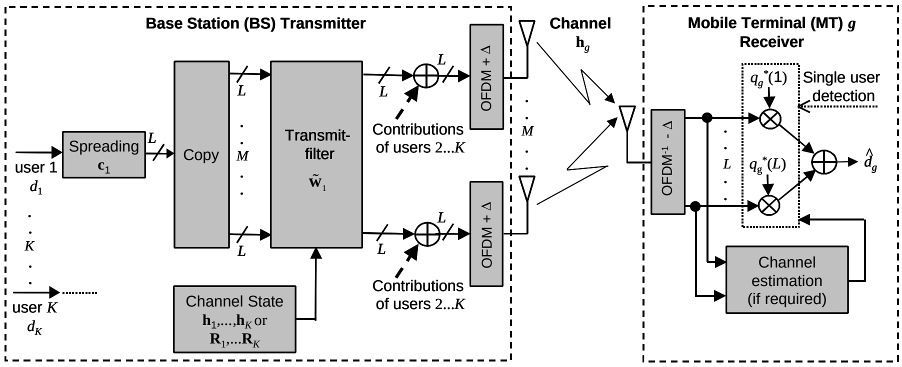

Figure 1: Downlink MC-CDMA system using an antenna array at the BS

3. TRANSMISSION SCENARIOS

The transmit filtering strategy depends on the transmission scenario and the related channel knowledge available at the BS. In the sequel, two different scenarios are identified.

3.1 Indoor scenario

The indoor scenario is characterised by a low mobility of the MT, typically less than 10 km/h. This means that the coherence time of the channel is much greater than the frame duration of the transmission. Hence, in a TDD system, the fading coefficients estimated during the UL transmission slot are still valid for the following DL transmission slot. We can thus assume to have perfect knowledge of the channel coefficient vector $\mathbf{h}_g$. Concerning the spatial properties, the BS may be surrounded by many obstacles so that the Directions of Arrival (DOAs) of the multipath components are spread over a large Angular Sector (AS). For instance, we will assume a uniform distribution of the DOAs of all users within 120°. Finally, due to the stationarity of the channel the DOAs of the UL are equal to the DODs in the DL.

3.2 Outdoor scenario

As a consequence of the potentially high mobility of MTs in an outdoor scenario, we cannot assume that the channel fading is constant during consecutive UL and DL slots. A measure of these channel variations is the correlation of a given element of $\mathbf{h}_g$ taken at two instants separated by a duration t. Denoting this element by $h_{m,\ell}(t)$ and assuming a Jakes Doppler spectrum [12], the correlation r is given by the following expression:

$$\mathrm{E}\left[h_{m,\ell}(t)\cdot h_{m,\ell}^{\ *}(t+t)\right]= r = J_0(2pt\,v_{MT}\,/\,l_c) \tag{2}$$

Here, v_{MT} is the mobile speed, l_C the carrier wavelength and J_0 denotes the zero order Bessel function of the first kind.

Concerning the spatial channel characteristics, the DOAs of each user generally lay in a smaller AS than in the indoor scenario, since the reflecting obstacles may be located far from the isolated BS. For instance, we will assume a uniform distribution of the DOAs in an AS of 10° around the main DOA, which itself is randomly chosen in a 120° sector for each user. Even in the high mobility case, we can still assume that the DOAs of the UL slot are identical to the DODs of the following DL slot.

As a consequence, we assume that only long term channel knowledge is available at the BS in the outdoor scenario. This means that the BS knows the spatial covariance matrices, $\mathbf{R}_g$. Let $\mathbf{h}'_g(\ell)$ gather the M coefficients of the channel between the BS array and MT g on subcarrier ℓ, then $\mathbf{R}_g$ is given by

$$\mathbf{R}_g = \mathrm{E}\left[\sum_{\ell=1}^{L}\mathbf{h}'_{\mathbf{g}}(\ell)\cdot\mathbf{h}'^{H}_{g}(\ell)\right] \tag{3}$$

where $\mathrm{E}[x]$ denotes the expected value of x. Note that $\mathbf{R}_g$ is simply averaged over the L subcarriers, because we assume that, with frequency interleaving, there is no correlation of the fading on the L subcarriers. It has been shown that these matrices can also be obtained from the UL even in FDD systems [11].

4. TRANSMIT FILTERING STRATEGIES

The transmit filtering strategies use the available channel knowledge at the BS to form the transmitted signal of each user with respect to Single-User (SU) and Multi-User (MU) optimisation criteria. The SU criteria are low complex and simply aim at maximising the desired signal part of the decision variable without taking into account the MAI. However, due to the spatial selectivity of the transmitted signal, SU techniques implicitly reduce the MAI generated at other MTs. The MU criteria explicitly reduce the MAI at the cost of a higher complexity, but this computational burden may be tolerable at the BS. It has to be noted that, in the DL, the spatial dimension can only be used for MAI cancellation at the BS side. Indeed, at a given MT, all signals have passed through the same space-frequency channel.

As it can be seen in figure 1, the transmit filters are placed before the OFDM modulation, which corresponds to filtering in the frequency domain. A time domain approach would imply a distinct OFDM operation for each user and, therefore, would be disadvantageous in terms of computational complexity.

We consider two distinct approaches: A joint optimisation of the transmit filter in space and frequency, called Space-Frequency Transmit Filtering (SFTF), when instantaneous channel knowledge is available, and an optimisation in space only, i.e. Beamforming (BF), if the BS has long-term channel knowledge only.

When modifying the transmit signal, the resulting signal power is likely to vary as well. Since power control is out of the scope of this paper, we ensure that all transmit filters are power normalised:

$$\left|\tilde{\mathbf{w}}_k\right|^2 = \tilde{\mathbf{w}}_k \tilde{\mathbf{w}}_k^H = 1 \quad \forall k = 1...K \tag{4}$$

Yet, it is worth noting that the presented techniques can readily be used in a joint transmit filtering and power control scheme, e.g. [13].

4.1 Space-Frequency Transmit Filtering (SFTF)

Space Frequency Transmit Filtering (SFTF) is a combination of spatially selective transmission and pre-filtering in the frequency domain. It is intended for the indoor scenario exposed in section 3.1. Hence, we assume that we have perfect knowledge of the channel vectors $\mathbf{h}_g$. We already presented several versions of SFTF in [8]. Here, we will focus on the SU criterion called Maximum Ratio Transmission (MRT) and a MU criterion based on a maximisation of the Signal over Interference plus Noise Ratio (SINR). Since instantaneous channel knowledge is available, we can include pre-equalisation in the transmit filter. This allows to simplify the detection at the MT to a pure despreading, i.e. $\tilde{\mathbf{q}}_g^H = \tilde{\mathbf{c}}_g^H$. So, there is no more need for channel estimation on each subcarrier at the MT side.

For the SU-SFTF criterion, the optimisation of the filter is based on the maximisation of the signal to noise ratio (SNR) after despreading. Since the noise term in (1) is not affected by the transmit filter, this is equivalent to maximising the desired signal part with a given transmit power. This criterion is well known and analogous to maximum ratio combining at the receiver. Hence, at the transmitter we may call it MRT. The corresponding transmit filtering vector is

$$\text{SU-SFTF:} \qquad \tilde{\mathbf{w}}_g = k_g \mathbf{h}_g \tag{5}$$

where the scalar k_g is used to meet the power constraint in (4).
The MU-SFTF criterion is based on maximising the SINR. Since the MAI in (1) is a function of the transmit weights of all other users $k \neq g$, direct SINR maximisation would lead to a joint optimisation problem for the transmit filters of all users at once. [8] presents an approach using a modified SINR (m-SINR) and decoupled optimisation, which leads to a closed form solution for the transmit filters of each user. The basic idea of this m-SINR is to replace the interference term of MT g by the sum of the interference that user g creates at the other MTs $k \neq g$. Assuming that the power of the data symbols is unity, the m-SINR is given as:

$$m-SINR_g = \frac{\left|\tilde{\mathbf{w}}_g^H \mathbf{h}_\mathbf{g}\right|^2}{\tilde{\mathbf{w}}_g^H \left(\sum_{k=1, k\neq g}^{K} \left(\tilde{\mathbf{c}}_k^H \circ \mathbf{h}_k \circ \tilde{\mathbf{c}}_g\right)\left(\tilde{\mathbf{c}}_k^H \circ \mathbf{h}_k \circ \tilde{\mathbf{c}}_g\right)^H \right) \tilde{\mathbf{w}}_g + \sigma_n^2} \tag{6}$$

with σ_n^2 being the noise variance per subcarrier. (6) has to be maximised under the power constraint in (4). When we include this constraint, we see that the maximisation problem itself gets independent of a scalar factor in $\tilde{\mathbf{w}}_g$. This means that we can add the term $k=g$ to the sum in the denominator without any impact. Defining the $ML \times K$ matrix $\mathbf{A}_g = \left[\tilde{\mathbf{c}}_1^H \circ \mathbf{h}_1 \circ \tilde{\mathbf{c}}_g, \ldots, \tilde{\mathbf{c}}_k^H \circ \mathbf{h}_k \circ \tilde{\mathbf{c}}_g, \ldots, \tilde{\mathbf{c}}_K^H \circ \mathbf{h}_K \circ \tilde{\mathbf{c}}_g\right]$ and a vector $\bar{\mathbf{w}}_g = \tilde{\mathbf{w}}_g / k_g$, where the scalar k_g is used to ensure (4), the optimum transmit filtering vector is obtained from

$$\max_{\bar{\mathbf{w}}_g} \frac{\left|\bar{\mathbf{w}}_g^H \mathbf{h}_\mathbf{g}\right|^2}{\bar{\mathbf{w}}_g^H \left(\mathbf{A}_g \mathbf{A}_g^H + \sigma_n^2 \mathbf{I}_{ML}\right) \bar{\mathbf{w}}_g} \quad \text{subject to } \left|\bar{\mathbf{w}}_g^H \mathbf{h}_g\right| = 1 \tag{7}$$

where $\mathbf{I}_X$ denotes the identity matrix of size X. After solving (7) with the method of Lagrange multipliers and some simplifications we finally get:

$$\text{MU-SFTF:} \quad \tilde{\mathbf{w}}_g = k_g \mathbf{A}_g (\mathbf{A}_g^H \mathbf{A}_g + \sigma_n^2 \mathbf{I}_K)^{-1} \mathbf{b}_g \text{ , with } \mathbf{b}_g = [0, \ldots, \underset{g\text{-th element}}{1}, \ldots, 0]^T \tag{8}$$

In contrast to SU-SFTF, MU-SFTF provides explicit MAI reduction at the price of a higher complexity, since it implies a matrix inversion of size $K \times K$.

4.2 Beamforming (BF)

In the second scenario, where the BS has long term channel knowledge only, spatial transmit filtering, i.e. BF, can be performed. We here assume that the BS has knowledge about the spatial covariance matrices $\mathbf{R}_g$ defined in (3). The optimisation criteria for BF are very close to those described for SFTF, with the only difference that they are averaged over the channel fading. We represent $\tilde{\mathbf{w}}_g$ as a repetition of BF vector $\mathbf{w}_g=[w_g(1),\ldots,w_g(M)]^T$, which is identical for all carriers, i.e. $\tilde{\mathbf{w}}_g = [w_g(1) \xleftrightarrow{L \text{ times}} w_g(1),\ldots, w_g(M) \xleftrightarrow{L \text{ times}} w_g(M)]^T$. After despreading, the power of the desired signal averaged over the channel fading and the subcarriers can be expressed by

$$\mathrm{E}\left[\tilde{\mathbf{w}}_g^H \mathbf{h}_g \mathbf{h}_g^H \tilde{\mathbf{w}}_g\right] = \mathrm{E}\left[\sum_{\ell=1}^{L} \mathbf{w}_g^H \mathbf{h}'_g(\ell) \mathbf{h}'^H_g(\ell) \mathbf{w}_g\right] = \mathbf{w}_g^H \mathbf{R}_g \mathbf{w}_g \tag{9}$$

(9) is maximised by the principal eigenvector (m_eig(.)) [14] of $\mathbf{R}_g$ for BF, i.e.

SU-BF: $$\mathbf{w}_g = k_g \ \mathrm{m\_eig}(\mathbf{R}_g) \tag{10}$$

The scalar k_g is used to meet the power constraint in (4) and ensure a common phase at a given array element for all users. Like SU-SFTF, SU-BF yields no explicit but implicit interference reduction.

We can also formulate a MU-BF approach and obtain the interference plus noise covariance matrix $\mathbf{R}_{\mathbf{I}g}$ by averaging the denominator of the fraction we had to maximise in (7). Hence, we get

$$\mathbf{R}_{\mathbf{I}g} = \sum_{k,k\neq g} \mathbf{R}_{\mathbf{k}} + s_n^2 \mathbf{I}_M \tag{11}$$

The MU-BF vector aims then at maximising the averaged m-SINR:

$$\max_{\mathbf{w}_g} \frac{\mathbf{w}_g^H \mathbf{R}_g \mathbf{w}_g}{\mathbf{w}_g^H \mathbf{R}_{\mathbf{I}g} \mathbf{w}_g} \quad \text{subject to (4)} \tag{12}$$

The solution is the principal generalised eigenvector (gm_eig(.)) of the matrix pair [14] formed by the signal and the interference plus noise covariance matrices:

MU-BF: $$\mathbf{w}_g = k_g \ \mathrm{gm\_eig}(\mathbf{R}_g, \mathbf{R}_{\mathbf{I}g}) \tag{13}$$

As before, k_g ensures (4) and a common phase at a given array element for all users. Note that the only degree of freedom available for MU-BF is the number of antennas. As, in practice, M may be quite low compared to the number of DOAs multiplied by the number of users, the MU-BF vector may be sub-optimum and even lead to a loss of desired signal power.

When BF is performed at transmission, channel estimation and frequency domain equalisation is required in addition to despreading at the MT side. As presented in section 2, we chose in this case the EGC SUD technique [1] for its low complexity, where the coefficient on the ℓ-th subcarrier is $q_g(\ell)=c_g(\ell)e^{jj_g(\ell)}$ and $j_g(\ell)$ is the phase of the estimated channel fading on subcarrier ℓ.

5. NUMERICAL RESULTS

The numerical results for the two scenarios exposed in section 3 were obtained with a system configuration similar to the European BRAN Hiperlan/2 standardisation project. The channel model is based on the channel A defined in [15]. This channel model comprises 18 paths with a multipath spread of 390 ns. We extended this time model to a space-time model by allocating a DOA to each of the paths. The DOAs are randomly chosen within an AS of 120° (indoor) and 10° (outdoor). The channel response is recalculated at each OFDM symbol to ensure a good average over the spatial configurations. The BS is equipped with a half wavelength spaced uniform linear array. The system bandwidth is 20 MHz and the OFDM system comprises 64 subcarriers. Walsh-Hadamard spreading codes of size L=8 are considered. We perform subcarrier interleaving to avoid that the chips of a symbol are transmitted on carriers lying within the coherence bandwitdth of the channel for the sake of diversity. All plots show the uncoded average Bit Error Rate (BER) computed over all active users versus the E_t/N_0, where E_t is the energy per bit transmitted over all antennas and N_0 the noise spectral density.

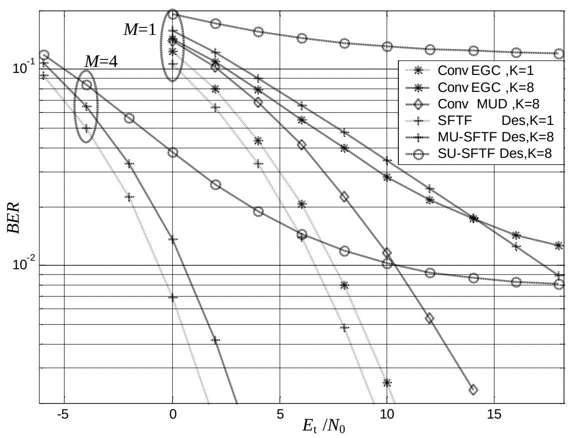

Figure 2: Comparison of SFTF to conventional systems in an indoor scenario (M=4)

Figure 2 compares the performance of SFTF, i.e. SU-SFTF (5) and MU-SFTF (8), in the indoor scenario to the Conventional (Conv) system (no transmit filtering and

M=1) with EGC and optimum linear MUD [4] at the MT receiver. For K=1 (dotted curves) SU- and MU-SFTF vectors are identical. With M=1, MU-SFTF already gains with respect to Conv EGC, because of the power weighting over the subcarriers. For M=4, the gain of MU-SFTF is about 7 dB at BER=10^{-2}, which is higher than the antenna gain itself (10log$M$=6 dB) and shows that MU-SFTF exploits space-frequency diversity in this case. For full load (K=8) and M=1, SU-SFTF has poor performance, because user-specific SU transmit filtering without spatial separation aggravates the loss of orthogonality among users' signals and thus increases the MAI. MU-SFTF has better performance, but shows no significant gain compared to Conv EGC, which means that transmit filtering without spatial separation is almost useless for the considered system. However, for K=8 and M=4, we clearly see the advantage of the antenna gain together with the spatial separation of users. Here, even low-complexity SU-SFTF outperforms Conv EGC. Compared to the Conv. System with MUD, MU-SFTF yields a gain of about 10 dB at BER=10^{-2}. Hence, joint SFTF yields a considerable MAI reduction.

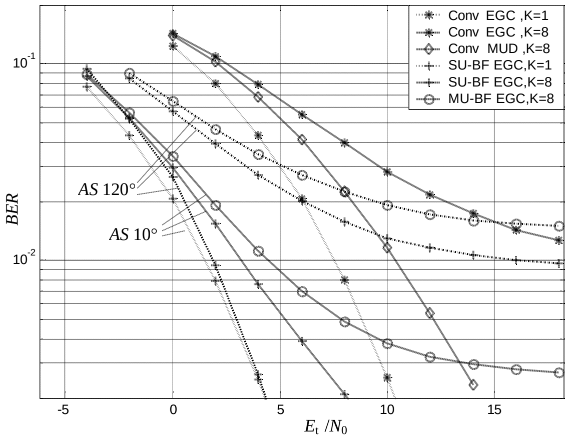

Figure 3: Comparison of BF to conventional systems in an outdoor scenario (M=4)

In figure 3, the outdoor scenario is considered and we compare the proposed BF schemes, i.e. SU-BF (10) and MU-BF (13), with M=4 to the conventional system with EGC or opt. lin. MUD at MTs. For comparison, we also plotted the curves with AS=120°. For K=1 (dotted curves) SU-BF yields the expected antenna gain, i.e. 6 dB, for AS=10°. With AS=120°, the gain is less, because the main lobe of the antenna diagram cannot cover all multipath components. For full load (K=8) and AS=10°, SU-BF outperforms Conv. MUD by about 7 dB at BER=10^{-2}. With AS=120°, SU-BF only leads an advantage at low SNR, as it experiences an error floor already for medium SNR. Thus, separation of users' signals in space only yields no advantage for a wide AS. For MU-BF, the fact that M is smaller than K and that there are multiple DOAs per user degrade the desired signal power as already explained in section 4. Therefore, the MU-BF actually has poorer performance than SU-BF.

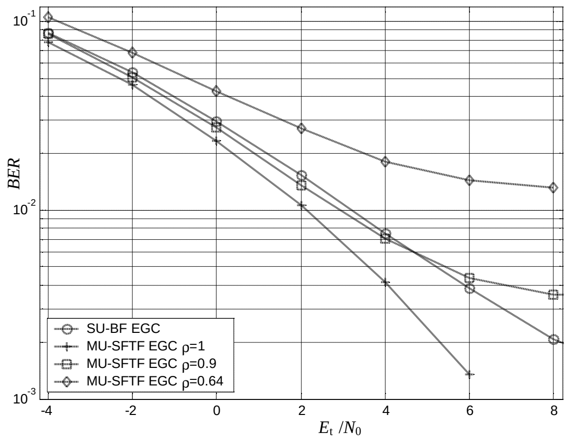

Figure 4: Comparison of MU-SFTF to SU-BF in an outdoor scenario (M=4, K=8)

In figure 4, we compare MU-SFTF and SU-BF in the outdoor scenario, i.e. AS=10° and imperfect channel estimates at the transmitter due to Doppler variations. Here, frequency equalisation is not only required for BF but also for SFTF. Note that the channel is still perfectly known at the receiver. The mobile speed considered here is in a range, where the Doppler effect has no impact on the underlying OFDM transmission. So the only effect is a mismatch in the channel estimates at the BS as quantified by (2). In this case, there is no impact on the performance of BF neither. We assume a worst case delay corresponding to a typical slot duration, i.e. 1 ms, between estimation and usage of the channel coefficients and mobile speeds of 20 km/h and 45 km/h. This leads to a correlation of r=0.9 and r=0.64, respectively. Since SFTF exploits the channel diversity, its performance is reduced by the smaller AS. Thus, even with perfect channel knowledge (r=1) for K=8 and M=4, there is a loss of about 1.5 dB at BER=10^{-2} compared to figure 2. But it remains a gain of 1 dB compared to SU-BF. This gain vanishes at a mobile speed of 22 km/h, where we have similar performance of MU-SFTF and SU-BF. Besides, MU-SFTF experiences an error floor for $E_t/N_0 \geq 5$ dB. For higher speeds, SFTF looses its superiority and BF becomes more appropriate.

6. CONLUSION

We considered MC-CDMA DL transmission systems using an antenna array at the BS in order to mitigate MAI. Two different scenarios have been distinguished: a low-mobility TDD system with perfect instantaneous channel knowledge the BS and a high-mobility TDD or FDD scheme where only spatial information is known in advance. The presented results show that for both cases, there are appropriate transmitter strategies for efficient interference mitigation in the DL, which not only allow the transfer of the computational effort from the MTs to the BS, but also outperform a conventional system with high complex MUD at the MT.

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