A Turbo SDMA Receiver for Strongly Nonlinearly Distorted MC-CDMA Signals

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***Abstract*—This paper considers the use of SDMA (Space Division Multiple Access) techniques for the uplink of MC-CDMA systems (MultiCarrier - Coded Division Multiple Access) where the transmitted signals face strong nonlinear distortion effects. The signal transmitted by each MT (Mobile Terminal) is submitted to a nonlinear operation consisting on a clipping device, followed by a frequency-domain filtering operation, so as to reduce the envelope fluctuations and PMEPR (Peak-to-Mean Envelope Power Ratio) while maintaining the spectral occupation of conventional MC-CDMA signals. At the BS (Base Station), an iterative receiver employing multiple antennas jointly performs turbo MUD (MultiUser Detection) and the estimation and cancellation of nonlinear distortion effects.**

Our performance results show that the proposed receiver structures allows good performance, that can be very close to the ones obtained with linear transmitters, even for high system load and/or when a very low-PMEPR MC-CDMA transmission is intended for each MT.

***Index Terms*—Space Division Multiple Access (SDMA), Multicarrier-Coded Division Multiple Access (MC-CDMA), turbo equalization, multiuser detection, nonlinear effects.**

I. INTRODUCTION

SDMA (Space Division Multiple Access) techniques employ multiple antennas to substantially increase the system capacity [1]. In this paper we consider the use of SDMA techniques for the uplink of MC-CDMA systems (MultiCarrier - Coded Division Multiple Access). As with other multicarrier schemes, MC-CDMA signals have strong envelope fluctuations and high PMEPR values (Peak-to-Mean Envelope Power Ratio), which lead to amplification difficulties. For this reason, it is desirable to reduce the envelope fluctuations of the transmitted signals. This is particularly important for the uplink transmission, since an efficient, low-cost power amplification is desirable at the MT (Mobile Terminal). Several techniques have been recommended for reducing the envelope fluctuations of multicarrier signals (see [2] and references within). A promising approach is to employ clipping techniques, combined with a frequency-domain filtering so as to reduce the envelope fluctuations of the transmitted signals while maintaining the spectral occupation of conventional schemes [2]. However, the nonlinear distortion effects can be severe when a low-PMEPR transmission is intended [2], [3].

As with other CDMA schemes, since the transmission over time-dispersive channels destroys the orthogonality between users, an FDE (Frequency-Domain Equalizer) is required before the despreading operation [4]. To avoid significant noise enhancement for channels with deep in-band notches, the FDE is usually optimized under an MMSE criterion (Minimum Mean-Squared Error) [4]. However, as FDE/MMSE does not perform an ideal channel inversion we are not able to fully orthogonalize the different spreading codes of an MC-CDMA signal[1]. This means that we can have severe interference levels, especially for fully loaded systems and/or when different powers are assigned to different spreading codes. To improve the performance several turbo MUD receivers were proposed for conventional CDMA systems [5], [6], as well as MC-CDMA [7]. A promising technique for MC-CDMA with nonlinear transmitters was proposed in [8], where nonlinear distortion effects are iteratively estimated and compensated. However, for low SNR (Signal-to-Noise Ratio) the error decisions might lead to error propagation effects, since errors in the estimation of nonlinear distortion effects can preclude an efficient cancellation. This is particularly serious for high system load and/or when no space diversity is used [8]. This is especially important when the spreading factor is small and/or if we decrease the clipping level, to reduce further the PMEPR of the transmitted signals.

To reduce error propagation effects in the typical region of operation we use channel decoder outputs in the feedback loop, in a turbo-like fashion (a similar approach was proposed for OFDM schemes [9]). We define an iterative receiver that jointly performs turbo MUD as well as estimation and cancellation of nonlinear distortion effects, taking in account the distortion's frequency distribution that is inherent to the transmitted signals.

This paper is organized as follows: the system characterization considered here is described in Sec. II. In Sec. III we describe the iterative receivers proposed in this paper. Sec. IV presents a set of performance results and Sec. V is concerned with the conclusions of the paper.

II. SYSTEM CHARACTERIZATION

We consider the uplink transmission of MC-CDMA signals employing frequency-domain spreading. We have an SDMA architecture depicted in Fig. 1, corresponding to a MIMO (Multiple-Input, Multiple-Output) system with P users (MTs), transmitting independent data blocks, and L receive antennas at the BS (Base Station). It is assumed that each MT has

[1]An MMSE FDE might also lead to noise correlations, creating unwanted dependencies between the decisions made for each data symbol associated to a given spreading code. This is usually not a problem in coded systems, provided that a suitable interleaving is employed between the channel encoder and the symbol mapper.

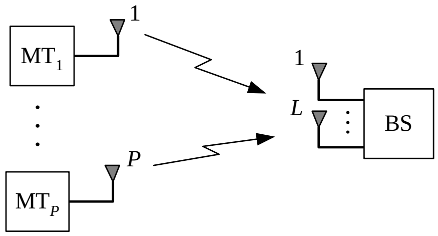

Fig. 1. System characterization.

a single transmit antenna. The coded bits are interleaved and mapped, leading to the block of data symbols to be transmitted by the pth MT $\{A_{k,p}; k = 0, 1, \ldots, M-1\}$, where M is the number of data symbols for that MT. The frequency-domain block to be transmitted by the pth MT is $\{S_{k,p}; k = 0, 1, \ldots, N-1\}$ where $N = KM$, with K denoting the spreading factor. The frequency-domain symbols are given by $S_{k,p} = \xi_p C_{k,p} A_{k \bmod M,p}$, ($x \bmod y$ is the reminder of the division of x by y) where ξ_p is an appropriate weighting coefficient that accounts for the propagation losses and $\{C_{k,p}; k = 0, 1, \ldots, N-1\}$ is the corresponding spreading sequence[2] (a pseudo-random spreading is assumed, with $C_{k,p}$ belonging to a QPSK constellation; without loss of generality, it is assumed that $|C_{k,p}| = 1$). The transmitter structure depicted in Fig. 2 is based on the nonlinear signal processing schemes proposed in [2] for reducing the PMEPR of OFDM signals. Within that transmitter, $N' - N$ zeros are added to the original frequency-domain block (i.e., $N' - N$ idle subcarriers), followed by an IDFT operation so as to generate a sampled version of the time-domain MC-CDMA signal, with an oversampling factor $M_{Tx} = N'/N$. Each time-domain sample is submitted to a nonlinear device so as to reduce the envelope fluctuations on the transmitted signal. In this paper we assume that the nonlinear device is an ideal envelope clipping with clipping level s_M, i.e., the output samples are

$$s^c_{n,p} = \begin{cases} s'_{n,p}, & s'_{n,p} \leq s_M \\ s_M, & s'_{n,p} > s_M \end{cases} \tag{1}$$

with $\{s'_{n,p}; n = 0, 1, ..., N'-1\}$ denoting the input samples. After a DFT operation the clipped signal is then submitted to a frequency-domain filtering procedure, through the set of multiplying coefficients $G_k, k = 0, 1, \ldots, N'-1$, in order to reduce the out-of-band radiation levels inherent to the nonlinear operation.

Using Price's Theorem [10], it is shown in [3] that the frequency-domain block to be transmitted by the pth MT $\{S^{Tx}_{k,p} = S^C_{k,p} G_k; k = 0, 1, \ldots, N'-1\}$ can be decomposed into the sum of two uncorrelated components: a useful one, proportional to $\{S_{k,p}; k = 0, 1, \ldots, N-1\}$, and a nonlinear self-interference one, i.e., $S^{Tx}_{k,p} = \alpha_p S_{k,p} G_k + D_{k,p} G_k$, where α_p is a scalar factor, defined in [2], [3], $G_k, k = 0, 1, ..., N'-1$, are the frequency-domain filtering coefficients in order to reduce the out-of-band radiation levels inherent to the nonlinear operation and $\{D_{k,p}; k = 0, 1, ..., N'-1\}$ is the frequency-domain block of nonlinear self-interference components associated to the pth MT. Unless otherwise stated, we will assume that $G_k = 1$ for the N in-band subcarriers and 0 for the $N' - N$ out-of-band subcarriers, i.e., the spectral occupation of the transmitted signal is similar to the spectral occupation of conventional MC-CDMA signals. In this case $S^{Tx}_{k,p} = \alpha_p S_{k,p} + D_{k,p}$, for the N in-band subcarriers and 0 otherwise. It can be shown that $D_{k,p}$ is approximately Gaussian-distributed, with zero mean; moreover, $E[D_{k,p} D^*_{k',p}]$ can be computed analytically, as described in [2], [3].

[2]This corresponds to uniformly spreading the chips associated to a given symbol within the transmission band, i.e., to employ a rectangular interleaver with dimensions $K \times M$.

III. Receiver Structure

A. Linear Transmitters

As usual, it is assumed that the length of the CP (Cyclic Prefix) is higher than the length of the overall channel impulse response. The received time-domain block associated to the lth diversity branch, after discarding the samples associated to the CP, is $\{y^{(l)}_n; n = 0, 1, \ldots, N-1\}$. The corresponding frequency-domain block (i.e., the length-N DFT (Discrete Fourier Transform) of the block $\{y^{(l)}_n; n = 0, 1, \ldots, N-1\}$) is $\{Y^{(l)}_k; k = 0, 1, \ldots, N-1\}$.

Let us consider first a linear transmitter. In this case, the frequency-domain block transmitted by the pth MT is $\{S^{Tx}_{k,p} = S_{k,p}; k = 0, 1, \ldots, N'-1\}$ and

$$\begin{aligned} Y^{(l)}_k &= \sum_{p=1}^{P} S_{k,p}\ H^{Ch(l)}_{k,p} + N^{(l)}_k = \\ &= \sum_{p=1}^{P} A_{k \bmod M,p}\ C_{k,p}\ \xi_p\ H^{Ch(l)}_{k,p} + N^{(l)}_k = \\ &= \sum_{p=1}^{P} A_{k \bmod M,p}\ H^{(l)}_{k,p} + N^{(l)}_k \end{aligned} \tag{2}$$

with $H^{Ch(l)}_{k,p}$ denoting the channel frequency response between the pth MT and the lth diversity branch, at the kth subcarrier, $N^{(l)}_k$ the corresponding channel noise and $H^{(l)}_{k,p} = \xi_p H^{Ch(l)}_{k,p} C_{k,p}$. To detect the kth symbol of the pth MT we will use the set of subcarriers $\Psi_k = \{k, k+M, \ldots, k+(K-1)M\}$.

By defining $\mathbf{Y}(k) = [\mathbf{Y}^{(1)}(k) \cdots \mathbf{Y}^{(L)}(k)]^T$, with $\mathbf{Y}^{(l)}(k)$ denoting the column vector with the received samples associated to the set of frequencies Ψ_k, for the lth antenna, and $\mathbf{A}(k) = [A_{k \bmod M,1} \ldots A_{k \bmod M,P}]^T$, we have

$$\mathbf{Y}(k) = \mathbf{H}^T(k)\mathbf{A}(k) + \mathbf{N}(k) \tag{3}$$

($(\cdot)^T$ denote the transpose matrix), where $\mathbf{N}(k) = [\mathbf{N}^{(1)}(k) \cdots \mathbf{N}^{(L)}(k)]^T$, with $\mathbf{N}^{(l)}(k)$ denoting the column vector with the noise samples associated to the set of frequencies Ψ_k, for the lth antenna. In (3), $\mathbf{H}(k)$ is the overall channel matrix associated to $\mathbf{A}(k)$, i.e., $\mathbf{H}(k) = [\mathbf{H}^{(1)}(k) \cdots \mathbf{H}^{(L)}(k)]$, with $\mathbf{H}^{(l)}(k)$ denoting a $(P \times K)$ matrix, with lines associated to the different MTs and columns associated to the set of frequencies Ψ_k, for the lth antenna.

Our receiver, depicted in Fig. 3, can be described as follows. For a given iteration, the detection of $\mathbf{A}(k)$ employs L feedforward filters (one for each receive antennas) and P feedback loops. The feedforward filters are designed to minimize the

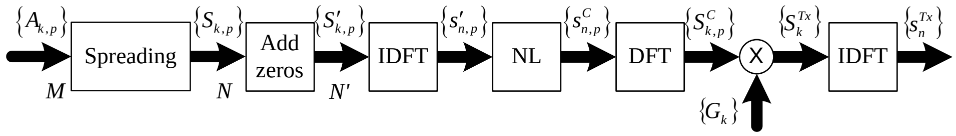

Fig. 2. Transmitter model considered in this paper.

MAI (Multiple Access Interference) that cannot be cancelled by the feedback loops. This means that our receiver can be regarded as an iterative multiuser detector with PIC (Parallel Interference Cancelation). For the first iteration we do not have any information about the MT's symbols and the receiver reduces to a linear multiuser receiver.

For each iteration, the samples associated to $\mathbf{A}(k)$, $\tilde{\mathbf{A}}(k)$ are given by

$$\tilde{\mathbf{A}}(k) = \mathbf{F}^T(k)\mathbf{Y}(k) - \mathbf{B}^T(k)\overline{\mathbf{A}}(k) \tag{4}$$

where $\tilde{\mathbf{A}}(k)$ is defined as $\mathbf{A}(k)$, $\mathbf{F}(k) = [\mathbf{F}^{(1)}(k) \cdots \mathbf{F}^{(L)}(k)]^T$ is the matrix of the feedforward filters' coefficients, with $\mathbf{F}^{(l)}(k)$ denoting a $(P \times K)$ matrix, with lines associated to the different MTs and columns associated to the set of frequencies Ψ_k, for the lth antenna, and $\mathbf{B}(k)$ is a $(P \times P)$ matrix with the feedback filters' coefficients. $\overline{\mathbf{A}}(k)$ is a vector with the LLRs of $\mathbf{A}(k)$ corresponding to the "soft decisions" from the SISO (Soft-In, Soft-Out) channel decoder, from the previous iteration. The SISO block, that can be implemented as defined in [11], provides the LLRs (LogLikelihood Ratios) of both the "information bits" and the "coded bits". The input of the SISO block are LLRs of the "coded bits" at the multiuser detector.

It can be shown that (see [8]), when the transmitted symbols are selected from a QPSK constellation under a Gray mapping rule, the complex value of the LLRs for the in-phase and quadrature bits of $A_{k,p} = A^I_{k,p} + jA^Q_{k,p}$ are given by

$$\overline{A}_{k,p} = \tanh\left(\frac{L^I_{k,p}}{2}\right) + j\tanh\left(\frac{L^Q_{k,p}}{2}\right) \tag{5}$$

(the generalization to other cases is straightforward), where $L^I_{k,p} = 2\tilde{A}^I_{k,p}/\sigma^2_p$ and $L^Q_{k,p} = 2\tilde{A}^Q_{k,p}/\sigma^2_p$, are the LLRs of the "in-phase bit" and the "quadrature bit", associated to $A^I_{k,p} = \mathrm{Re}\{A_{k,p}\}$ and $A^Q_{k,p} = \mathrm{Im}\{A_{k,p}\}$, respectively, with

$$\sigma^2_p = \frac{1}{2}E[|A_{k,p} - \tilde{A}_{k,p}|^2] \approx \frac{1}{2}E[|\hat{A}_{k,p} - \tilde{A}_{k,p}|^2], \tag{6}$$

and $\hat{A}_{k,p}$ denoting the "hard decisions" associated to $\tilde{A}_{k,p}$.

The hard decisions $\hat{A}^I_{k,p} = \pm 1$ and $\hat{A}^Q_{k,p} = \pm 1$ are defined according to the signs of $L^I_{k,p}$ and $L^Q_{k,p}$, respectively; $\rho^I_{k,p} = \tanh\left(|L^I_{k,p}|/2\right)$ and $\rho^Q_{k,p} = \tanh\left(|L^Q_{k,p}|/2\right)$ can be regarded as the reliabilities associated to the "in-phase" and "quadrature" bits of the kth symbol of the pth MT. For the first iteration, $\rho^I_{k,p} = \rho^Q_{k,p} = 0$ and $\overline{A}_{k,p} = 0$. We can also define the blockwise reliability

$$\rho_p = \frac{1}{M}\sum_{k=0}^{M-1}\frac{E[A^*_{k,p}\hat{A}_{k,p}]}{E[|A_{k,p}|^2]} = \frac{1}{2M}\sum_{k=0}^{M-1}(\rho^I_{k,p} + \rho^Q_{k,p}). \tag{7}$$

The feedforward and feedback filters' coefficients matrixes, $\mathbf{F}(k)$ and $\mathbf{B}(k)$, respectively, are chosen so as to maximize the $SNIR$ (Signal-to-Noise plus Interference Ratio), at the decoder's input, for all MTs, at a particular iteration. For the pth MT, the $SNIR$ is defined as

$$SNIR_p = \frac{E[|A_{k,p}|^2]}{\sigma^2_p}. \tag{8}$$

It can be shown that, by employing the Lagrangian's multipliers method and after some straightforward but lengthy manipulation, the optimum feedforward coefficients in the MMSE sense can be written as

$$\mathbf{F}(k) = \mathbf{F}^I(k)\mathbf{\Gamma}^{-1} \tag{9}$$

where $\mathbf{\Gamma} = \mathrm{diag}(\gamma_1, \cdots, \gamma_P)$, with

$$\gamma_p = \frac{1}{M}\sum_{k'\in\Psi_k}\sum_{l=1}^{L}F^{I(l)}_{k',p}H^{(l)}_{k',p}. \tag{10}$$

and

$$\mathbf{F}^I(k) = [\mathbf{H}^H(k)(\mathbf{I}_P - \mathbf{P}^2)\mathbf{H}(k) + \beta\mathbf{I}_{KL}]^{-1}\mathbf{H}^H(k), \tag{11}$$

($(\cdot)^H$ denotes the Hermitian matrix and $\mathbf{I}_X$ being the X-by-X identity matrix), with $\mathbf{P} = \mathrm{diag}(\rho_1, \ldots, \rho_P)$ and $\beta = E[|N^{(l)}_k|^2]/E[|A_{k,p}|^2]$ is the inverse of the signal-to-noise ratio of the pth MT (SNR_p). Equation (11) is the Wiener-Hopf equation with the covariance of the available measurements given by the matrix being inverted. γ_p is a normalization factor that can be regarded as the average overall channel frequency response, for the pth MT, after the feedforward filters' coefficients $\{F_{k',p}, k' = k, k+M, ..., k+(K-1)M\}$.

The optimum feedback coefficients are given by

$$\mathbf{B}(k) = \mathbf{H}(k)\mathbf{F}(k) - \mathbf{I}_P. \tag{12}$$

If we do not have data estimates for the different MTs, $\rho_p = 0$ ($p = 1, 2, \ldots, P$), and the feedback coefficients are zero. Therefore, (4) reduces to

$$\tilde{\mathbf{A}}(k) = \mathbf{F}^T(k)\mathbf{Y}(k), \tag{13}$$

which corresponds to the linear receiver.

It can be shown that the optimum feedforward coefficients can be written in the form

$$\mathbf{F}(k) = \mathbf{H}^H(k)\mathbf{V}(k), \tag{14}$$

apart a normalization factor as in (9), with $\mathbf{V}(k)$ given by

$$\mathbf{V}(k) = [(\mathbf{I}_P - \mathbf{P}^2)\mathbf{H}(k)\mathbf{H}^H(k) + \beta\mathbf{I}_P]^{-1}. \tag{15}$$

The computation of the feedforward coefficients from (14) is simpler than the direct computation, from (11), especially when $P < KL$ (i.e., when the system is not fully loaded).

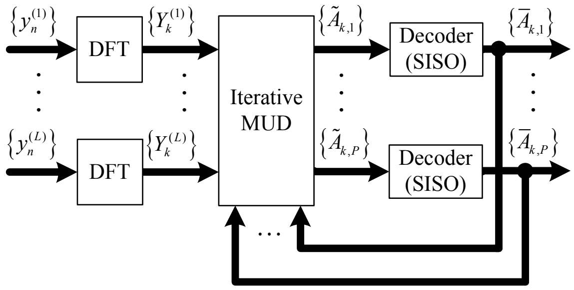

Fig. 3. Iterative receiver for a linear transmitter.

B. Nonlinear Transmitters

It was shown in [9] that we can improve significantly the performance of OFDM schemes submitted to nonlinear devices by employing a receiver with iterative cancelation of nonlinear distortion effects. This concept can be extended to MC-CDMA, leading to the receiver structure of Fig. 4. The basic idea behind this receiver is to use an estimate of the nonlinear self-distortion $\{\overline{D}_{k,p}; k = 0, 1, \ldots, N-1\}$ provided by the preceding iteration to remove the nonlinear distortion effects in the received samples. Therefore, the received frequency-domain block associated to the lth diversity antenna, $\{Y_k^{(l)}; k = 0, 1, \ldots, N-1\}$, is replaced by the corrected block $Y_k^{Corr(l)} = Y_k^{(l)} - \sum_{p=1}^{P} H_{k,p}^{Ch(l)} D_{k,p},\ k = 0, 1, \ldots, N-1$. The remaining of the receiver is similar, with the optimum feedforward coefficients given by (9), but with

$$\mathbf{F}^I(k) = \Big[\mathbf{H}^H(k)\mathbf{U}^2\left(\mathbf{I}_P - \mathbf{P}^2\right)\mathbf{H}(k) + \beta_k \mathbf{I}_{KL} + \eta_k \mathbf{H}^{Ch^H}(k)\mathbf{H}^{Ch}(k)\Big]^{-1}\mathbf{H}^H(k), \quad (16)$$

where $\mathbf{U} = \mathrm{diag}(\alpha_1, ..., \alpha_P)$, $\mathbf{H}^{Ch}(k)$ denote the channel frequency response matrix, defined as $\mathbf{H}(k)$,

$$\beta_k = \frac{E[|N_k^{(l)}|^2]}{E[|\alpha_p A_{k,p}|^2] + E[|D_{k,p}^{eq}|^2]} \quad (17)$$

is, again, the inverse of the signal (plus nonlinear self-distortion)-to-noise ratio of the pth MT, and

$$\eta_k = \frac{\mathrm{diag}(E[|D_{k,p}^{eq}|^2])}{E[|\alpha_p A_{k,p}|^2]}, \quad k = 0, M-1, ..., KM-1 \quad (18)$$

with

$$D_{k,p}^{eq} = \frac{1}{K}\sum_{k' \in \Psi_k} D_{k',p} C_{k',p} \quad (19)$$

represents the inverse of the signal-to-distortion ratio of the pth MT.

Again, it can be shown that the optimum feedforward coefficients can be written as (14) but, in this case, with

$$\mathbf{V}(k) = \Big[(\mathbf{I}_P - \mathbf{P}^2)\mathbf{H}(k)\mathbf{H}^H(k)\mathbf{U}^2 + \beta_k \mathbf{I}_P + \eta_k \mathbf{H}^{Ch}(k)\mathbf{H}^{Ch^H(k)}\Big]^{-1}. \quad (20)$$

For a given iteration, $\{\overline{D}_{k,p}; k = 0, 1, \ldots, N-1\}$ can be estimated from $\{\overline{A}_{k,p}; k = 0, 1, \ldots, M-1\}$ as follows: $\{\overline{A}_{k,p}; k = 0, 1, \ldots, M-1\}$ is re-spread to generate the "average block to be transmitted" $\{\overline{S}_{k,p}; k = 0, 1, \ldots, N-1\}$;

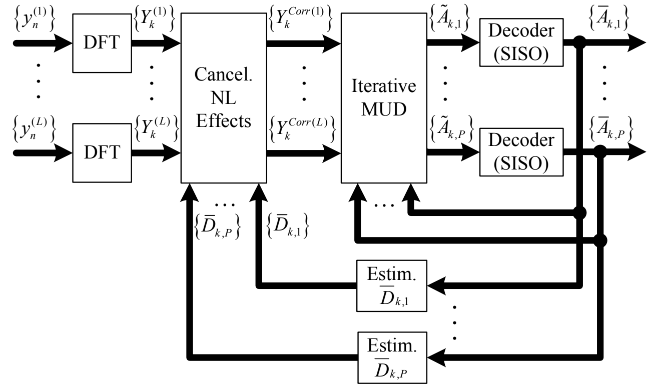

Fig. 4. Iterative receiver with cancelation of nonlinear distortion effects.

$\{\overline{S}_{k,p}; k = 0, 1, \ldots, N-1\}$ is submitted to a replica of the nonlinear signal processing scheme employed in the pth transmitter so as to form the "average transmitted block" $\{\overline{S}_{k,p}^{Tx}; k = 0, 1, \ldots, N-1\}$; $\overline{D}_{k,p}$ is given by $\overline{D}_{k,p} = \overline{S}_{k,p}^{Tx} - \alpha_p \overline{S}_{k,p}$ (naturally, for the first iteration, $\overline{D}_{k,p} = 0$).

IV. Performance Results

In this section we present a set of performance results concerning the iterative receiver structures proposed in this paper for the uplink of MC-CDMA systems with frequency-domain spreading. The spreading factor is $K = 4$ and we have $M = 64$ data symbols for each user, corresponding to blocks with length $N = KM = 256$, plus an appropriate CP. QPSK constellations, with Gray mapping, are employed. To reduce the envelope fluctuations of the transmitted signals (and the PMEPR) while maintaining the spectral occupation of conventional MC-CDMA schemes, each MT employs the clipping techniques combined with a frequency-domain filtering proposed in [2] (the power amplifiers are assumed to be linear for the (reduced) dynamic range of the envelope fluctuations of the transmitted signals). The receiver (i.e., the BS) knows the characteristics of the PMEPR-reducing signal processing technique employed by each MT.

We consider the power delay profile type C for the HIPERLAN/2 (HIgh PERformance Local Area Network) [12], with uncorrelated Rayleigh fading for the different MTs and for the different paths. The duration of the useful part of the block is 4μs and the CP has duration 1.25μs. We consider coded BER performances under perfect synchronization and channel estimation conditions[3]. We consider the well-known rate-1/2, 64-state convolutional code with generators $1+D^2+D^3+D^5+D^6$ and $1+D+D^2+D^3+D^6$. A random intrablock interleaving of the coded bits is assumed before the mapping procedure. The SISO decoder is implemented using the Max-Log-MAP approach. We have $\xi_p = 1$ for all MTs, i.e., we have a perfect power control. At the BS we have $L = 1, 2,$ or 4 uncorrelated receive antennas, for diversity purposes.

[3]It should be noted that perfect time synchronization between the blocks associated to different MTs is not required since some time mismatches can be absorbed by the CP.

TABLE I
PMEPR OF THE TRANSMITTED SIGNALS.

s_M/σ	PMEPR (dB)	
	Just clipping	Clipping and filtering†
∞	8.4	8.4
1	1.0	4.3
0.5	0.5	4.1

†Filtering operation is required if we want to maintain the spectral occupation of conventional MC-CDMA signals.

Let us first consider that we have $P = 4$ MTs and a normalized clipping level, identical for all MTs, of $s_M/\sigma = 1$. This allows the PMEPR values shown in Table I [2]. Fig. 5 and 6 show the corresponding coded BER performances for each iteration (averaged over all MTs), when the receiver's feedback loop uses the soft decisions from the multiuser detector and the channel decoder, respectively, as well as the performance for a linear transmitter. From this figures, it is clear that the performance of the linear receiver (first iteration) is very poor, with high irreducible error floors due to the nonlinear distortion effects. This is especially serious for low clipping levels and/or when the system is fully loaded ($P = KL$). As we increase the number of iterations improve significantly the performances, that can be close to the ones obtained with linear transmitters. We can also see that the use of turbo receivers (with channel decoder outputs in the feedback loop) leads to better performance relative to that with multiuser detector outputs in the feedback loop.

Let us assume now that we have $P = KL = 8$ MTs, corresponding to a fully loaded scenario and a very low clipping level of $s_M/\sigma = 0.5$. Fig. 7 shows the corresponding coded BER performances as well as the performance for $s_M/\sigma = 1$ and for a linear transmitter. Once again, with the increase of the iterations number, the performances are close to the ones obtained with linear transmitters, even for high system load and/or when a low-PMEPR is intended for each MT, although with a little performance degradation. As we can see from Fig. 8, the increase of the receiver's antennas from $L = 2$ to $L = 4$ allows a significant improve in performance results.

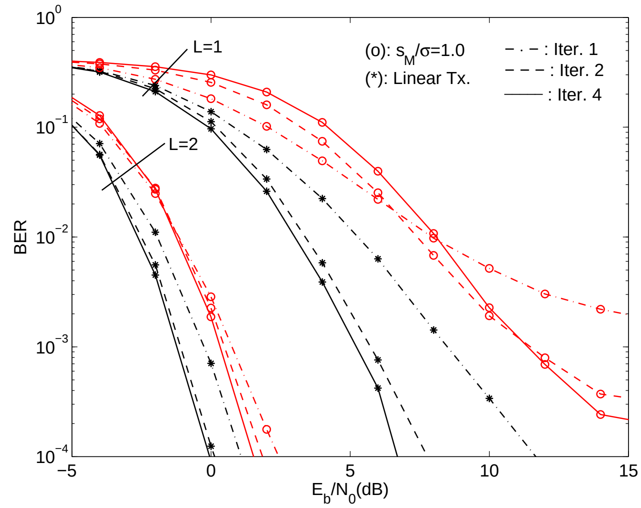

Fig. 5. Average coded BER performance with the soft decisions from the multi-user detector in the receiver's feedback loop, for iterations 1, 2 and 4 (better performances as we increase the number of iterations), for $P = 4$ and $L = 1$ and 2, when linear and nonlinear transmitters are considered.

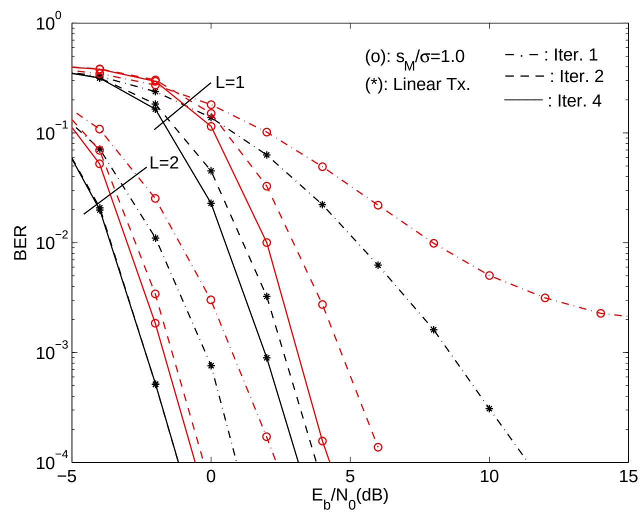

Fig. 6. As in Fig. 5, but with a turbo receiver (channel decoder outputs in the feedback loop).

V. CONCLUSIONS

In this paper we considered the use of SDMA techniques for the uplink transmission of MC-CDMA signals employing strongly nonlinear transmitters. We proposed an iterative receiver structure employing multiple antennas that combine turbo MUD and estimation and cancellation of the nonlinear distortion effects that are inherent to the transmitted signals.

Our performance results show that the proposed receiver structures allows good performance, that can be very close to the ones obtained with linear transmitters, even for high system loads and/or when a low-PMEPR MC-CDMA transmission is intended for each MT.

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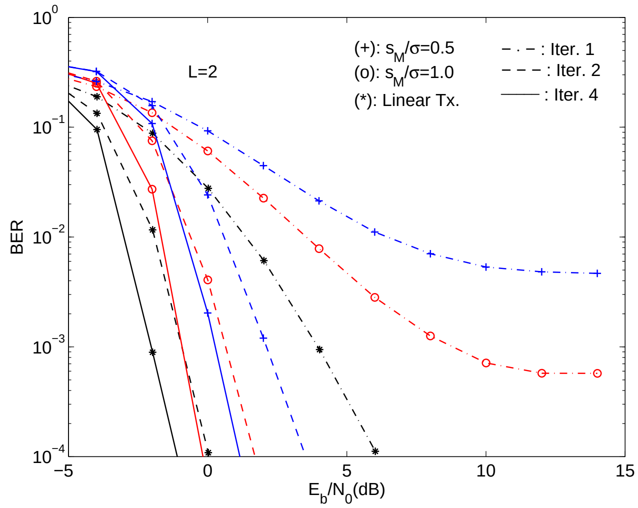

Fig. 7. Average coded BER performance for iterations 1, 2 and 4, for $P = 8$ and $L = 2$, when linear and nonlinear transmitters are considered.

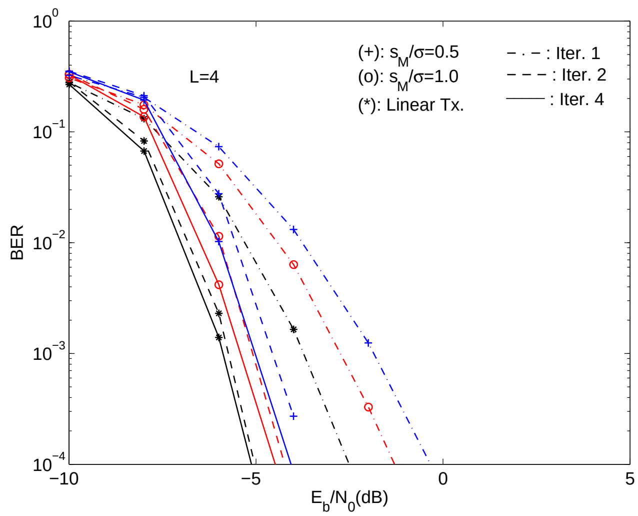

Fig. 8. As in Fig. 7, but with $L = 4$.

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