Design of a Multi-Carrier CDMA Downlink with Different Transmit Antenna Array Strategies

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Abstract— **Multi-Carrier CDMA combines Orthogonal Frequency Division Multiplex with CDMA through symbol spreading in the frequency domain. Depending on the transmission scenario (indoor, outdoor urban or sub-urban) and the channel state information that may be available at the base station, several transmit antenna array strategies may be employed to mitigate the downlink multiple access interference while limiting the terminal complexity. Here, we present optimized system designs for several pre-filtering approaches such as joint space-frequency pre-filtering and space-only pre-filtering, *i.e.* beamforming.**

I. INTRODUCTION

Multi-Carrier CDMA (MC-CDMA) is currently considered as a versatile air interface for future mobile communication systems [1]-[4]. Based on Orthogonal Frequency Division Multiplex (OFDM), MC-CDMA applies symbol spreading in the frequency domain and, thus, provides a flexible multi-user access and a robustness to multipath propagation and cellular interference. The chips obtained after symbol spreading are generally transmitted on different sub-carriers of the OFDM, which provides frequency diversity but at the same time introduces a loss of orthogonality between different users' signals, *i.e.* Multiple Access Interference (MAI). A good trade-off between the negative impact of MAI and the positive effect of frequency diversity on the system performance can be achieved by an appropriate positioning of the chips of a given symbol on the sub-carriers with respect to the channel frequency selectivity.

To further mitigate MAI, conventional single-antenna systems often employ multi-user detection at the receiver side, which significantly increases the complexity and power consumption of the Mobile Terminal (MT). Alternatively, if the base station is equipped with multiple antennas, mitigation of MAI in the downlink can be obtained with appropriate transmit antenna array strategies. The additional spatial dimension is exploited to separate users’ signals already at the transmission so as to allow a higher system capacity even in case of low-complexity MTs. Antenna arrays have already been studied by several authors to improve the performance of MC-CDMA base stations. In [5], Kim et al. considered beamforming as a solution to improve the performance of uplink MC-CDMA communications. In [6]-[9], we demonstrated through several solutions that adaptive user separation with transmit antenna arrays is an interesting approach for the downlink as well. These latter schemes, which require partial or full knowledge of Channel State Information (CSI) prior to transmission, are particularly suited in Time Division Duplex (TDD) systems. Indeed, CSI may be obtained from channel estimation during the uplink time-slot and reused for transmission during the following downlink time-slot. Thus, the reliability of CSI for transmit filtering depends not only on the accuracy of uplink channel estimates but also on the mobility of the MT, which may introduce a mismatch between the available estimate and the real channel state at the instant of transmission. Therefore, we proposed in [7] different strategies of MC-CDMA downlink pre-filtering with multiple transmit antennas according to the propagation environment. For indoor communications and hotspots, space-frequency pre-filtering using instantaneous knowledge of the channel can be considered due to the relatively low mobility of MTs. In contrast for outdoor scenarios, where MTs may move quite fast, pre-filtering may only be applied in the space dimension, *i.e.* using spatial covariance knowledge for transmit beamforming. These solutions have been assessed with a spreading over the total number of sub-carriers.

In this paper, we focus on future mobile MC-CDMA communication systems, which require a spreading factor much lower than the total number of sub-carriers to limit the despreading complexity at the MT side. This offers a degree of freedom for placing the chips on the sub-carriers. We investigate how the selection of a transmit antenna array strategy optimized for a given communication environment impacts the overall system design and particularly the mapping of chips on the total set of OFDM sub-carriers. We propose practical combinations of pre-filtering and chip mapping for indoor and outdoor scenarios.

The remainder of this paper is organized as follows: A general description of the proposed MC-CDMA downlink transmission system with transmit antenna arrays at the base station is presented in section II. The different options for chip mapping are also introduced. In section III, the transmit antenna array optimization is presented for both indoor and outdoor environments and considerations are given on the influence of chip mapping in each case. Section IV evaluates the impact of chip mapping on the proposed transmit antenna array approaches with channel coding and section V concludes the paper.

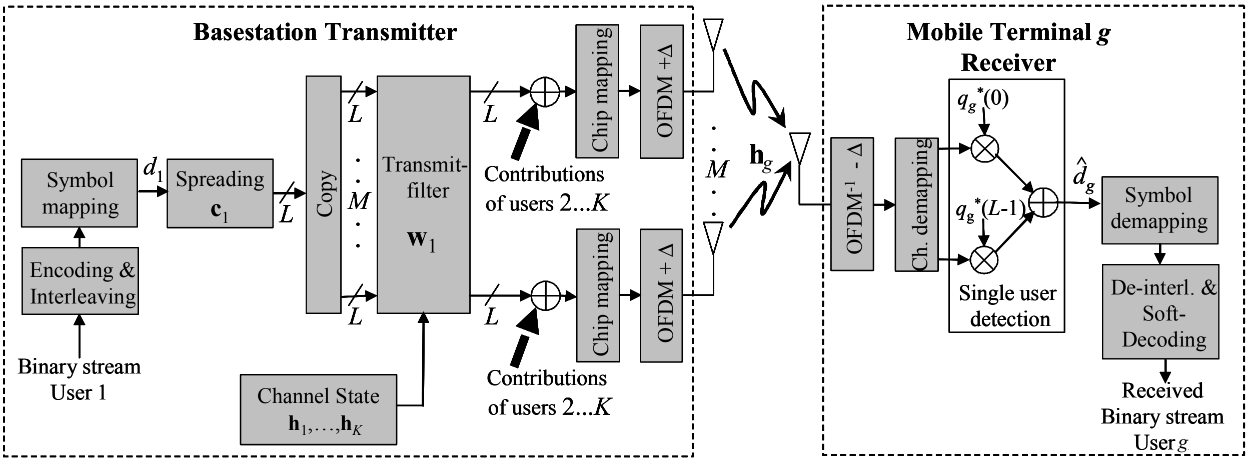

Figure 1: MC-CDMA system with a transmit antenna array.

II. GENERAL MC-CDMA SYSTEM

Figure 1 shows the structure of the proposed downlink system. At the Base Station (BS) side, the binary stream of each user k=1..K is first encoded, interleaved, and mapped to data symbols d_k, *e.g.* QPSK symbols. Each symbol d_k is then spread into L chips using a user-specific code vector $\mathbf{c}_k=[c_k(1),\ldots,c_k(L)]^T$ taken from an orthogonal, *e.g.* Walsh-Hadamard, set. These chips are copied to each of the M antenna branches. Both operations can be represented by extended vector $\vec{\mathbf{c}}_g=[\mathbf{c}_k^T,\ldots,\mathbf{c}_k^T]^T$ of size ML, which is a repetition of the code vector $\mathbf{c}_k$ of size L. $(.)^T$ denotes vector transposition. The so-obtained ML chips of each data symbol are weighted by the user-specific transmit filter vector $\mathbf{w}_k$ of size ML. Adaptive optimization of each $\mathbf{w}_k$ is based on CSI, which is assumed available for all users prior to transmission. Processing identically for each user in parallel, the contributions of all K users are then summed up chip-by-chip. The L cumulated pre-filtered chips of each antenna branch are finally mapped to L sub-carriers of the underlying OFDM, which is composed of $N_C=BL$ sub-carriers (B is an integer value).

We assume an ideal OFDM transmission with a proper dimensioning of both, the cyclic prefix Δ and the sub-carrier spacing. Thus, the system can absorb the multi-path spread of the propagation channel and avoid variations of the channel fading within the OFDM symbol duration. As a result, the channel between the M transmit antennas of the BS and the single antenna at the g-th MT on the L selected sub-carriers can be represented in the frequency domain by a vector $\mathbf{h}_g=[\mathbf{h}_{g,1}^T,\ldots,\mathbf{h}_{g,M}^T]^T$ of ML flat fading coefficients. Here, $\mathbf{h}_{g,m}$ is the vector of L fading coefficients representing the channel from antenna m to MT g.

For complexity reasons, we consider MTs equipped with a single antenna, which implicitly recombines the signals from the M transmit antennas in space. After cyclic prefix removal and OFDM demodulation, the chip de-mapping collects the L chip observations of each data symbol from the N_C sub-carriers. Then, assuming that Single-User Detection (SUD) is employed to further limit the receiver complexity [1], equalization and despreading explicitly recombine these observations using vector $\mathbf{q}_g=[q_g(1),\ldots,q_g(L)]^T$ of size L. The mathematical representation of signal recombination in space and frequency involves the expanded vector $\vec{\mathbf{q}}_g=[\mathbf{q}_k^T,\ldots,\mathbf{q}_k^T]^T$ of size ML. The resulting decision variable $\hat{d}_g$ for MT g is given by:

$$\hat{d}_g=\vec{\mathbf{q}}_g^H\left(\mathbf{h}_g\circ\mathbf{w}_g^*\circ\vec{\mathbf{c}}_g\right)d_g+\vec{\mathbf{q}}_g^H\left(\mathbf{h}_g\circ\sum_{k=1,k\neq g}^{K}\left(\mathbf{w}_k^*\circ\vec{\mathbf{c}}_k\right)d_k\right)+\mathbf{q}_g^H\mathbf{n}_g \quad (1)$$

where vector $\mathbf{n}_g$ gathers the noise samples on the L sub-carriers, the superscripts * and ^{H} are respectively the complex conjugate and conjugate transpose operators, and $\circ$ is the element-wise vector multiplication. Finally, soft de-mapping, de-interleaving, and soft-decoding yields the received binary stream for MT g.

From (1), the decision variable gathers the contributions of the desired signal, the MAI and the noise. Particularly, it has to be noted that the MAI, *i.e.* the loss of orthogonality originally installed by the spreading codes, results from the variations between the components of each vector $\mathbf{h}_g\circ\mathbf{w}_k$ combining channel fading and transmit filtering. Due to transmit filtering, this combination is specific to each user and a SUD technique aiming at restoring the orthogonality among users' signals is not suited. Therefore, we use Equal Gain Combining (EGC), *i.e.* phase equalization and despreading, as low complexity SUD scheme. The corresponding weight vector $\mathbf{q}_g$, which depends on the transmit filtering strategy, is detailed in section III.

Since the spreading factor L is generally much lower than the total number of used sub-carriers N_C in the OFDM [2]-[4], two basic options can be considered for chip mapping as illustrated in Figure 2. With an adjacent mapping (a), each vector $\mathbf{h}_{g,m}$ involves

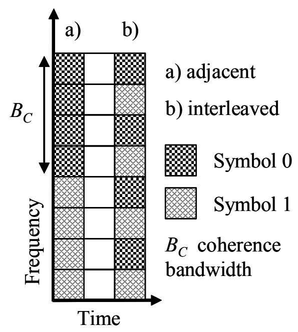

Figure 2 : Adjacent (a) and interleaved (b) chip mapping.

highly correlated fading coefficients. This minimizes MAI but induces a loss of frequency diversity after despreading. In contrast, with an interleaved mapping (b), the transmission benefits from a large frequency diversity thanks to the highly decorrelated components of each vector $\mathbf{h}_{g,m}$. This however introduces a loss of orthogonality among users' signals and thus an increased MAI.

Note that in both cases channel coding and bit interleaving also allows to benefit from frequency diversity to some extent. Especially adjacent chip mapping, which does not exploit frequency diversity at the chip level will benefit from this effect. It is thus essential to include channel coding in the comparison of both options.

III. TRANSMITTER OPTIMIZATION

Depending on the CSI available at the BS, two different strategies are investigated for transmit filtering using multiple antennas in the MC-CDMA downlink [7].

A. Strategy for Indoor Environments

In low-mobility indoor transmissions, the duration of uplink or downlink transmission slots can be chosen to be much shorter than the coherence time of the propagation channel. Thus, in a TDD system, the fading coefficients estimated by the BS during the reception of the uplink slot are still valid for optimizing the transmission during the following DL slot. Therefore, transmit filtering can be optimized assuming perfect knowledge of the channel coefficient vectors $\mathbf{h}_k$ for each MT k.

As a result, we propose Space Frequency Transmit Filtering (SFTF) *i.e.* a pre-filter optimized jointly in the space and frequency dimensions. Two different optimization criteria are employed. The Single-User (SU) criterion maximizes the Signal to Noise Ratio (SNR) in the decision variable. Here, the transmit filter is adapted to the channel of the considered MT g only. This results in a maximum ratio transmission whose transmit vector $\mathbf{w}_k$ is given by

$$\text{SU-SFTF:}\quad \mathbf{w}_k = \kappa_k \mathbf{h}_k^H \tag{2}$$

where the scalar κ_k ensures a constant transmit power.

The Multi-User (MU) criterion explicitly mitigates the interference created at other MTs by maximizing a modified Signal to Interference plus Noise Ratio (SINR) [7] defined for MT g as

$$\mathrm{SINR}_g = \frac{\left|\mathbf{w}_g^H \mathbf{h}_g\right|^2}{\mathbf{w}_g^H \left(\sum_{k=1,k\neq g}^{K} \mathbf{v}_{k,g}\mathbf{v}_{k,g}^H\right)\mathbf{w}_g + \sigma^2} \tag{3}$$

where $\mathbf{v}_{k,g} = \vec{\mathbf{c}}_k^* \circ \mathbf{h}_k \circ \vec{\mathbf{c}}_g$ and σ^2 is the noise variance per sub-carrier.

The solution that maximises (3) under the constraint of a normalised transmit power is given as

$$\text{MU-SFTF: } \mathbf{w}_k = \kappa_k \left(\sum_{k'=1,k'\neq k}^{K} \mathbf{v}_{k',k}\mathbf{v}_{k',k}^H + L\sigma^2 \mathbf{I}_{ML}\right)^{-1} \mathbf{h}_k^H \tag{4}$$

where the scalar κ_k ensures a constant transmit power and $\mathbf{I}_{ML}$ is the identity matrix of size ML.

As pre-equalization in the frequency dimension is included in (2) or (4), detection at MT g only consists in a despeading with, in theory, no need of channel estimation nor equalization, *i.e.* $\vec{\mathbf{q}}_g = \vec{\mathbf{c}}_g$. However, in practice, some slight remaining mismatches, *e.g.* due to imperfections of the RF front ends, may still require a very basic channel estimation and equalization process at the receiver side for reliability reasons.

The performance of SFTF may drastically depend on the way the chips are mapped. From (2), SU-SFTF applies to the transmitted signal an amplitude distortion equal to the amplitude distortion created by the channel. With several active users, this results in an increased MAI, which increases with the decorrelation of the frequency components of channel vectors $\mathbf{h}_{g,m}$ (and then $\mathbf{h}_g$), *i.e.* when interleaved chip mapping is used. On the contrary, as long as the number of users is not higher than the number of antennas times the spreading factor, an explicit MAI mitigation pre-filtering using (4) has sufficiently degrees of freedom to separate users' signals whatever the frequency correlation among these signals, *i.e.* whatever the chip mapping.

B. Strategy for Outdoor Environments

In outdoor transmissions with a potentially high mobility of MTs, the channel fading coefficients cannot be assumed constant over consecutive uplink and downlink slots. Therefore, even in a TDD system, only average channel knowledge can be exploited with enough reliability by the BS for pre-filtering. Hence, the pre-filtering proposed here is based on the spatial characteristics of users' signals, *e.g.* directions of departure (DODs) or covariance matrices, which vary on larger time scales, for beamforming (BF). In this paper, we assume that the BS can have knowledge of any of the following expressions related to the MxM spatial covariance matrice $\mathbf{R}_k$ [8]:

$$\mathbf{R}_{k,long} = \mathbf{E}\left[\mathbf{h}_k(\ell)\mathbf{h}_k^H(\ell)\right] \tag{5}$$

$$\mathbf{R}_{k,Nc} = \frac{1}{N_C}\sum_{\ell=1}^{N_C}\mathbf{h}_k(\ell)\mathbf{h}_k^H(\ell) \tag{6}$$

where $\mathbf{E}[x]$ denotes the statistical expectation of x and $\mathbf{h}_k(\ell)$ gather the M coefficients of the channel between the BS array and MT k on sub-carrier ℓ.

Based on any of the covariance matrices given in (5)-(6), we consider a SU eigen-beamforming approach [10] with the M-sized vector $\mathbf{w}'_k$ defined as

$$\text{BF:}\quad \mathbf{w}'_k = \kappa_k\, \mathrm{m\_eig}(\mathbf{R}_k) \tag{7}$$

where m_eig($\mathbf{X}$) denotes the principal eigenvector of matrix $\mathbf{X}$ and the scalar κ_k ensures a constant transmit power.

Since for a given MT the same beamforming is applied to the L chips involved in the transmission of a data symbol, the ML-sized vector $\mathbf{w}_k$ of (1) is just an L-times repetition of each component of $\mathbf{w}'_k$, *i.e.*

$$\mathbf{w}_k = \mathbf{w}'^T_k \begin{bmatrix} \underbrace{11\cdots1}_{L\ times} & 00\cdots0 & \cdots & 00\cdots0 \\ 00\cdots0 & 11\cdots1 & \cdots & 00\cdots0 \\ \vdots & \vdots & \ddots & \vdots \\ 00\cdots0 & 00\cdots0 & \cdots & 11\cdots1 \end{bmatrix} \tag{8}$$

Note the difference between the matrices given in (5) and (6): $\mathbf{R}_{k,long}$ is obtained by averaging over a duration much longer than the coherence time of the channel. In contrast, $\mathbf{R}_{k,Nc}$ is averaged over a much shorter duration than the coherence time of the channel and involves the total number N_C of sub-carriers used for the transmission of each OFDM symbol. As a result, $\mathbf{R}_{k,long}$ only depends on the spatial properties and the average power of each channel path whereas $\mathbf{R}_{k,Nc}$ is more representative of the current channel realization which involves not only the space properties but also the current complex fast fading coefficients of each path [8]. Therefore, applying (5) in (7) may thus achieve a better match between the BF and the current channel realization. However, when estimating $\mathbf{R}_{k,Nc}$ from the uplink slot for the next downlink slot transmission, a BF mismatch may be introduced in case of mobility. On the contrary, no mismatch is introduced by using $\mathbf{R}_{k,long}$ as it is only linked to the average channel realization which does not take fast fading variations into account.

Besides, by using (5) (respectively (6)) in (7), the same covariance matrix $\mathbf{R}_{g,Nc}$ (respectively $\mathbf{R}_{g,long}$) is used on the whole frequency axis of N_C sub-carriers, which leads to a BF efficiency independent of the chip mapping. However, the chip mapping trade-off MAI versus diversity obviously still exists.

In contrast to SFTF, pre-equalization is not included in BF so that both equalization and channel estimation are required at the MT side. Then, the SUD coefficient $q_g(\ell)$ is given by:

$$q_g(\ell) = c_g(\ell)\frac{\mathbf{w}_g'^H \mathbf{h}_g(\ell)}{\left|\mathbf{w}_g'^H \mathbf{h}_g(\ell)\right|} \qquad (9)$$

IV. NUMERICAL RESULTS

The proposed pre-filtering strategies, *i.e.* SFTF on one hand and BF on the other hand, have been assessed numerically in an indoor and an outdoor scenario, respectively. Single-user detection (EGC) is considered at the MT side, which results either in a pure despreading in case of SFTF or in the application of (9) for BF. UMTS convolutional channel coding is included in both cases. The system parameters are given in Table 1.

Table 1: System parameters.

Carrier frequency	5 GHz
Bandwidth	57.6 MHz
FFT-size / Used sub-carriers, N_C	1024 / 736
Carrier spacing, Δf	56.3 kHz
Symbol time / Cyclic prefix length	21.5 µs / 3.75 µs
Symbol alphabet	QPSK
Coding	UMTS convol., rate 2/3
Spreading factor, L	16
Uniform Linear Array	half-wavel. spaced, M=1,4
Indoor Channel Model	Spatially extended BRAN A, Max. delay 390 ns, AS = 120° No mobility
Outdoor Channel Model	Spatially extended BRAN E, Max. delay 1720 ns, AS = 30° Mobile speed 45 km/h

The channel models are based on the BRAN channels A (indoor) and E (outdoor) with 18 path components defined in [11]. We extended these time models to space-time models by allocating a DOD to each of the paths.

A. Indoor scenario

In typical indoor environments, the BS is surrounded by many obstacles which are responsible for multipath components spread over a large Angular Sector (AS). Here, we assume a uniform distribution of the Directions of Arrival (DOAs) of all users' signals within a sector of 120°. Thanks to the low mobility of MTs, perfect channel knowledge is assumed at the BS prior to transmission. Due to the stationarity of the channel, the DOAs of the uplink are equal to the DODs in the downlink.

Figure 3 shows the Bit Error Rate (BER) averaged over all active users as a function of the Signal to Noise Ratio (SNR) for a typical indoor scenario with the different chip mapping strategies. We compare the conventional single antenna system (M=1) with EGC single user detection (SUD) to the proposed system with joint transmit filtering in space and frequency dimensions (SFTF) and 4 antennas (M=4). Since channel coding is involved, we compare the performance at an operation point of BER=10^{-4}. As a reference, the conventional system for a single user (K=1) with interleaved mapping gains almost 2 dB compared to adjacent mapping due to a higher frequency diversity. However, for full system load (K=L=16), the interleaved scheme severely suffers from MAI. Hence, for single-user detection, adjacent mapping is preferable at higher loads, which confirms the results in [2].

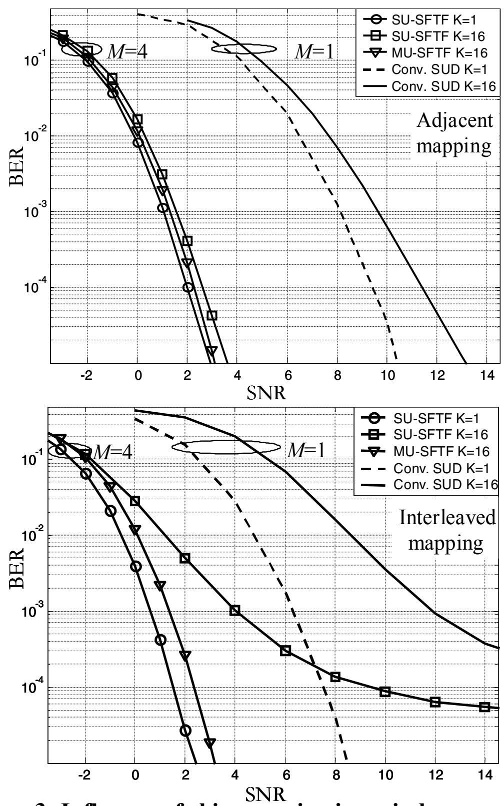

Figure 3: Influence of chip mapping in an indoor scenario.

With multiple transmit antennas, the situation is different. For K=1, the additional diversity gain obtained by chip interleaving is relatively small (around 0.5 dB). If only the complexity of a single-user approach (SU-SFTF) is tolerable, the system suffers again from a high MAI with interleaved mapping at full load since the SU criterion does not consider MAI. In contrast, with adjacent chip mapping, the loss at full load compared to K=1 is less than 1 dB, which is even lower than for the conventional system. As expected, the much more complex MU-SFTF performs efficient MAI mitigation whatever the chip mapping since $K<ML$. As a result, the overall system slightly benefits from an interleaved chip mapping (0.4 dB gain at BER=10^{-4}) by better exploiting the frequency diversity at both the despreading and the channel decoding stages.

B. Outdoor scenario

In a typical outdoor environment, the reflecting obstacles may be located far from the BS leading to a significantly smaller AS than in the indoor scenario. Here, we assume a uniform distribution of the DOAs in an AS of 30° around the main DOA, which itself is randomly chosen in the 120° sector for each MT. The velocity of MTs corresponds to a typical city traffic, *i.e.* 45 km/h. Note that even in the high mobility case, we can still assume that the DOAs of the uplink slot are identical to the DODs of the following downlink slot. Channel estimation at the MT receiver is assumed perfect. In contrast, at the BS, we assume that the covariance matrices defined in (5)-(6) are estimated from the uplink and used for downlink BF after a typical delay of 1 ms. Hence, the short-term spatial covariance matrix in (6) may suffer from a mismatch due to Doppler variations.

The BER performance of the proposed BF strategies with M=4 transmit antennas using (5) or (6) are represented in figure 4 according to the SNR for the different chip mapping options. In these cases, the low-complexity single-user detection as given by (9) is applied at the MT side. For the reference system, we consider a single-antenna transmission with minimum mean square error (MMSE) multi-user detection (MUD) at the receiver side. Full system load is assumed. It has to be noted that such a conventional system requires at the MTs a LxL matrix inversion (here 16x16) whose large complexity may not be tolerable in hardware implementations.

First, since BF is a single user technique where MAI cannot be mitigated, we find again that adjacent chip mapping outperforms interleaved chip mapping. When using a BF based on long-term average channel knowledge (5), the proposed system hardly achieves BER=10^{-2} with an interleaved chip mapping. With an adjacent chip mapping, performance is improved (BER=6.10^{-4}) but still with an error floor. On the contrary, if the covariance matrix given by (6) is used for BF, efficient users' signals

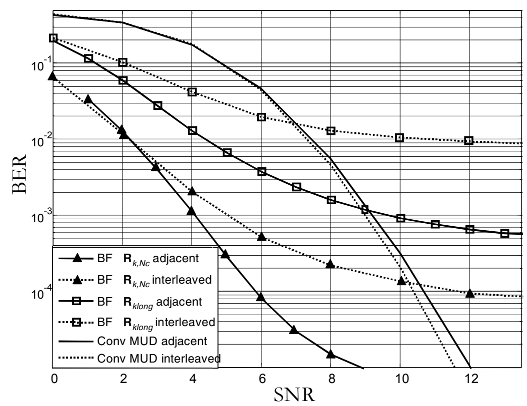

Figure 4: Influence of chip mapping in an outdoor scenario (*K=L=16*).

separation is achieved thanks to the better match of BF to the actual channel realization, even if the covariance is outdated due to the 45 km/h velocity of each MT. Thus, a SNR of 6 dB is sufficient to achieve BER=10^{-4} with an adjacent chip mapping. For comparison, the performance of the conventional system with single-transmit antenna at BS and very complex MMSE MUD at MT requires around 5 additional dB to achieve such a performance. Note that in this case, there is a slight benefit to use interleaved chip mapping due to the good robustness to MAI of MUD schemes, which can thus benefit from an increase of frequency diversity.

V. CONCLUSION

An MC-CDMA system with transmit antenna arrays can benefit from spatial diversity, which reduces the advantage of frequency diversity obtained by interleaved chip mapping. We presented several antenna array strategies suited to indoor and outdoor MC-CDMA systems and proposed appropriate chip mapping accordingly. In particular, we show that single-user transmit filtering techniques, *i.e.* SU-SFTF and Eigen-BF, combined with adjacent chip mapping are a good trade-off between performance and complexity.

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