Turbo Multiuser Detection for MC-CDMA Signals with Strongly Nonlinear Transmitters

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***Abstract*—In this paper we consider the uplink transmission in MC-CDMA (MultiCarrier - Coded Division Multiple Access) systems. Since MC-CDMA signals are OFDM-like multicarrier signals, they have high envelope fluctuations and a high PMEPR (Peak-to-Mean Envelope Power Ratio) which leads to amplification difficulties. To reduce the envelope fluctuations of the transmitted signals, while maintaining the spectral efficiency, the MC-CDMA signal associated to each MT (Mobile Terminal) is submitted to a clipping device, followed by a frequency-domain filtering operation. However, the nonlinear distortion effects can be high when an MC-CDMA transmitter with reduced envelope fluctuations is intended.**

In this paper, we define an iterative receiver that jointly performs a turbo-MUD (MultiUser Detection) and the estimation and cancellation of the nonlinear distortion effects.

The set of simulation results presented shows that the proposed receiver structure allows good performances, very close to the linear receiver ones, even for high system load and/or when a low-PMEPR is intended for each MT. [1]

I. INTRODUCTION

MC-CDMA schemes (MultiCarrier - Coded Division Multiple Access) have been proposed for future broadband wireless systems [1]. The basic idea behind these schemes is to combine a CDMA scheme with an OFDM modulation (Orthogonal Frequency Division Multiplexing) [2]. Therefore, we can have high transmission rates over severely time-dispersive channels without the need of complex receiver implementations. Since the spreading is made in frequency domain, the time synchronization requirements are much lower than with conventional direct sequence CDMA schemes. Moreover, the diversity effect inherent to the spreading allows good uncoded performances, as well as good performances with high code rates.

The transmission over time-dispersive channels destroys the orthogonality between spreading codes. For this reason, an FDE (Frequency-Domain Equalizer) is required before the de-spreading operation [3], [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) [3], [4]. Since an MMSE FDE does not perform an ideal channel inversion, we are not able to fully orthogonalize the different spreading codes of an MC-CDMA signal. 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 iterative receiver for multicode MC-CDMA signals was proposed in [8], based on the IB-DFE (Iterative block Decision Feedback Equalizer) concept [9], [10]. That receiver allows significant performance improvements as we increase the number of iterations, especially for fully loaded systems and high spreading factors.

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 [11]-[14]. 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 [14]. However, the nonlinear distortion effects can be severe when a low-PMEPR transmission is intended [14], [15]. To improve the performances, we can employ the receivers proposed in [16], [17] where the nonlinear distortion effects are iteratively estimated and compensated [18]. 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 its cancellation. This is particularly serious for high system load [18]. 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.

In this paper we consider the uplink transmission in MC-CDMA systems. To allow an efficient power amplification, the PMEPR-reducing techniques of [14] are adopted by each MT. The BS (Base Station) has several receive antennas, so as to reduce the transmit power requirements of each MT. To reduce error propagation effects in the typical region of operation we

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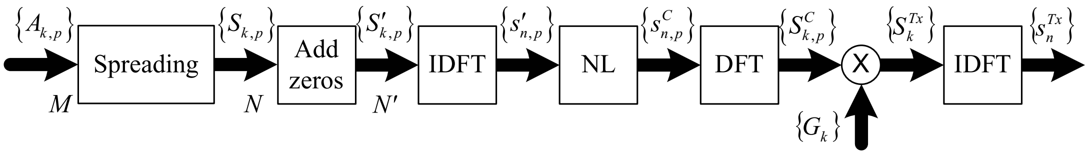

Fig. 1. Transmitter model considered in this paper (NL- nonlinear device).

use channel decoder outputs in the feedback loop, in a turbo-like fashion (a similar approach was proposed for OFDM schemes [19]). We define an iterative receiver that jointly performs a turbo-MUD and the estimation and cancellation of nonlinear distortion effects that are inherent to the transmitted signals.

This paper is organized as follows: the transmitter structure considered in this paper 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. Transmitter Structure

In this paper we consider the uplink transmission in MC-CDMA systems employing frequency-domain spreading. 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 and M the number of data symbols for that MT. The frequency-domain symbols are given by $S_{k,p} = \xi_p C_{k,p} A_{k\mathrm{Mod}M,p}$, ($x$ Mod 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, $\{A_{k,p}; k = 0,1,\ldots,M-1\}$ is the block of data symbols associated to the pth MT 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 is depicted in Fig. 1, which is based on the nonlinear signal processing schemes proposed in [14] for reducing the PMEPR of OFDM signals. It is shown in [15] 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 useful and nonlinear self-interference components: $S^{Tx}_{k,p} = \alpha_p S_{k,p} G_k + D_{k,p} G_k$, where $G_k, k = 0,1,...,N'-1$, are the frequency-domain filtering coefficients, in order to reduce the out-of-band radiation levels inherent to the nonlinear operation, and α_p defined in [14], [15]. Throughout this paper we assume that $G_k = 1$ for the N in-band subcarriers and 0 for the $N'-N$ out-of-band subcarriers. In this case

$$S^{Tx}_{k,p} = \begin{cases} \alpha_p S_{k,p} + D_{k,p}, & k \text{ in band} \\ 0, & k \text{ out of band.} \end{cases} \tag{1}$$

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 [14], [15].

[2] This corresponds to uniformly spread 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. We will assume that the BS has L receive antennas and 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

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

with $H^{Ch(l)}_{k,p}$ denoting the channel frequency response between the pth MT and the lth diversity branch, at the kth sub-carrier, $N^{(l)}_k$ 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 line vector with the received samples associated to the set of frequencies Ψ_k, for the lth antenna, and $\mathbf{A}(k) = [A_{k\mathrm{Mod}M,1} \ldots A_{k\mathrm{Mod}M,P}]^T$, 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 line 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.

This receiver can be regarded as an iterative multiuser detector with PIC (Parallel Interference Cancelation). The receiver can be described as follows. For a given iteration, the detection of $\mathbf{A}(k)$ employs the structure depicted in Fig.

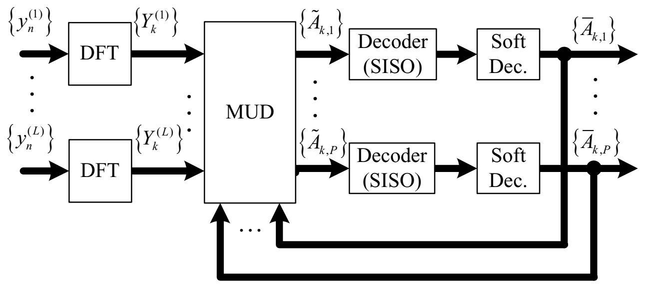

Fig. 2. Iterative receiver for a linear transmitter.

2, where we have L feedforward filters (one for each receive antennas) and P feedback loops. The feedforward filters are designed to minimize the MAI (Multiple Access Interference) that cannot be cancelled by the feedback loops. 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

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

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. $\overline{\mathbf{A}}(k)$, also defined as $\mathbf{A}(k)$, is the "soft decision" of $\mathbf{A}(k)$ from the SISO (Soft-In, Soft-Out) channel decoder, from the previous iteration. The SISO block, that can be implemented as defined in [20], 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.

In [18] it is shown that, assuming that the transmitted symbols are selected from a QPSK constellation under a Gray mapping rule (the generalization to other cases is straightforward), $\overline{A}_{k,p}$ is given by

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

where $L^I_{k,p} = 2\tilde{A}^I_{k,p}/\sigma_p^2$ and $L^Q_{k,p} = 2\tilde{A}^Q_{k,p}/\sigma_p^2$, 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

$$\sigma_p^2 = \frac{1}{2}E[|A_{k,p} - \tilde{A}_{k,p}|^2] \approx \frac{1}{2M}\sum_{k=0}^{M-1} E[|\hat{A}_{k,p} - \tilde{A}_{k,p}|^2], \tag{6}$$

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

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

It can be shown that the optimum feedforward coefficients in the MMSE sense can be written as

$$\mathbf{F}(k) = \mathbf{F}^I(k)\mathbf{\Gamma}^{-1} \tag{8}$$

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

$$\gamma_p = \frac{1}{M}\sum_{k' \in \Psi_k}\sum_{l=1}^{L} F^{(l)}_{k',p} H^{(l)}_{k',p} \tag{9}$$

and

$$\mathbf{F}^I(k) = [\mathbf{H}^H(k)(\mathbf{I}_P - \mathbf{P}^2)\mathbf{H}(k) + \beta\mathbf{I}_{KL}]^{-1}\mathbf{H}^H(k), \tag{10}$$

($(\cdot)^H$ 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 $\beta = E[|N^{(l)}_k|^2]/E[|A_{k,p}|^2]$.

The optimum feedback coefficients are given by

$$\mathbf{B}(k) = \mathbf{H}(k)\mathbf{F}(k) - \mathbf{I}_P. \tag{11}$$

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

$$\tilde{\mathbf{A}}(k) = \mathbf{F}^T(k)\mathbf{Y}(k), \tag{12}$$

which corresponds to the linear receiver.

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

$$\mathbf{F}(k) = \mathbf{H}^H(k)\mathbf{C}(k), \tag{13}$$

apart a normalization factor as in (8), with $\mathbf{C}(k)$ given by

$$\mathbf{C}(k) = [(\mathbf{I}_P - \mathbf{P}^2)\mathbf{H}(k)\mathbf{H}^H(k) + \beta\mathbf{I}_P]^{-1}. \tag{14}$$

The computation of the feedforward coefficients from (13) is simpler than the direct computation, from (10), especially when $P < K$ and/or $L > 1$.

B. Nonlinear Transmitters

It was shown in [16], [17] 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. 3. 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, $\{Y^{(l)}_k; k = 0, 1, \ldots, N-1\}$, is replaced by the corrected block $\{Y^{Corr(l)}_k; k = 0, 1, \ldots, N-1\}$, where

$$Y^{Corr(l)}_k = \frac{1}{\alpha_p}\left(Y^{(l)}_k - \sum_{p=1}^{P} H^{Ch(l)}_{k,p}\overline{D}_{k,p}\right). \tag{15}$$

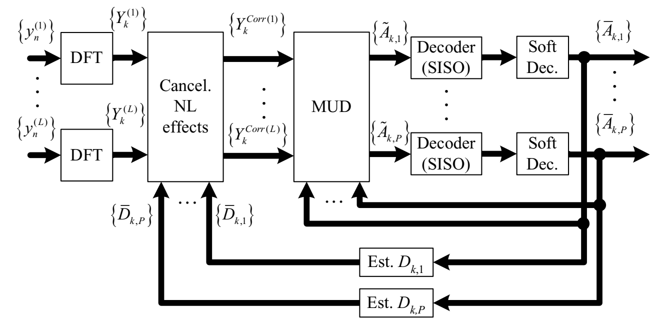

Fig. 3. Iterative receiver with cancelation of nonlinear distortion effects.

The remaining of the receiver is similar, but with β given by

$$\beta = \frac{E[|N_k^{(l)}|^2]}{E[|\alpha_p A_{k,p}|^2] + E[|D_{k,p}^{eq}|^2]}, \tag{16}$$

where

$$D_{k,p}^{eq} = \frac{1}{K} \sum_{k' \in \Psi_k} D_{k',p} C_{k',p}. \tag{17}$$

For the first iteration, the expectations of (16) can easily be obtained by using the analytical approach of [14], [15]; for the remaining iterations they have to be obtained by simulation. However, in most cases of interest we cancel a significant part of the nonlinear self-distortion at the first iteration; therefore, after the first iteration, we can ignore $E[|D_{k,p}^{eq}|^2]$ in (16).

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" $\{\overline{S}_{k,p}; k = 0, 1, \ldots, N-1\}$; $\{\overline{S}_{k,p}; k = 0, 1, \ldots, N-1\}$ 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 $K = 8$ and we have $M = 32$ 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 [14] (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) [21], with uncorrelated Rayleigh fading for the different MTs and for the different paths (similar results were obtained for other severely time-dispersive channels). 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 interblock interleaving of the coded bits is assumed before the mapping procedure. The SISO decoder is implemented using the Max-Log-MAP approach. We have $P = K = 8$ MTs, corresponding to a fully loaded scenario and $\xi_p = 1$ for all MTs, i.e., we have a perfect power control (our simulations showed that the iterative receivers considered here are still suitable for scenarios without power control). At the BS we have L uncorrelated receive antennas, for diversity purposes.

Let us first consider that we have linear transmitters at each MT. Fig. 4 shows the coded BER performance for each iteration (averaged over all MTs) when either coded and uncoded soft decisions are used in the feedback loop, for different values of L (naturally, the first iteration corresponds to a linear receiver). From this figure it is clear that the iterative receiver allows significant performance improvements relatively to the linear receiver. We can also see that the turbo receiver outperform the one that uses uncoded soft decisions in the feedback loop, especially for $L = 1$. The higher BERs for the receiver with uncoded feedback are due to error propagation effects.

Let us assume now that the normalized clipping level, identical to all MTs, is $s_M/\sigma = 1.0$. This allows PMEPRs between 1.0dB and 4.3dB, depending on the out-of-band radiation levels (if we want to maintain the spectral occupation of conventional MC-CDMA signals the PMEPR is 4.3dB) [14] (for conventional MC-CDMA signals with a large number of subcarriers the PMEPR is about 8.4dB). Fig. 5 shows the corresponding coded BER performances. Clearly, the performance of the linear receiver is very poor, with high irreducible error floors due to the nonlinear distortion effects. This is especially serious for the case where $L = 1$. As we increase the number of iterations and/or we increase L improve significantly the performances, that can be close to the ones obtained with linear transmitters if $L > 1$.

V. CONCLUSIONS

In this paper we considered the uplink transmission in MC-CDMA systems employing clipping techniques. We proposed an iterative receiver structure that combine turbo-MUD and estimation and cancellation of the nonlinear distortion effects that are inherent to the transmitted signals.

Our performance results showed that the use of the channel decoder outputs instead of the coded MUD outputs, in the feedback loop, allow a significant performance improve, even

[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.

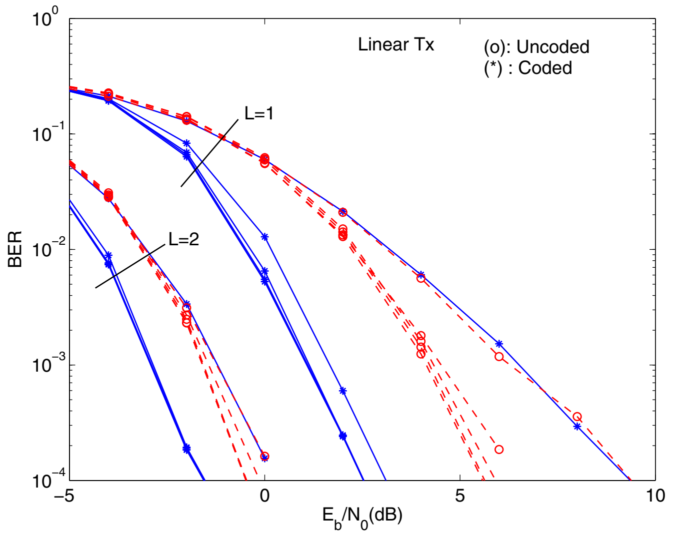

Fig. 4. Average coded BER performance for iterations 1 to 5 (better performances as we increase the number of iterations), for either coded and uncoded soft decisions in the receivers' feedback loop, when linear transmitters are considered.

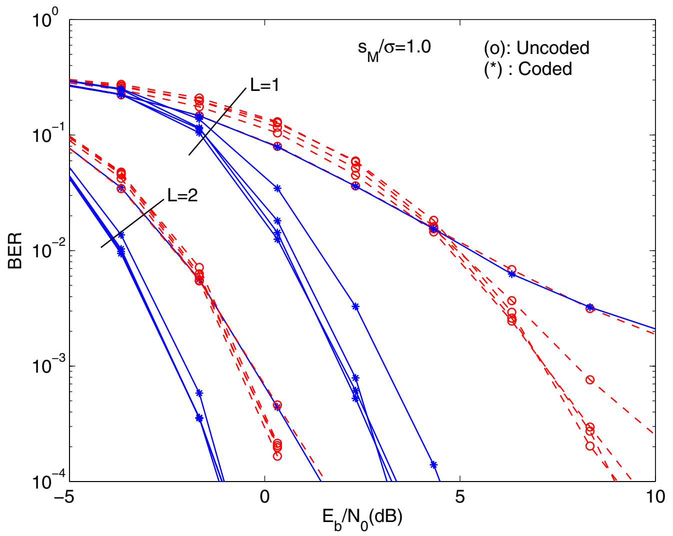

Fig. 5. As in Fig. 4, but for nonlinear transmitters.

for severely time-dispersive channels and/or when we decrease the clipping level of the transmitted signals.

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