Turbo Detection for the Uplink of CP-Assisted DS-CDMA Systems

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***Abstract*—In this paper we consider the use of CP-assisted (Cyclic Prefix) DS-CDMA schemes (Direct Sequence Code Division Multiple Access) in the uplink of broadband wireless systems. We present frequency-domain receivers that combine turbo equalization and multiuser detection. The performance of the proposed receiver can be close to the single-user MFB (Matched Filter Bound), even for fully loaded systems and severely time-dispersive channels. Moreover, our receivers can also cope with overloaded scenarios.**

I. Introduction

In DS-CDMA systems (Direct Sequence Code Division Multiple Access) all users transmit continuously, regardless of the bit rates. Therefore, the peak power requirements for the amplifiers are significantly reduced. Moreover, since DS-CDMA schemes can be regarded as single-carrier modulations, the transmitted signal associated to each spreading code can have low envelope fluctuations. For these reasons, DS-CDMA schemes are good candidates for broadband wireless systems, especially at the uplink. Since the optimum multiuser receiver is too complex [1], especially for severely time-dispersive channels, CP-assisted (Cyclic Prefix) DS-CDMA schemes were proposed, with a frequency-domain receiver design that have relatively low complexity, even for severely time-dispersive channels [2]. By taking advantage of the spectral correlations that result from the cyclostationarity of the signal transmitted by each MT [3], we can define a frequency-domain linear multiuser receiver. However, to avoid noise enhancement effects, this receiver is usually optimized in the MMSE sense (Minimum Mean Squared Error). This means that the linear MMSE receiver is not able to fully orthogonalize the different users (in fact, it is not even able to eliminate completely the ISI (Inter-Symbol Interference) in a single-user scenario). Therefore, we can have significant residual interference levels, especially for fully loaded scenarios and/or in the presence of strong interference levels. To overcome this problem, a promising multiuser receiver for the uplink of CP-assisted DS-CDMA systems was recently proposed [4], [5]. The asymptotic performance of the receivers can be close to the MFB (Matched Filter Bound) after just a few iterations. However, for low and moderate SNR (Signal-to-Noise Ratio), the region of operation of most systems, the performance of those receivers is not too different from the linear receiver.

Most digital transmission schemes employ some sort of channel coding and the typical detection strategy is to perform separately the channel equalization and channel decoding procedures. However, it is known that high performance gains can be achieved if these procedures are performed jointly. An effective way of achieving this is by employing the so-called turbo equalization schemes where the equalization and decoding procedures are repeated, in an iterative way, with some soft information being passed between them [6]. Although initially proposed for time-domain receivers, turbo equalizers allow also frequency-domain implementations [7], [8].

In this paper we consider the uplink transmission within a DS-CDMA system employing CP-assisted block transmission techniques. We propose frequency-domain receiver structures that combine turbo equalization and multiuser detection. Thanks to the frequency-domain implementation of our receivers, the required signal processing complexity is moderate, even for severely time-dispersive channels.

This paper is organized as follows: a linear receiver for multiuser detection of CP-assisted block transmission DS-CDMA schemes is described in Sec. II. In Sec. III we describe the frequency-domain receivers considered in this paper that combine turbo equalization with multiuser detection. Sec. IV presents a set of performance results and sec. V is concerned with the conclusions of the paper.

II. Linear Receiver for CP-Assisted DS-CDMA Schemes

In this paper we consider the uplink transmission in DS-CDMA systems employing CP-assisted block transmission techniques. We have a spreading factor K and P users. It is assumed that the received blocks associated to each user are synchronized in time (in practice, this means that there is a suitable "time-advance" mechanism, although just a coarse synchronization is required since some time misalignments can be absorbed by the CP).

The size-M data block to be transmitted by the pth user is $\{a_{n,p}; n = 0, 1, \ldots, M-1\}$, with $a_{n,p}$ selected from a given constellation. The corresponding chip block to be transmitted is $\{s_{n,p}; n = 0, 1, \ldots, N-1\}$, where $N = MK$ and $s_{n,p} = a_{\lfloor n/K \rfloor,p} c_{n,p}$ ($\lfloor x \rfloor$ denotes "larger integer not higher than x"), with $c_{n,p}$ denoting the spreading symbols.

It is assumed that the spreading sequences are periodic, with period K, i.e., $c_{n+K,p} = c_{n,p}$.

The signal received at the BS is sampled at the chip rate (the generalization for multiple samples per chip is straightforward) and the CP is removed, leading to the time-domain block $\{y_n; n = 0, 1, \ldots, N-1\}$. It can be shown that, when the CP is longer than the overall channel impulse response for each user, the corresponding frequency-domain block is $\{Y_k; k = 0, 1, \ldots, N-1\}$, where

$$Y_k = \sum_{p=1}^{P} S_{k,p} \xi_p H_{k,p}^{Ch} + N_k \tag{1}$$

with $H_{k,p}^{Ch}$ denoting the channel frequency response for the pth user and the kth frequency and N_k the channel noise for that frequency. ξ_p is a suitable scale factor that accounts for the overall attenuation between the pth MT and the BS (without loss of generality, it is assumed that $E[|H_{k,p}^{Ch}|^2] = 1$). The frequency-domain block $\{S_{k,p}; k = 0, 1, \ldots, N-1\}$ is the DFT of the chip block transmitted by the pth user, $\{s_{n,p}; n = 0, 1, \ldots, N-1\}$. It is shown it [5] that $S_{k,p} = A'_{k,p} C'_{k,p}$, where $\{C'_{k,p}; k = 0, 1, \ldots, N-1\}$ = DFT $\{c'_{n,p}; n = 0, 1, \ldots, N-1\}$, with $c'_{n,p} = c_{n,p}$ for $0 \le n < K$ and 0 otherwise, and $A'_{k,p} = \frac{1}{K} A_{k \bmod M,p}$, $k = 0, 1, \ldots, N-1$, with $\{A_{k,p}; k = 0, 1, \ldots, M-1\}$ = DFT $\{a_{n,p}; n = 0, 1, \ldots, M-1\}$. This means that, apart a constant, the block $\{A'_{k,p}; k = 0, 1, \ldots, N-1\}$ is the size-N periodic extension of the DFT of the data block associated to the pth user $\{A_{k,p}; k = 0, 1, \ldots, M-1\}$. This multiplicity in the $A'_{k,p}$ is related to the spectral correlations that are inherent to the cyclostationary nature of the transmitted signals [3]. Therefore, $Y_k = \sum_{p=1}^{P} A_{k,p} H_{k,p} + N_k$, with $H_{k,p} = \frac{1}{K} \xi_p H_{k,p}^{Ch} C'_{k,p}$ denoting the equivalent channel frequency response for the pth user and the kth frequency.

When we have $P \le K$ users, the K replicas associated to a given frequency domain sample $A_{k,p}$ can be employed to separate them. Therefore, the detection of the pth user could be made based on $\{\tilde{a}_{n,p}; n = 0, 1, \ldots, M-1\}$ = DFT $\{\tilde{A}_{k,p}; k = 0, 1, \ldots, M-1\}$, where $\tilde{A}_{k,p} = \sum_{l=0}^{K-1} F_{k+lM,p} Y_{k+lM}$. For each k ($k = 0, 1, \ldots, M-1$), the K coefficients $F_{k+lM,p}$, $l = 0, 1, \ldots, K-1$, are obtained by solving a system of K equations, as in [5].

III. Joint Turbo Equalization and Multiusers Detection

In this paper we consider iterative frequency-domain receivers based on the receiver proposed in [5]. Each iteration consists of P detection stages, one for each user. It is assumed that the users are ordered in descending order of their power and, when detecting a given user, we cancel the MAI (Multiple Access Interference) from all users, as well as the residual ISI for the user that is being detected. For each iteration, the frequency-domain samples associated with the pth user at the detector output are given by[1]

$$\tilde{A}_{k,p} = \sum_{l=0}^{K-1} F_{k+lM,p} Y_{k+lM} - \sum_{p'=1}^{P} B_{k,p}^{(p')} \rho_{p'} \hat{A}_{k,p'} \tag{2}$$

where $F_{k,p}$ ($k = 0, 1, \ldots, N-1$) and $B_{k,p}^{(p')}$ ($k = 0, 1, \ldots, M-1$; $p = 1, 2, \ldots, P$) denote the feedforward and the feedback coefficients, respectively, and the correlation coefficient ρ_p is defined as $\rho_p = E[\hat{a}_{n,p} a_{n,p}^*]/E[|a_{n,p}|^2]$. The block $\{\hat{A}_{k,p'}; k = 0, 1, \ldots, M-1\}$ is the DFT of the block $\{\hat{a}_{n,p'}; n = 0, 1, \ldots, M-1\}$, where the time-domain samples $\hat{a}_{n,p'}, n = 0, 1, \ldots, M-1$, are the latest estimates for the p'th user transmitted symbols, i.e., the hard-decisions associated with the block of time-domain samples $\{\tilde{a}_{n,p'}; n = 0, 1, \ldots, M-1\}$ = IDFT $\{\tilde{A}_{k,p'}; k = 0, 1, \ldots, M-1\}$. As in [4], either a SIC (Successive Interference Cancellation) or a PIC (Parallel Interference Cancellation) receiver can be adopted. For the ith iteration of a SIC receiver, $\hat{a}_{n,p'}$ is associated with the ith iteration for $p' < p$ and with the $(i-1)$th iteration for $p' \ge p$ (in the first iteration, we do not have any information for $p' \ge p$ and $\hat{a}_{n,p'} = 0$); for the PIC receiver, $\hat{a}_{n,p'}$ is always associated with the previous iteration (for the first iteration $\hat{a}_{n,p'} = 0$ for all p).

It can be shown that the optimum feedforward coefficients are [4][2]

$$F_{k+lM,p} = \frac{F_{k+lM,p}^I}{\gamma_p}, \quad l = 0, 1, \ldots, K-1, \tag{3}$$

where $F_{k+lM,p}^I$, $l = 0, 1, \ldots, K-1$ are the solution of the system of K equations[3]

$$\begin{gathered}\sum_{p'=1}^{P} (1 - \rho_{p'}^2) H_{k+lM,p'}^* \sum_{l'=0}^{K-1} F_{k+l'M,p}^I H_{k+l'M,p'} \\ + \alpha F_{k+lM,p}^I = H_{k+lM,p}^*, \quad l = 0, 1, \ldots, K-1,\end{gathered} \tag{4}$$

with $\alpha = E[|N_k|^2]/E[|A_{k,p}|^2]$ and

$$\gamma_p = \frac{1}{M} \sum_{k=0}^{M-1} \sum_{l=0}^{K-1} F_{k+lM,p}^I H_{k+lM,p}. \tag{5}$$

The optimum feedback coefficients are given by

$$B_{k,p}^{(p')} = \sum_{l'=0}^{K-1} F_{k+l'M,p} H_{k+l'M,p'} - \delta_{p,p'}, \tag{6}$$

$p' = 1, 2, \ldots, P$, ($\delta_{p,p'} = 1$ if $p = p'$ and 0 otherwise).

[1]It should be pointed out that, although our basic receiver is formally equivalent to the one proposed in [4], the feedback coefficients considered here are slightly different, since $\rho_{p'}$ is implicit in the definition of the $B_{k,p}^{(p')}$ in [4], instead of being explicit as in (2).

[2]Our feedforward coefficients are slightly different from the ones of [4], since we are considering a normalized receiver.

[3]As in [5], the feedforward coefficents can also be obtained by solving a set of P equations.

A. Soft Decisions

If we define the average time-domain symbols associated to a given iteration as

$$\overline{a}_{n,p} = \rho_p \hat{a}_{n,p}, \tag{7}$$

and the corresponding average frequency-domain block $\{\overline{A}_{k,p} = \rho_p \hat{A}_{k,p}; k = 0,1,\ldots,M-1\}$ = DFT $\{\overline{a}_{n,p}; n = 0,1,\ldots,M-1\}$, then (2) could be written as

$$\tilde{A}_{k,p} = \sum_{l=0}^{K-1} F_{k+lM,p} Y_{k+lM} - \sum_{p'=1}^{P} B_{k,p}^{(p')} \overline{A}_{k,p'} \tag{8}$$

($l = 0,1,\ldots,K-1$). This means that the feedback loop is fed with blockwise averages and ρ_p can be regarded as the blockwise reliability of the estimates $\{\hat{a}_{n,p}; n = 0,1,\ldots,M-1\}$.

To improve the performances, we could replace the "blockwise averages" by "symbol averages", which can be done as described in the following.

Let us assume that the transmitted symbols are selected from a QPSK (Quaternary Phase Shift Keying) constellation under a Gray mapping rule (the generalization to other cases is straightforward). We will define $a_{n,p} = \pm 1 \pm j = a_{n,p}^I + j a_{n,p}^Q$, with $a_{n,p}^I = \mathrm{Re}\{a_{n,p}\} = \pm 1$ and $a_{n,p}^Q = \mathrm{Im}\{a_{n,p}\} = \pm 1$, $n = 0,1,\ldots,M-1$, (similar definitions can be made for $\tilde{a}_{n,p} = \tilde{a}_{n,p}^I + j\tilde{a}_{n,p}^Q$, $\hat{a}_{n,p} = \hat{a}_{n,p}^I + j\hat{a}_{n,p}^Q$ and $\overline{a}_{n,p} = \overline{a}_{n,p}^I + j\overline{a}_{n,p}^Q$).

The LLRs (LogLikelihood Ratios) of the "in-phase bit" and the "quadrature bit", associated to $a_{n,p}^I$ and $a_{n,p}^Q$, respectively, are given by

$$L_{n,p}^I = \frac{2}{\sigma_p^2}\tilde{a}_{n,p}^I \quad \text{and} \quad L_{n,p}^Q = \frac{2}{\sigma_p^2}\tilde{a}_{n,p}^Q, \tag{9}$$

respectively, where

$$\sigma_p^2 = \frac{1}{2}E[|a_{n,p} - \tilde{a}_{n,p}|^2] \approx \frac{1}{2M}\sum_{n=0}^{M-1} E[|\hat{a}_{n,p} - \tilde{a}_{n,p}|^2]. \tag{10}$$

Under a Gaussian assumption, it can be shown that the mean value of $a_{n,p}$ is

$$\overline{a}_{n,p} = \tanh\left(\frac{L_{n,p}^I}{2}\right) + j\tanh\left(\frac{L_{n,p}^Q}{2}\right). \tag{11}$$

Clearly, the hard decisions $\hat{a}_{n,p}^I = \pm 1$ and $\hat{a}_{n,p}^Q = \pm 1$ are defined according to the signs of $L_{n,p}^I$ and $L_{n,p}^Q$, respectively. Therefore, $\overline{a}_{n,p} = \rho_{n,p}^I \hat{a}_{n,p}^I + j\rho_{n,p}^Q \hat{a}_{n,p}^Q$, where $\rho_{n,p}^I = E[a_{n,p}^I \hat{a}_{n,p}^I]/E[|a_{n,p}^I|^2] = \tanh\left(|L_{n,p}^I|/2\right)$ and $\rho_{n,p}^Q = E[a_{n,p}^Q \hat{a}_{n,p}^Q]/E[|a_{n,p}^Q|^2] = \tanh\left(|L_{n,p}^Q|/2\right)$ can be regarded as the reliabilities associated to the "in-phase" and "quadrature" bits of the nth symbol of the pth user (naturally, $0 \le \rho_{n,p}^I \le 1$ and $0 \le \rho_{n,p}^Q \le 1$) (for the first iteration, $\rho_{n,p}^I = \rho_{n,p}^Q = 0$). The feedforward coefficients are still obtained from (3)-(5), but with the blockwise reliability given by

$$\rho_p = \frac{1}{M}\sum_{n=0}^{M-1} \frac{E[a_{n,p}^* \hat{a}_{n,p}]}{E[|a_{n,p}|^2]} = \frac{1}{2M}\sum_{n=0}^{M-1} (\rho_{n,p}^I + \rho_{n,p}^Q). \tag{12}$$

Therefore, the receiver with "blockwise reliabilities", denoted in the following as IMUD-HD (Iterative MUD with Hard Decisions), and the receiver with "symbol reliabilities", denoted in the following as IMUD-SD (Iterative MUD with Soft Decisions), employ the same feedforward coefficients; however, in the first the feedback loop uses the "hard-decisions" on each data block, weighted by a common reliability factor, while in the second the reliability factor changes from symbol to symbol (in fact, the reliability factor is different in the real and imaginary component of each symbol).

B. Use of Channel Decoder Outputs in the Feedback Loop

We can define a receiver with joint equalization and multiuser detection that, as the turbo equalizers, employs the channel decoder outputs instead of the uncoded "soft decisions" in the feedback loop. The receiver structure, that will be denoted Turbo MUD-SD (Turbo MUD with Soft Decisions), is similar to the IMUD-SD, but with a SISO channel decoder (Soft-In, Soft-Out) employed in the feedback loop. The SISO block, that can be implemented as defined in [9], provides the LLRs 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, given by (9). Once again, the feedforward coefficients are obtained from (3)-(5), with the blockwise reliability given by (12).

As an alternative, we could employ a conventional decoder, unable to provide the LLRs of the "coded bits" but allowing a simpler implementation (e.g., a Viterbi decoder in the case of convolutional codes). Since we are not able to obtain the reliabilities of each bit, we can assume that $\rho_p = \rho_{p,p}^I = \rho_{p,p}^Q = 1$, which is a simple, but suboptimal, solution. This receiver will be denoted Turbo MUD-HD (Turbo MUD with Hard Decisions).

IV. Performance Results

In this section, we present a set of performance results concerning the proposed receivers with joint turbo equalization and multiuser detection. The iterative SIC structure was considered. We consider the uplink transmission within a CP-assisted DS-CDMA system with spreading factor $K = 4$ and $M = 256$ data symbols for each user, corresponding to $N = KM = 1024$. The blocks have length 4μs. QPSK constellations, with Gray mapping, are employed. The radio channel is characterized by the power delay profile type C for HIPERLAN/2 (HIgh PERformance Local Area Network) [10], with uncorrelated Rayleigh fading on the different paths. We consider perfect synchronization and channel estimation conditions. The signals associated to all users have the same average power at the receiver (i.e., the BS), which corresponds to a scenario where an "ideal average power control" is implemented.

Let us first consider uncoded performances and $P = 4$ users (i.e., a fully loaded system). In Fig. 1 we compare the average performance for each iteration (i.e., the average over all the users) when either hard decisions or soft decisions are used in the feedback loop, i.e., for the IMUD-HD and the IMUD-SD,

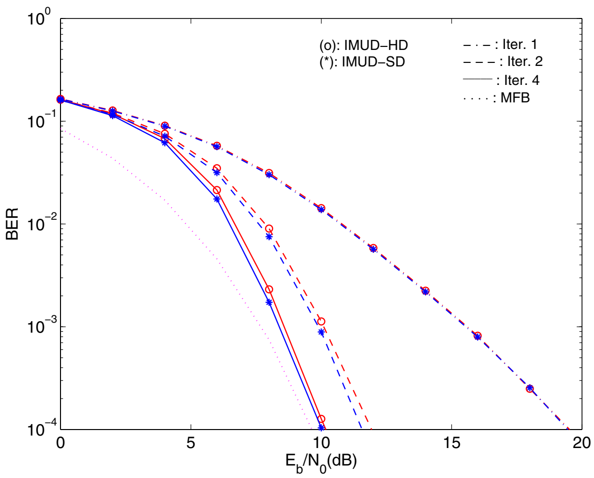

Fig. 1. Average uncoded BER performances for each iteration and a receiver employing either hard decisions or soft decisions in the feedback loop.

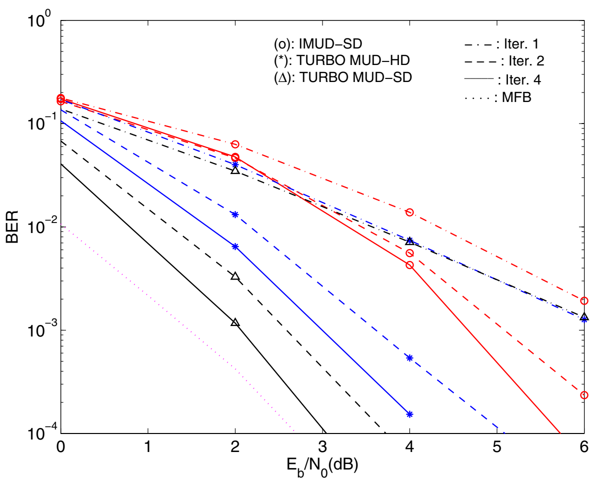

Fig. 2. Average coded BER performances for each iteration.

respectively. Clearly, the use of soft decisions allows just a slight performance improvement (a few tens of dB, at most). In both cases, the performances are asymptotically close to the MFB; however, for low SNR we are still far from the MFB.

Let us consider now the impact of channel coding. We consider the well-known rate-1/2 64-state convolutional code with generators $1+D^2+D^3+D^5+D^6$ and $1+D+D^2+D^3+D^6$. For Fig. 2 we consider, once again, a fully loaded system ($P=4$), in the following cases: uncoded feedback with soft decisions (IMUD-SD); coded feedback with a conventional Viterbi decoder, assuming $\rho_p = \rho^I_{p,p} = \rho^Q_{p,p} = 1$ (Turbo MUD-HD); coded feedback with a SISO decoder, implemented using the Max-Log-MAP approach (Turbo MUD-SD). From this figure, it is clear that the turbo receivers (Turbo MUD-HD and Turbo MUD-SD) outperform the receiver that uses uncoded soft decisions in the feedback loop (IMUD-SD), even the sub-optimum Turbo MUD-HD based on the Viterbi decoder. Moreover, the performance of Turbo MUD-SD approaches the corresponding MFB, contrarily to the other schemes.

Let us consider now an overloaded scenario, i.e., when $P > K$. In Fig. 3 we show the average performances for the receivers of Fig. 2, when $P = 6$. In this case, we are able

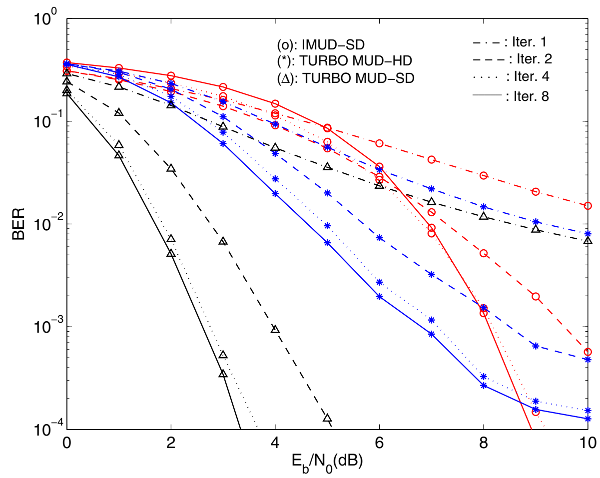

Fig. 3. Average coded BER performances when $P = 6$ and $K = 4$.

to separate the different users, even when using an uncoded feedback, although the required number of iterations is higher, and with some performance degradation. If we increase the number of users the only receiver able to separate them is the turbo receiver that uses the SISO block in the feedback receiver (Turbo MUD-SD).

V. Conclusions

In this paper we considered the use of CP-assisted DS-CDMA schemes in the uplink of broadband wireless systems and we presented frequency-domain receivers that combine turbo equalization and multiuser detection.

The proposed receivers can have performances close to the single-user performance, even for fully loaded systems and severely time-dispersive channels. Moreover, our receivers can also cope with overloaded scenarios, where the number of users exceeds the spreading factor.

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