SINR-BASED CHANNEL PRE-EQUALIZATION FOR UPLINK MULTI-CARRIER CDMA SYSTEMS

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Abstract - This paper investigates channel pre-equalization for the up-link of multi-carrier code division multiple access (MC-CDMA) mobile cellular systems. This technique solves the crucial problem of up-link channel estimation by keeping the high spectral efficiency of MC-CDMA with no need of pilot symbols. We study an optimisation criterion based on the maximization of the signal over interference plus noise ratio (SINR) at the base station while constraining the transmitted power by the mobile stations. The optimum algorithm is derived and a sub-optimum version is proposed to reduce the computational complexity in the mobile station. Simulation results on Rayleigh fading channels show that SINR-based pre-equalization systems achieve performance that are much better than systems using either conventional pre-equalization techniques at the transmitter or multi-user detection techniques at the receiver.

Keywords – MC-CDMA, Uplink, TDD, Pre-equalization.

I. INTRODUCTION

Among the different air interfaces that are candidate for the broadband component of a 4th generation mobile cellular system, the Multi-Carrier Code Division Multiple Access (MC-CDMA) technique [1][2][3] can provide high data rates in hostile propagation environments. In contrast with Direct Sequence CDMA (DS-CDMA) techniques, MC-CDMA uses the Orthogonal Frequency Division Multiplex (OFDM) modulation to perform spreading in the frequency domain. Thus, MC-CDMA, which is also referred to as OFDM-CDMA, benefits from OFDM characteristics such as high spectral efficiency and robustness against multi-path propagation while the CDMA features allow a flexible multiple access with good interference properties for cellular environments.

MC-CDMA received lots of interests from the research community during the past few years, especially for the down-link (DL), *i.e.* from Base Station (BS) to Mobile Terminals (MT), where the multiple access interference resulting from multi-path propagation can be mitigated by simple frequency domain equalization techniques [4]. In [5], MC-CDMA is shown to significantly increase the capacity of mobile cellular systems compared with 3rd generation technologies based on DS-CDMA.

For the up-link (UL), *i.e.* from MT to BS, MC-CDMA faces the difficulty that the receiver must perform channel equalization and thus channel estimation prior to despreading, at least to combine the frequency multiplex signal in phase as in equal gain combining (EGC) technique [4]. Then, in case of a pilot-based channel estimation, the pilots transmitted from distinct MTs cannot benefit from the spreading sequence orthogonality properties and must be transmitted at distinct time or frequency positions with very few interfering signals in order to make possible at the BS the estimation of the different channel coefficients. This principle drastically decreases the spectral efficiency of the system as the number of users increases.

Considering a Time Division Duplex (TDD) mode, another solution consists in performing channel pre-equalization at the transmitter side using the TDD channel reciprocity between alternative UL and DL transmission periods, *i.e.* slots. The principle is to use the same channel estimates obtained during a reception slot, for instance thanks to a pilot-based channel estimation, in order to pre-compensate the signal during the following transmission slot. This principle has already been proposed for MC-CDMA systems [6] [7] at the BS with the aim of canceling the multiple access interference by restoring the orthogonality among users' signal in the frequency domain. However, pre-equalizing each user's signal at the BS can only achieve such an orthogonality for a single MT. Indeed, an MT receives a multi-user signal where each user's contribution is corrupted by the same multi-path channel, which involves a specific pre-equalization for all MTs, but the multi-path channels from BS to other MTs are different. Besides, the pre-equalization technique in [6] [7] is based on a zero-forcing algorithm, which may lead to non-tolerable increase of the transmitted power in case of deep fading.

By contrast, this paper proposes new channel pre-equalization techniques for the UL of MC-CDMA/TDD systems that solve the channel estimation problem at the BS, ensures a high performance in terms of bit-error-rate, minimizes the transmitted power of MTs and induces a very low increase of computational complexity. Regarding 4th generation system constraints in terms of asymmetric traffic, pre-equalization is more suitable in UL than in DL: the higher number of DL slots in a duplex frame allows to benefit from more frequent channel estimates for pre-equalizing in the UL. After a description of the MC-CDMA/TDD transmission system in section II, this paper proposes a theoretical criterion based on a modified definition of the Signal-to-Interference plus Noise Ratio (SINR) in order to optimize the channel pre-equalization in section III and derives optimum as well as sub-optimum

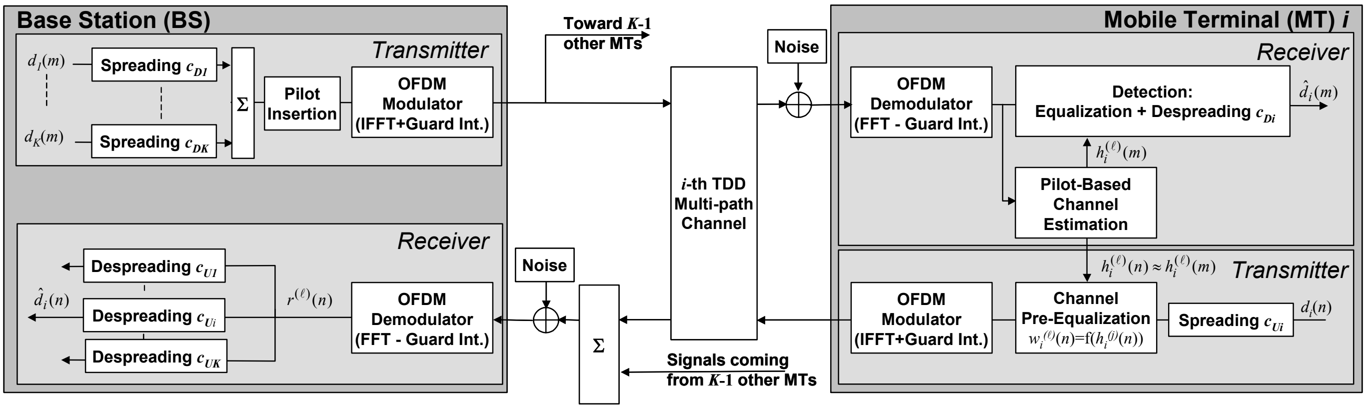

Fig. 1: Block diagram of the MC-CDMA/TDD transmission system between the base station and mobile terminal i (K users).

algorithms. Simulation results comparing the SINR-based pre-equalization techniques with other pre-equalization schemes and with advanced detection schemes at the BS with no pre-equalization in MTs are presented in section IV. Finally, section V gives concluding remarks.

II. MC-CDMA/TDD SYSTEM DESCRIPTION

We consider the MC-CDMA/TDD transmission system depicted in Fig. 1 with K active MTs. Both UL and DL are represented for MT i. During DL slot at time m, the BS generates for each MT k (k=1,…,K) the data symbol $d_k(m)$, which is spread with spreading sequence c_{Dk} over L_D chips. The chips of each user's signal are summed and common pilot chips are inserted at both edges of the DL slot to allow channel estimation at the MT side (Cf. Fig. 2).

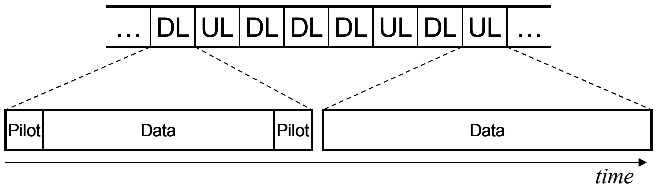

Fig. 2: Example of TDD frame with DL and UL slots.

The data and pilot chip sequences are then mapped on the N_C subcarriers of the OFDM multiplex, the subcarrier separation being equal to $\Delta f=1/T_s$, where T_s is the useful duration of the OFDM symbol. After OFDM modulation performed by Inverse Fast Fourier Transform (IFFT) and guard interval insertion, the multi-user OFDM signal is sent through the multi-path channels of the different MTs. The i-th MT performs guard interval removal and OFDM demodulation on the received signal corrupted by an additive white Gaussian noise (AWGN). As we assume that the guard interval is large enough to absorb the delay spread of the channel, the channel effect on subcarrier ℓ can be represented by a single complex fading coefficient $h_i^{(\ell)}(m)$. This assumes that the OFDM symbol duration is fixed to a value much smaller than the channel coherence time, to ensure negligible inter-carrier interference due to the channel non-stationnarity. Thanks to the common pilot chips, the MT channel estimator provides the detection stage with estimates of channel coefficients $h_i^{(\ell)}(m)$ to decode the data symbols in the DL.

If the channel is stationary over a period including time m dedicated to DL and time n dedicated to UL, the DL channel estimates $h_i^{(\ell)}(m)$ can be used at time n to pre-equalize the signal for UL transmission.

Thus, during UL slot at time n, the normalized data symbol $d_i(n)$ from the i-th MT is first spread over L_U chips with spreading sequence c_{Ui}, which is selected from a set of orthogonal spreading sequences, *e.g.* Walsh-Hadamard spreading sequences. These chips are mapped on the N_C subcarriers of the OFDM multiplex. For simplicity reasons, we consider the UL spreading factor L_U equal to N_C and normalized spreading sequence elements $c_{Uk}^{(\ell)}$ of user k at subcarrier ℓ taking value in {-1/√L_U, 1/√L_U}. On each subcarrier ℓ, the chip $c_{Ui}^{(\ell)}d_i(n)$ is multiplied by a pre-equalization coefficient $w_i^{(\ell)}(n)$, which is a function of channel estimates $h_i^{(j)}(n)$ (j=1,…,L_U). After IFFT and guard interval insertion, the single user OFDM signal is sent to the BS through the multi-path channel, which is assumed to have a channel transfer function close to the one estimated during DL transmission. We assume that the BS receives the sum of K synchronized MT signals, which were propagated through K distinct channels, corrupted by AWGN. A guard interval removal and a single FFT provide us with a multi-user received sample $r^{(\ell)}(n)$ on each subcarrier ℓ that is defined as:

$$r^{(\ell)}(n) = \sum_{k=1}^{K} h_k^{(\ell)}(n) w_k^{(\ell)*}(n) c_{Uk}^{(\ell)} d_k(n) + b^{(\ell)}(n) \tag{1}$$

where $b^{(\ell)}(n)$ is an AWGN component of variance σ^2 and the superscript * is the complex conjugate operator.

Denoting vectors as lower case underlined boldface and forgetting the time index for clarity reasons, the received sample vector can also be written as:

$$\underline{\mathbf{r}} = \left[r^{(1)},\cdots,r^{(L_U)}\right]^T = \sum_{k=1}^{K} \left(\underline{\mathbf{h}}_k \circ \underline{\mathbf{w}}_k^* \circ \underline{\mathbf{c}}_{Uk}\right) d_k + \underline{\mathbf{b}} \tag{2}$$

where $\circ$ defines the vector multiplication processed element by element, $\underline{\mathbf{h}}_k$, $\underline{\mathbf{w}}_k$, $\underline{\mathbf{c}}_{Uk}$ and $\underline{\mathbf{b}}$ are column vectors of size L_U containing respectively the channel response between BS and MT k, the pre-equalization coefficients computed at MT k, the spreading sequence c_{Uk} and the AWGN components.

As pre-equalization has been processed, the detection at the BS is reduced to a simple despreading by the i-th spreading sequence to yield an estimate of the transmitted data symbol:

$$\hat{d}_i = \underline{\mathbf{c}}_{Ui}^H \underline{\mathbf{r}} = \sum_{k=1}^{K} \underline{\mathbf{c}}_{Ui}^H \left(\underline{\mathbf{h}}_k \circ \underline{\mathbf{w}}_k^* \circ \underline{\mathbf{c}}_{Uk} \right) d_k + \underline{\mathbf{c}}_{Ui}^H \underline{\mathbf{b}} \tag{3}$$

where the superscript ^{H} denotes the conjugate-transpose complex operator.

By introducing b_i and vector $\underline{\mathbf{v}}_{ik}$ respectively defined as $b_i = \underline{\mathbf{c}}_{Ui}^H \underline{\mathbf{b}}$ and $\underline{\mathbf{v}}_{ik} = \underline{\mathbf{c}}_{Ui}^* \circ \underline{\mathbf{h}}_k \circ \underline{\mathbf{c}}_{Uk}$, equation (3) can be simplified in:

$$\begin{aligned} \hat{d}_i &= \underline{\mathbf{c}}_{Ui}^H \underline{\mathbf{r}} = \sum_{k=1}^{K} \left(\underline{\mathbf{w}}_k^H \underline{\mathbf{v}}_{ik} \right) d_k + b_i \\ &= \underbrace{\frac{1}{L_U} \left(\underline{\mathbf{w}}_i^H \underline{\mathbf{h}}_i \right) d_i}_{\text{desired signal}} + \underbrace{\sum_{k \neq i}^{K} \left(\underline{\mathbf{w}}_k^H \underline{\mathbf{v}}_{ik} \right) d_k}_{\text{multiple access interference}} + \underbrace{b_i}_{\text{noise}} \end{aligned} \tag{4}$$

Equation (4) emphasizes three terms: a signal from the desired MT with power $P_{Di}=(|\underline{\mathbf{w}}_i^H \underline{\mathbf{h}}_i|/L_U)^2$, a signal coming from other MTs that is Multiple Access Interference (MAI), and a noise with variance σ^2 since spreading sequences are properly normalized.

III. SINR-BASED PRE-EQUALIZATION

Different criteria can be considered to perform pre-equalization in UL. We may first aim at only performing a phase pre-equalization ($\underline{\mathbf{w}}_k=\underline{\mathbf{h}}_k/|\underline{\mathbf{h}}_k|$), which is mandatory to combine the signal in phase at the BS. Additionally, we may try to completely cancel the MAI term of equation (4), which consists in applying the well-known zero-forcing algorithm ($\underline{\mathbf{w}}_k=1/\underline{\mathbf{h}}_k^*$) at each MT. However, in case of deep fades that occur during mobile transmission, this technique may result in a large increase in transmitted power $P_{Tk}=|\underline{\mathbf{w}}_k|^2$. Therefore, criteria taking into account the transmitted power have to be investigated.

A. Modified SINR Criterion

It seems natural to design channel pre-equalization in order to maximize, for each MT i and for a fixed value of the transmitted power P_{Ti}, the SINR defined at the BS as:

$$\text{SINR}_i = \frac{P_{Di}}{\sum_{k \neq i} P_{\text{MAI}}(k/i) + \sigma^2} \tag{5}$$

where $P_{\text{MAI}}(k/i)$ defines the MAI power contribution of MT k on MT i.

However, for MT i, the sum of $P_{\text{MAI}}(k/i)$ terms depends on channel and pre-equalization coefficients of other MTs. Then, a pre-equalization algorithm maximizing equation (5) for MT i represents a global optimization problem involving not only $\underline{\mathbf{h}}_i$ but also all other channel responses $\underline{\mathbf{h}}_k$ (k=1,…,K, $k\neq i$). Whatever the complexity of this mathematical problem, it seems unlikely that MT i could know other channel response $\underline{\mathbf{h}}_k$ in an operational cellular system.

Assuming that all MTs use the same pre-equalization strategy and that all channel responses are statistically equal, we can consider that, on average, the interfering effect that undergoes the BS for the detection of MT i is equal to the interfering effect caused by MT i on the detection of other MTs. So, we can substitute the SINR of (5) by a modified SINR (m-SINR) defined for MT i at the BS as:

$$m\text{-SINR}_i = \frac{P_{Di}}{\sum_{k \neq i} P_{\text{MAI}}(i/k) + \sigma^2} \tag{6}$$

where the sum of $P_{\text{MAI}}(i/k)$ terms may be expressed as:

$$\sum_{k \neq i} P_{\text{MAI}}(i/k) = \underline{\mathbf{w}}_i^H \left(\sum_{k \neq i} \underline{\mathbf{v}}_{ki} \underline{\mathbf{v}}_{ki}^H \right) \underline{\mathbf{w}}_i = \underline{\mathbf{w}}_i^H \underline{\mathbf{\Phi}}_i \underline{\mathbf{w}}_i \tag{7}$$

where the $[L_U \times L_U]$ Hermitian matrix $\underline{\mathbf{\Phi}}_i=\Sigma_{k\neq i}\underline{\mathbf{v}}_{ki}\underline{\mathbf{v}}_{ki}^H$ represents the quadratic form associated with the generated interfering power of MT i.

From (7), m- SINR can finally be rewritten as:

$$m - \text{SINR}_i = \frac{1}{L_U^2} \frac{\left| \underline{\mathbf{w}}_i^H \underline{\mathbf{h}}_i \right|^2}{\underline{\mathbf{w}}_i^H \underline{\mathbf{\Phi}}_i \underline{\mathbf{w}}_i + \sigma^2} \tag{8}$$

Thus, in contrast with equation (5), maximizing equation (8) with a fixed transmitted power P_{Ti} for MT i only involves $\underline{\mathbf{h}}_i$ channel coefficients, which offers practical advantages.

B. m-SINR–Based Optimum Pre-Equalization

The optimization problem may be viewed as a maximization of equation (8) under the constraint $P_{Ti}=|\underline{\mathbf{w}}_i|^2=L_U$, which can be expressed under the same constraint as:

$$\max_{\underline{\mathbf{w}}_i} \frac{\left| \underline{\mathbf{w}}_i^H \underline{\mathbf{h}}_i \right|^2}{\underline{\mathbf{w}}_i^H \underline{\mathbf{\Phi}}_i \underline{\mathbf{w}}_i + \underline{\mathbf{w}}_i^H \underline{\mathbf{w}}_i \dfrac{\sigma^2}{L_U}} = \max_{\underline{\mathbf{w}}_i} \frac{\left| \underline{\mathbf{w}}_i^H \underline{\mathbf{h}}_i \right|^2}{\underline{\mathbf{w}}_i^H \left(\underline{\mathbf{\Phi}}_i + \dfrac{\sigma^2}{L_U} \underline{\mathbf{I}}_{Lu} \right) \underline{\mathbf{w}}_i} \tag{9}$$

where $\underline{\mathbf{I}}_{Lu}$ is the $[L_U \times L_U]$ identity matrix.

If $\underline{\mathbf{w}}_i$ is multiplied by a scalar β, expression (9) remains unchanged so that it is equivalent to find an optimum $\underline{\tilde{\mathbf{w}}}_i = \beta \underline{\mathbf{w}}_i$ that verifies:

$$\max_{\underline{\tilde{\mathbf{w}}}_i} \frac{\left| \underline{\tilde{\mathbf{w}}}_i^H \underline{\mathbf{h}}_i \right|^2}{\underline{\tilde{\mathbf{w}}}_i^H \left(\underline{\mathbf{\Phi}}_i + \dfrac{\sigma^2}{L_U} \underline{\mathbf{I}}_{Lu} \right) \underline{\tilde{\mathbf{w}}}_i} \quad \text{with} \quad \underline{\tilde{\mathbf{w}}}_i^H \underline{\mathbf{h}}_i = 1 \tag{10}$$

If $\underline{\mathbf{A}}_i=\underline{\mathbf{\Phi}}_i+(\sigma^2/L_U)\underline{\mathbf{I}}_{Lu}$, this problem is analogous to:

$$\min_{\underline{\tilde{\mathbf{w}}}_i} \underline{\tilde{\mathbf{w}}}_i^H \underline{\mathbf{A}}_i \underline{\tilde{\mathbf{w}}}_i \quad \text{subject to} \quad \underline{\tilde{\mathbf{w}}}_i^H \underline{\mathbf{h}}_i = 1 \tag{11}$$

whose solution can be obtained through the Lagrange function J:

$$J = \underline{\tilde{\mathbf{w}}}_i^H \underline{\mathbf{A}}_i \underline{\tilde{\mathbf{w}}}_i - \alpha\left(\underline{\tilde{\mathbf{w}}}_i^H \underline{\mathbf{h}}_i - 1\right) \tag{12}$$

where α is a Lagrange multiplier.
By setting the gradient vector of J equal to zero, we get:

$$\nabla J_{\underline{\tilde{\mathbf{w}}}_i^*} = \underline{\mathbf{A}}_i \underline{\tilde{\mathbf{w}}}_i - \alpha \underline{\mathbf{h}}_i = \underline{0} \Leftrightarrow \underline{\tilde{\mathbf{w}}}_i = \alpha \underline{\mathbf{A}}_i^{-1} \underline{\mathbf{h}}_i \tag{13}$$

where α is chosen in order to respect the linear constraint of (11).
Finally, the optimum vector of channel pre-equalization coefficients according to the m-SINR criterion is:

$$\underline{\mathbf{w}}_i = \lambda\left(\underline{\mathbf{\Phi}}_i + \frac{\sigma^2}{L_U}\mathbf{I}_{Lu}\right)^{-1} \underline{\mathbf{h}}_i \tag{14}$$

where λ is chosen such that $|\underline{\mathbf{w}}_i|^2 = L_U$.
Applying equation (14) in MTs ensures maximizing the m-SINR at the BS defined by (6) for each MT with a constrained transmitted power. However, it requires solving a $[L_U \times L_U]$ linear system, whose complexity might be intolerable even for low spreading factors since MTs have drastic computational restrictions due to power consumption limitations. Therefore, we propose a simplification of the optimum m-SINR-based pre-equalization technique.

C. m-SINR–Based Sub-Optimum Pre-Equalization

Considering the matrix $\underline{\mathbf{\Phi}}_i$ defined in (7), we can rewrite it as:

$$\begin{aligned}\underline{\mathbf{\Phi}}_i &= \sum_{k \neq i} \underline{\mathbf{v}}_{ki} \underline{\mathbf{v}}_{ki}^H = \sum_{k \neq i} \left(\underline{\mathbf{c}}_{Uk}^* \circ \underline{\mathbf{h}}_i \circ \underline{\mathbf{c}}_{Ui}\right)\left(\underline{\mathbf{c}}_{Uk}^* \circ \underline{\mathbf{h}}_i \circ \underline{\mathbf{c}}_{Ui}\right)^H \\ &= Diag\left(h_i^{(\ell)} c_{Ui}^{(\ell)}\right) \cdot \sum_{k \neq i}\left(\underline{\mathbf{c}}_{Uk}^* \underline{\mathbf{c}}_{Uk}^T\right) \cdot Diag\left(c_{Ui}^{(\ell)*} h_i^{(\ell)*}\right)\end{aligned} \tag{15}$$

We observe that the diagonal terms of matrix $\Sigma_{k \neq i}(c_{Uk}^* \, c_{Uk}^T)$ are predominant compared with off-diagonal terms and are equal to $(K\text{-}1)/L_U$. So, we propose to approximate $\underline{\mathbf{\Phi}}_i$ by $\underline{\mathbf{\Psi}}_i$, which is defined as:

$$\begin{aligned}\underline{\mathbf{\Psi}}_i &= Diag\left(h_i^{(\ell)} c_{Ui}^{(\ell)}\right) \cdot Diag\left(\frac{K-1}{L_U}\right) \cdot Diag\left(c_{Ui}^{(\ell)*} h_i^{(\ell)*}\right) \\ &= Diag\left(\frac{(K-1) \cdot \left|h_i^{(\ell)}\right|^2}{L_U^{\,2}}\right)\end{aligned} \tag{16}$$

Coming back to equation (14) and replacing $\underline{\mathbf{\Phi}}_i$ by $\underline{\mathbf{\Psi}}_i$, we get a vector of pre-equalization coefficients which approaches optimality according to the m-SINR criterion:

$$\begin{aligned}&\underline{\mathbf{w}}_i = \lambda\left(\underline{\mathbf{\Psi}}_i + \frac{\sigma^2}{L_U}\mathbf{I}_K\right)^{-1} \cdot \underline{\mathbf{h}}_i \\ \Leftrightarrow\ & w_i^{(\ell)} = \frac{\mu \cdot h_i^{(\ell)}}{(K-1) \cdot \left|h_i^{(\ell)}\right|^2 + L_U \sigma^2}\end{aligned} \tag{17}$$

where μ is chosen such that $|\underline{\mathbf{w}}_i|^2 = L_U$.
Compared to equation (14), applying equation (17) in MTs induces a very low complexity. It only requires the knowledge of channel coefficients $\underline{\mathbf{h}}_i$ that were already computed during DL channel estimation period, the number of active users K and the noise variance σ^2 at the BS. K and σ^2 can be sent by BS as control information. However, σ^2 can also be set to an average value.
It can be demonstrated that (17) is equivalent to (14) for full load transmission (K=L_U).

IV. SIMULATION RESULTS

In this section, we evaluate the performance of the proposed pre-equalization algorithms, *i.e.* the optimum m-SINR-based pre-equalization (OMSP) and the sub-optimum m-SINR-based pre-equalization (S-OMSP) algorithms, by comparison with UL systems using other pre-equalization algorithms, namely the phase pre-equalization (PP) and the zero-forcing pre-equalization (ZFP). We also consider UL systems without pre-equalization in MTs but with optimum linear multi-user detection at the BS based on the minimization of the mean square error of the decision variable, assuming that some spectrally efficient multi-user channel estimation at the BS would be feasible. This latter technique will be referred to as MMSE MUD. Besides, the single user transmission with optimum detection will be considered as a performance bound.
As a simulation scenario, we consider an UL synchronized transmission using Walsh-Hadamard spreading sequences of length L_U=64. Then, the system can tolerate up to 64 distinct users, which corresponds to a full load transmission. The channel estimates required at the MT for pre-equalization or at the BS for multi-user detection are assumed to be perfect.
Performance of the transmission is evaluated over a theoretical Rayleigh fading channel according to the required transmitted signal-to-noise ratio Eb/N0, where Eb is the energy per bit transmitted for each user and N0 is the noise power spectral density.

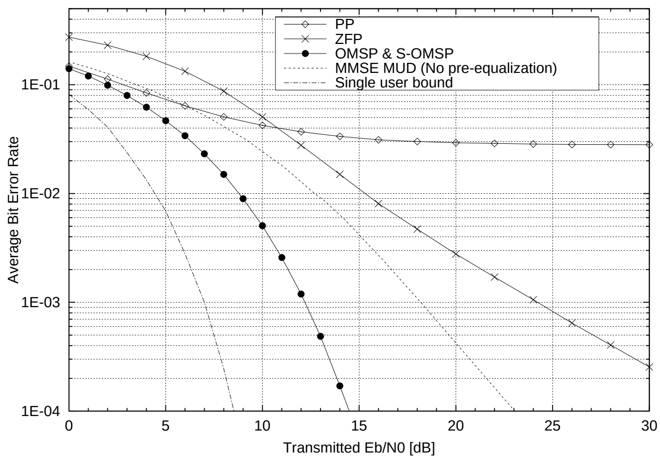

Fig. 3: Influence of the UL technique on the transmission performance (Full load).

Fig. 3 represents the BER performance of different UL transmission techniques as a function of the transmitted Eb/N0 for a full load transmission. At first, the PP scheme cannot compensate the large interference due to full load: an error floor is experienced at BER=3.10^{-2}. For low Eb/N0, the PP algorithm outperforms the ZFP scheme since ZFP allocates almost all the transmitted power to the subcarriers that encounter deep fades. Thus, the received power of the desired signal is too low compared with the noise level, even if the interference is completely cancelled. As mentioned in section III, OMSP and S-OMSP achieve identical performance in full-load. A 12 dB gain is shown compared with ZFP to achieve BER=10^{-3}. This large gain comes from the better allocation of the transmitted power to the different subcarriers, which is reduced for highly faded subcarriers by taking into account the expectable noise variance σ^2 at the BS (see equations (14) and (17)). Compared with the single user bound, the performance loss of OMSP and S-OMSP algorithms is only 5 dB at BER=10^{-3}, which is excellent for an UL full load transmission. Finally, OMSP and S-OMSP in MTs outperform MMSE-MUD at the BS with a 6 dB gain at BER=10^{-3}. In other words, processing pre-equalization in MTs achieves better performance than performing an optimum multi-user detection at the BS, independently of the spectral efficiency gain that can be obtained in addition by avoiding a pilot-based channel estimation in the BS.

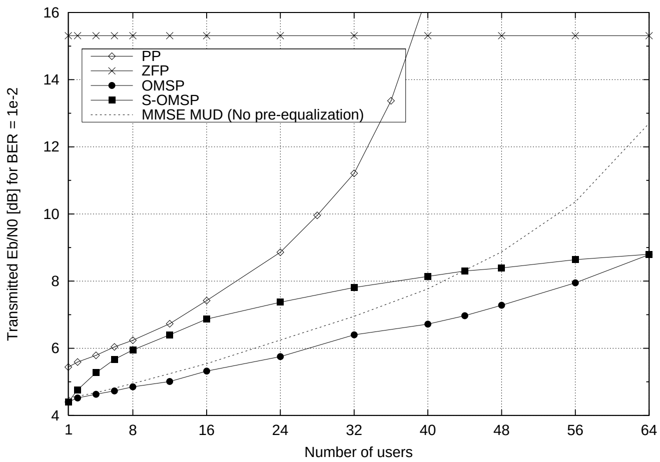

Fig. 4: Influence of the number of users on the transmission performance (BER=10^{-2}).

Fig.4 represents the transmitted Eb/N0 required by the different techniques to achieve a BER=10^{-2}, which may be considered as a realistic operating point before any channel decoding process. Curves are given according to the number of users. First, the ZFP scheme confirms its bad performance independently of the number of users. By contrast, the PP algorithm shows a good behavior for low loads that degrades as the load increases so that BER=10^{-2} cannot be achieved from 44 users. OMSP achieves the best performance with a required Eb/N0 ranging between 4.2 dB for one user and 8.8 dB for full load. The maximum loss of performance due to S-OMSP sub-optimality is 1.5 dB, which is reasonable regarding its very low complexity. Finally, even optimum MMSE MUD at the BS experiences performance loss compared to optimum pre-equalization in MTs whatever the number of users. This proves the need of pre-equalization to increase the performance of UL MC-CDMA transmission systems.

V. CONCLUSION

Channel pre-equalization is particularly suitable for the reverse link of MC-CDMA systems using TDD mode as it achieves excellent performance whatever the system load and can solve the channel estimation issue, this issue being very restrictive in UL MC-CDMA systems.

From an implementation point of view, UL channel pre-equalization can be interpreted as a transfer of complexity from BS to MTs, which goes against the current trend to limit the computational complexity in terminals. In contrast, we propose simple pre-equalization techniques that aim at maximizing the SINR at the BS by optimizing the allocation of the MT transmitted power among the different subcarriers. Thus, we show that the required transmitted power to achieve a given operating point is much lower for UL systems based on pre-equalization in MTs than for UL systems based on optimum multi-user detection in the BS, which leads to reduction of MT battery consumption.

Finally, it has already been proved that MC-CDMA technique is a good candidate for the DL broadband component of the 4th generation mobile cellular system [5]. From this paper, we show that the UL associated component can also be based on MC-CDMA technology.

ACKNOWLEDGEMENTS

This study was sponsored by the French RNRT (National Network for Telecommunications Research), in the framework of the project TURBO-ACCESS.

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