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Multi-group linear turbo equalization with intercell interference cancellation for MC-CDMA cellular systems

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Summary

In this paper, we investigate multi-group linear turbo equalization using single antenna interference cancellation (SAIC) techniques to mitigate the intercell interference for multicarrier code division multiple access (MC-CDMA) cellular systems. It is important for the mobile station to mitigate the intercell interference as the performance of the users close to cell edge is mainly degraded by the intercell interference. The complexity of the proposed iterative detector and receiver is low as the one-tap minimum mean square error (MMSE) equalizer is employed for mitigating the intracell interference, while a simple group interference canceller is used for suppressing the intercell interference. Simulation results show that the proposed iterative detector and receiver can mitigate the intercell interference effectively through iterations for both uncoded and coded signals.

KEY WORDS: iterative receiver; frequency-domain MMSE equalizer; single antenna interference cancellation (SAIC); multicarrier code division multiple access (MC-CDMA); intercell interference

1. Introduction

In most cellular systems, since the interference can degrade the performance it is important to avoid or mitigate the interference. There are two different types of interference: intercell and intracell interference. In downlink channels, some orthogonal multiple access schemes, including orthogonal frequency division multiple access (OFDMA), can avoid the intracell interference. In multicarrier code division multiple access (MC-CDMA) system (where the orthogonality can be destroyed by frequency selective fading), the intracell interference can also be effectively mitigated by the frequency domain equalizer provided that the spreading codes are orthogonal [1]. However, it is generally difficult to mitigate the intercell interference, especially at the cell boundary. Increasing frequency reuse factor can help at the expense of low spectral efficiency. Another solution is rate control via adaptive modulation and coding (AMC) schemes (e.g., in long-term evolution (LTE) systems [2]), where the base station adjusts transmission rate depending on the channel condition. For example, transmission rate with users at cell edge should be lower compared to that of users under better channel condition. Again, this solution suffers low spectral efficiency due to low transmission rate.

Alternatively, in order to have a high spectral efficiency, an intercell interference mitigation technique can be employed to provide satisfactory

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performance. In general, linear techniques can provide reasonably good performance when multiple received signals are available using multiple receive antennas [3]. Otherwise, nonlinear techniques should be used in single antenna receivers. Various single antenna interference cancellation (SAIC) techniques have been proposed for time division multiple access (TDMA) based cellular systems to mitigate the intercell interference [4]-[7]. By means of SAIC, interference is removed from the desired signals using either filter-based approaches or multi-user detection techniques [8] when only single antenna is available at the receiver. SAIC techniques are cost-effective and applicable to existing TDMA based cellular systems when the reuse factor decreases to improve the spectral efficiency.

To further improve performance of the interference canceller, soft-input soft-output (SISO) equalizers have been introduced [9, 10, 11] where soft information is exchanged between the detector/decoder and the canceller to improve overall performance through iterations. In multi-user environments, group cancellation based detection can be used either in parallel [12] or successive [22] fashions. Inspired by those approaches, in this paper, we propose a multi-group turbo equalization with intercell interference group cancellation for MC-CDMA cellular downlinks. The mitigation of intercell interference within MC-CDMA based cellular systems is not relatively well addressed except in a few literatures, including [13]-[15]. In our approach, each group is a number of user signals from the same base station. While the frequency domain equalizer can suppress the intracell interference, the soft group canceller is found to be efficient in mitigating the intercell interference. Note that the intercell interference and spreading code are jointly taken into account when obtaining the log-likelihood ratio (LLR) for the soft canceller and consequently the performance of the proposed iterative receiver is improved. It is shown in our simulation that the performance improvement is even more impressive when including the channel decoder into our design.

The rest of the paper is organized as follows. In Section 2, system models are described. In Section 3, we propose a multi-group turbo equalization based detector and present an approach to find the soft-decision for cancellation. An iterative receiver for coded signals is discussed in Section 4. We conclude this paper with some remarks in Section 5.

2. System Models

In this section, we describe downlink MC-CDMA in multi-cell environments, in which the detection performance would be degraded by the intercell interference if a mobile station is close to cell edge.

2.1. Downlink MC-CDMA

Throughout this paper, we consider an MC-CDMA based cellular system with a frequency reuse factor of 1. Assume that there are L subcarriers and spread signals are transmitted by L subcarriers. Throughout the paper, we consider a mobile user at cell edge, whose receiver can receive almost equally strong downlink signals from multiple BSs.

The signal to be transmitted by orthogonal frequency division multiplexing (OFDM) from the qth BS can be written as

$$\mathbf{s}_q = \sum_{k=1}^{K} \mathbf{c}_{q,k} b_{q,k}, \tag{1}$$

where $\mathbf{c}_{q,k}$ and $b_{q,k}$ denote the $L \times 1$ spreading code and data symbol of the kth user from the qth BS, and K denotes the number of signals (or users) in downlink. For convenience, we assume that binary phase shift keying (BPSK) is used: $b_{q,k} \in \{-1, +1\}$. Let

$$\begin{aligned} \mathbf{C}_q &= [\mathbf{c}_{q,1} \quad \mathbf{c}_{q,2} \quad \cdots \quad \mathbf{c}_{q,K}], \\ \mathbf{b}_q &= [b_{q,1} \quad b_{q,2} \quad \cdots \quad b_{q,K}]^{\mathrm{T}}. \end{aligned}$$

Then, we have $\mathbf{s}_q = \mathbf{C}_q \mathbf{b}_q$. With cyclic prefix (CP), $\mathbf{s}_q$ can be transmitted using OFDM. For OFDM, the inverse discrete Fourier transform (IDFT) can be used; for details, see [1].

At the receiver, after discrete Fourier transform (DFT) and deleting the CP part, the received signals over L subcarriers can be obtained. Letting $\mathbf{H}_q = \mathrm{Diag}(H_{q,0}, H_{q,1}, \ldots, H_{q,L-1})$, where $H_{q,l}$ denotes the channel frequency response corresponding to the lth subcarrier from the qth BS to the mobile station. For convenience, the channel from the qth BS to the mobile station is referred to as channel q. Suppose that the impulse response of channel q is written as

$$h_q(t) = \sum_{p=0}^{P-1} h_{q,p} \delta(t - qT_{\mathrm{c}}),$$

where P denotes the number of multipaths, $h_{q,p}$ denotes the pth multipath coefficient of channel q, $T_{\mathrm{c}} = T/L$, and $\delta(t)$ denotes the Dirac delta. Here,

T denotes the symbol duration. The relation between $h_{q,p}$ and $H_{q,l}$ is given by

$$H_{q,l} = \sum_{p=0}^{P-1} h_{q,p} e^{-j2\pi pl/L}.$$

2.2. Multicell Environments

Considering a cellular environment which includes Q cells (i.e., Q BSs), the received signal vector in the frequency domain (or after DFT) is given by

$$\mathbf{r} = \sum_{q=1}^{Q} \mathbf{H}_q \mathbf{C}_q \mathbf{b}_q + \mathbf{n}, \quad (2)$$

where $\mathbf{n}$ denotes the background noise, which is a circular complex Gaussian random vector with $E[\mathbf{n}] = \mathbf{0}$ and $E[\mathbf{n}\mathbf{n}^{\mathrm{H}}] = \sigma_n^2 \mathbf{I}$.

We assume that the spreading codes for downlink channels within a cell are orthogonal. That is, $\mathbf{C}_q^{\mathrm{H}} \mathbf{C}_q = \mathbf{I}$. However, the spreading codes among different cells may not be orthogonal, i.e., $\mathbf{c}_{q,k}^{\mathrm{H}} \mathbf{c}_{q',k'} \neq 0$ for $q \neq q'$. Due to the orthogonality of spreading codes within a cell, a simple one-tap frequency domain equalizer can be used to equalize the frequency-selective channel, and then a despreader can be followed to extract the desired signal [1]. However, the intercell interference may not be effectively mitigated by the simple one-tap frequency domain equalizer, because the orthogonality (between spreading codes of neighbor cells) is not guaranteed.

Note that the orthogonality of the spreading codes cannot be recovered by the channel equalization for uplink channels, because each user's channel is different. In this case, the minimum mean squared error (MMSE) detection (rather than the MMSE channel equalization) can be employed with the MSE cost, $E[|b_{q,k} - \mathbf{w}_{q,k}^H \mathbf{r}|^2]$ [17]. The same approach can also be applied for downlink to mitigate both intercell and intracell interference. However, it requires the matrix inversion of an $L \times L$ matrix. Since L is usually large (several hundreds or thousands), it seems the MMSE detection is impractical due to prohibitively high complexity. Thus, in the next section (i.e., Section 3), we consider the MMSE channel equalization, which has much less computational complexity than the MMSE detection, to build an iterative detector.

3. Iterative Detectors using MMSE Equalization with Group Cancellation

In this section, group cancellation is employed to suppress the intercell interference within the proposed iterative detector. Once the intercell interference is cancelled, the one-tap MMSE equalizer is applied to mitigate the intracell interference.

3.1. Proposed Iterative Detector and EXIT Charts

The proposed iterative detector consists of the group interference canceller and one-tap MMSE equalizer. The group interference canceller is adopted to cancel the intercell interference, and the MMSE equalizer is used to suppress the intracell interference together with despreaders using the orthogonality of spreading codes within a cell. This detector is different from the existing ones. For example, in [16], the cancellation is considered to suppress the intracell interference, while, in the proposed detector, the equalizer and despreader are used to suppress the intracell interference to take advantage of the orthogonality. The proposed detector can be seen as a generalized one of that proposed in [1, 18] with the canceller to mitigate the intercell interference. Fig. 1 shows the structure of the proposed iterative detector when $Q = 2$ (two cells).

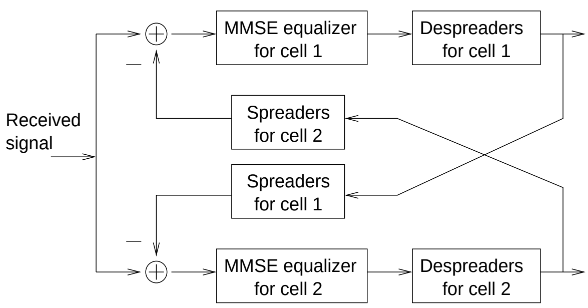


Fig. 1. Block diagram for the proposed iterative detector when $Q = 2$: SD and HD denote soft-decision and hard-decision, respectively.

The operation of the proposed receiver can be understood with the extrinsic information transfer (EXIT) chart [19] between the two detectors in Fig. 1. For convenience, the upper detector is referred to as detector 1, while the lower detector is referred to as detector 2, in Fig. 1. In the first iteration, detector 1 performs the detection (for the signals from BS 1) with no extrinsic bit information of the signals from BS

2. Thus, the cancellation is not performed. Note that the LLR can be used as the extrinsic bit information (an approach to obtain the LLR is given in Subsection 3.3). Then, in detector 2 (to detect the signals from BS 2), with the soft-decision of the signals from BS 1, the extrinsic bit information can be obtained and used for the (soft interference) cancellation to suppress the interfering signals from BS 1. The performance improvement through iterations can be seen with EXIT charts. For illustration purpose, consider

$$\begin{aligned}[h_{1,0}, h_{1,1}, \ldots, h_{1,4}] &= [0.227,\ 0.46,\ 0.688, \\ &\qquad\qquad 0.46,\ 0.227] \\ [h_{2,0}, h_{2,1}, \ldots, h_{2,4}] &= \frac{1}{\sqrt{5}}[1,\ 1,\ 1,\ 1,\ 1].\end{aligned}$$

With $L = 128$ and the signal to noise ratio (SNR) = 6 dB, we can have the results in Fig. 2. Fig. 2 (a) shows the empirical pdf of the LLRs from detector 1 (from histogram) and the fitted Gaussian pdf with the same mean and variance as those of the empirical pdf. Assume that the pdf of the LLRs follows the Gaussian pdf, the parametric approach in [20] is used to obtain the EXIT chart shown in Fig. 2 (b). According to the EXIT chart in Fig. 2 (b), we can see that a few iterations are required for convergence and the reliability is improved as the SNR increases.

For the case of SNR = 6 dB, we have three points in the EXIT charts, labelled by “A”, “B”, and “C” in Fig. 2 (b). Point B is the cross point of the two EXIT charts (which is the convergence point), while point A corresponds to the performance with all the interference (i.e., without interference cancellation) and point C corresponds to the performance without the interference. Point B is not close to point C, and this indicates that the performance of the iterative detector would not be the ideal one although the iterative detector converges. There would be the residual interference (as the soft-cancellation is used) after the convergence and the performance degradation (from the ideal one) is expected.

In general, it is possible to generalize the proposed iterative detector for multiuser detection in CDMA or MC-CDMA when the system is over-loaded. For convenience, consider a single cell system (i.e., there is no intercell interference). The signals can be divided into Q subgroups, and Q linear detectors (possibly MMSE detectors) can be designed to detect each subgroup's signals with assuming that the signals in other subgroups are interfering signals. If the number of users in each subgroup, say K/Q, is less than the processin gain, L, the linear detection becomes effective in suppressing the interference within the subgroup, while the interfering signals from the other subgroups are cancelled by group interference cancellers. Then, the proposed approach for the iterative detection can be used for the iteration. The resulting iterative detector consists of linear detectors and interference cancellers.

3.2. Intercell Interference Cancellation and MMSE Equalization

In this subsection, we consider the MMSE equalization for the intracell interference after soft cancellation of the intercell interference.

Suppose that the mean vector of $\mathbf{b}_q$, denoted by $\bar{\mathbf{b}}_q$, is available from prior information (or the previous detection in interative processing). Prior to the one-tap MMSE equalization, the soft cancellation for the group detection of $\mathbf{b}_q$ can be carried out as follows:

$$\mathbf{r}_q = \mathbf{r} - \sum_{m \neq q} \mathbf{H}_m \bar{\mathbf{s}}_m = \mathbf{H}_q \mathbf{s}_q + \sum_{m \neq q} \mathbf{H}_m \tilde{\mathbf{s}}_m + \mathbf{n}, \quad (3)$$

where $\bar{\mathbf{s}}_q = \mathbf{C}_q \bar{\mathbf{b}}_q$ and $\tilde{\mathbf{s}}_q = \mathbf{C}_q \tilde{\mathbf{b}}_q$. Here, $\tilde{\mathbf{b}}_q = \mathbf{b}_q - \bar{\mathbf{b}}_q$. After the soft cancellation, the MMSE channel equalization is applied. The one-tap MMSE equalizer coefficients can be obtained as follows:

$$\begin{aligned}\mathbf{G}_q &= \arg\min_{\mathbf{G}} E[||\mathbf{s}_q - \mathbf{G}\mathbf{r}_q||^2] \\ &= \arg\min_{\mathbf{G}} \sum_{l=0}^{L-1} E[|s_{q,l} - G_{q,l} r_{q,l}|^2], \quad (4)\end{aligned}$$

where the lth diagonal element of $\mathbf{G}_q$ is individually obtained as

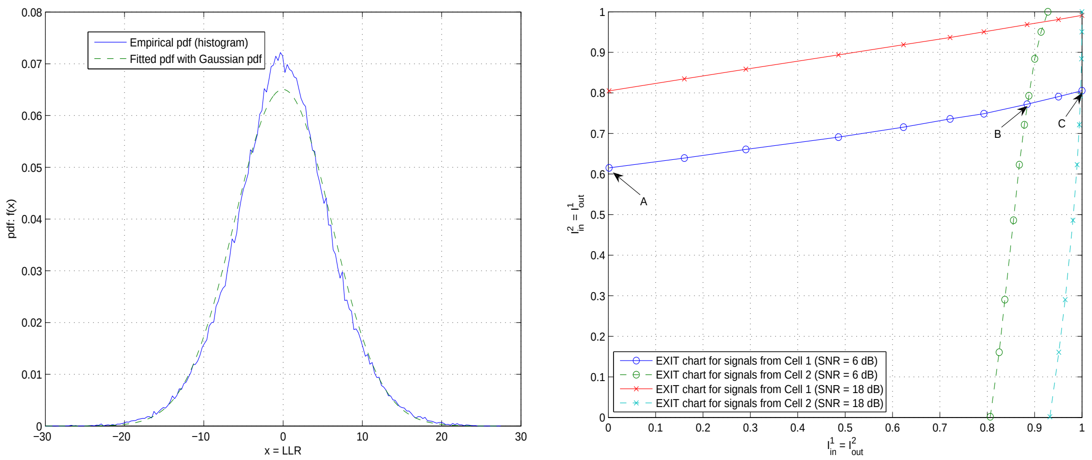
$$G_{q,l} = \frac{H_{q,l}^* E[|s_{q,l}|^2]}{|H_{q,l}|^2 E[|s_{q,l}|^2] + U_{m,l}}. \quad (5)$$

Here, $U_{m,l} = \sum_{m \neq q} |H_{m,l}|^2 E[|\tilde{s}_{m,l}|^2] + \sigma_n^2$ and $E[|\tilde{s}_{m,l}|^2]$ is the variance of the residual interference. Since no matrix inversion is required, the complexity of the one-tap MMSE equalizer is low (on the other hand, the cancellation based approach (e.g., [16]) requires a higher complexity due to matrix inversion). The fact that complexity of both intercell interference cancellation and one-tap MMSE equalization is low leads to a reduced overall complexity of the proposed iterative detector.

Once the channel response is equalized by the MMSE equalizer, the signals from the qth BS can be detected by despreading operation as follows:

$$\hat{\mathbf{b}}_q = \mathbf{C}_q^{\mathrm{H}} \hat{\mathbf{s}}_q, \quad (6)$$

where $\hat{\mathbf{s}}_q = \mathbf{G}_q \mathbf{r}_q$.



(a) (b)

Fig. 2. (a) Empirical pdf of the LLR with a fitted pdf with Gaussian pdf when SNR = 6 dB; (b) EXIT charts for the iterative detector when SNR = 6 and 18 dB.

3.3. Finding LLR for Soft Decision

For the soft interference cancellation, it is important to derive the LLR, because the mean value of the data symbol can be found from the LLR.

Consider the output of the despreader as follows:

$$\begin{aligned} \hat{b}_{q,k} &= \sum_{l=0}^{L-1} c^*_{q,k,l} G_{q,l} r_{q,l} \\ &= \sum_{l=0}^{L-1} c^*_{q,k,l} G_{q,l} (H_{q,l} c_{q,k,l} b_{q,k} + \\ &\qquad \sum_{k' \neq k} H_{q,l} c_{q,k',l} b_{q,k'} + \sum_{m \neq q} H_{m,l} \tilde{s}_{m,l} + n_l) \\ &= A_{q,k} b_{q,k} + \eta_{q,k}, \end{aligned} \tag{7}$$

where

$$\begin{aligned} A_{q,k} &= \sum_{l=0}^{L-1} G_{q,l} H_{q,l} c^*_{q,k,l} c_{q,k,l}, \\ \eta_{q,k} &= \sum_{k' \neq k} \Big(\sum_{l=0}^{L-1} G_{q,l} H_{q,l} c^*_{q,k,l} c_{q,k',l} b_{q,k'}\Big) \\ &\quad + \sum_{m \neq q} \Big(\sum_{l=0}^{L-1} G_{q,l} H_{m,l} c^*_{q,k,l} \tilde{s}_{m,l}\Big) \\ &\quad + \sum_{l=0}^{L-1} G_{q,l} c^*_{q,k,l} n_l. \end{aligned} \tag{8}$$

To obtain the LLR, we can assume that $\eta_{q,k}$ is a Gaussian random variable (this is called the Gaussian assumption). The Gaussian assumption is also used in obtaining the LLR in [18, 21, 22].

It can be readily shown that the signal gain, $A_{q,k}$, depends only on q:

$$A_{q,k} = A_q = \frac{1}{L} \sum_{l=0}^{L-1} G_{q,l} H_{q,l}.$$

There are three terms in $\eta_{q,k}$ as shown in (8): the first term is the self-interference, the second term is the intercell interference, and the third term is the background noise. Since they are uncorrelated, the variance of $\eta_{q,k}$ can be found from each term's variance. Denoting by σ^2_{SI} the variance of the self-interference, it can be shown that

$$\begin{aligned} \sigma^2_{SI} &= E\Big[\Big|\sum_{k' \neq k} \Big(\sum_{l=0}^{L-1} G_{q,l} H_{q,l} c^*_{q,k,l} c_{q,k',l} b_{q,k'}\Big)\Big|^2\Big] \\ &= \mathbf{c}^{\mathrm{H}}_{q,k} \mathbf{G}_q \mathbf{H}_q \bar{\mathbf{C}}_{q,k} \bar{\mathbf{C}}^{\mathrm{H}}_{q,k} \mathbf{H}^{\mathrm{H}}_q \mathbf{G}^{\mathrm{H}}_q \mathbf{c}_{q,k} \\ &= \mathbf{c}^{\mathrm{H}}_{q,k} \mathbf{G}_q \mathbf{H}_q (\mathbf{C}_q \mathbf{C}^{\mathrm{H}}_q - \mathbf{c}_{q,k} \mathbf{c}^{\mathrm{H}}_{q,k}) \mathbf{H}^{\mathrm{H}}_q \mathbf{G}^{\mathrm{H}}_q \mathbf{c}_{q,k} \\ &= [\mathbf{C}^{\mathrm{H}}_q \mathbf{G}_q \mathbf{H}_q \mathbf{C}_q \mathbf{C}^H_q \mathbf{H}^{\mathrm{H}}_q \mathbf{G}^{\mathrm{H}}_q \mathbf{C}_q]_{k,k} \\ &\qquad - |\mathbf{c}^{\mathrm{H}}_{q,k} \mathbf{G}_q \mathbf{H}_q \mathbf{c}_{q,k}|^2 \\ &= [\mathbf{C}^{\mathrm{H}}_q \mathbf{G}_q \mathbf{H}_q \mathbf{C}_q \mathbf{C}^H_q \mathbf{H}^{\mathrm{H}}_q \mathbf{G}^{\mathrm{H}}_q \mathbf{C}_q]_{k,k} \\ &\qquad - |A_q|^2, \end{aligned} \tag{9}$$

where $\bar{\mathbf{C}}_{q,k}$ is the submatrix of $\mathbf{C}_q$ obtained by deleting the kth column vector.

If we assume that the residual interference, $\tilde{s}_{m,l}$, has the same statistical properties for all l, we have

$$E[|\tilde{s}_{m,l}|^2] = \sigma^2_{\tilde{s},m}, \quad \text{for all } l. \tag{10}$$

This can reduce the computational complexity to find the variance of the intercell interference after soft cancellation. Using (10) we can show that

$$E[|\sum_{l=0}^{L-1} G_{q,l} H_{m,l} c^*_{q,k,l} \tilde{s}_{m,l}|^2] \\ = \sigma^2_{\tilde{s},m} [\mathbf{C}^{\mathrm{H}}_q \mathbf{G}_q \mathbf{H}_m \mathbf{H}^{\mathrm{H}}_m \mathbf{G}^{\mathrm{H}}_q \mathbf{C}_q]_{k,k}. \quad (11)$$

Denoting by σ^2_{II} the variance of the intercell interference, it then follows that

$$\begin{aligned} \sigma^2_{II} &= E[|\sum_{m \neq q} (\sum_{l=0}^{L-1} G_{q,l} H_{m,l} c^*_{q,k,l} \tilde{s}_{m,l})|^2] \\ &= \sigma^2_{\tilde{s},m} [\mathbf{C}^{\mathrm{H}}_q \mathbf{G}_q (\sum_{m \neq q} \mathbf{H}_m \mathbf{H}^{\mathrm{H}}_m) \mathbf{G}^{\mathrm{H}}_q \mathbf{C}_q]_{k,k}. \end{aligned}$$

Note that $\mathbf{D}_{q,m} = \mathbf{G}_q (\sum_{m \neq q} \mathbf{H}_m \mathbf{H}^{\mathrm{H}}_m) \mathbf{G}^{\mathrm{H}}_q$ is diagonal, because $\mathbf{G}_q$ and $\mathbf{H}_m$ are diagonal. Then, it follows that

$$\begin{aligned} \sigma^2_{II} &= \sigma^2_{\tilde{s},m} [\mathbf{C}^{\mathrm{H}}_q \mathbf{D}_{q,m} \mathbf{C}_q]_{k,k} \\ &= \sigma^2_{\tilde{s},m} \sum_{l=1}^{L} \left|[\mathbf{C}_q]_{l,k}\right|^2 [\mathbf{D}_{q,m}]_{l,l}. \end{aligned} \quad (12)$$

Thus, the complexity is low (its complexity is the same as the complexity of inner product of two vectors rather than that of multiplication of two matrices).

The variance of the background term, denoted by σ^2_{BG}, is given by

$$\begin{aligned} \sigma^2_{BG} &= E[|\sum_{l=0}^{L-1} G_{q,l} c^*_{q,k,l} n_l|^2] \\ &= \sigma^2_n [\mathbf{C}^{\mathrm{H}}_q \mathbf{G}_q \mathbf{G}^{\mathrm{H}}_q \mathbf{C}_q]_{k,k} \\ &= \frac{\sigma^2_n |G_{q,k}|^2}{L}. \end{aligned} \quad (13)$$

Taking the summation of the three variances obtained in (9), (12), and (13), the variance of $\eta_{q,k}$, denoted by $\sigma^2_\eta(q,k)$, can be found. Then, using the Gaussian assumption, the LLR can be given by

$$\mathrm{LLR}(\hat{b}_{q,k}) = \frac{4A_q}{\sigma^2_\eta(q,k)} \hat{b}_{q,k}. \quad (14)$$

Note that there are two major differences in obtaining the LLR in this section from [18]: (i) the intercell interference is taken into account; (ii) the spreading codes are also taken into account. On the other hand, in [18], the LLR is obtained without the intercell interference. In addition, the LLR is found under the assumption of random spreading codes. Thus, the LLR expression does not include the spreading codes.

Finally, the mean value of $b_{q,k}$ from the LLR is found as follows:

$$\bar{b}_{q,k} = \tanh\left(\mathrm{LLR}(\hat{b}_{q,k})/2\right). \quad (15)$$

The variance of $b_{q,k}$ becomes

$$\sigma^2_{q,k} = 1 - (\bar{b}_{q,k})^2.$$

Since $\tilde{\mathbf{s}}_q = \mathbf{C}_q \tilde{\mathbf{b}}_q$, we have

$$E[\tilde{\mathbf{s}}_q \tilde{\mathbf{s}}^{\mathrm{H}}_q] = \mathbf{C}_q \mathrm{Diag}(\sigma^2_{q,1}, \sigma^2_{q,2}, \ldots, \sigma^2_{q,K_q}) \mathbf{C}^{\mathrm{H}}_q.$$

According to (10), we can obtain the variance of $\tilde{s}_{q,l}$ using the average of the diagonal elements of $E[\tilde{\mathbf{s}}_q \tilde{\mathbf{s}}^{\mathrm{H}}_q]$:

$$\begin{aligned} \sigma^2_{\tilde{s},q} &= \frac{1}{L} \mathrm{Tr}(E[\tilde{\mathbf{s}}_q \tilde{\mathbf{s}}^{\mathrm{H}}_q]) \\ &= \frac{1}{L} \mathrm{Tr}\left(\mathbf{C}_q \mathrm{Diag}(\sigma^2_{q,1}, \sigma^2_{q,2}, \ldots, \sigma^2_{q,K_q}) \mathbf{C}^{\mathrm{H}}_q\right) \\ &= \frac{1}{L} \sum_{k=1}^{K_q} \sigma^2_{q,k}. \end{aligned} \quad (16)$$

3.4. Simulation Results

For simulations, we consider an MC-CDMA based cellular system with two cells ($Q = 2$). The number of subcarriers is set to $L = 128$. For Rayleigh multipath fading channels, the following power delay profile is used:

$$E[|h_{q,p}|^2] = 1/P, \; p = 0, 1, \ldots, P-1,$$

where P is set to 6. The two channels have identical statistical properties. Thus, the two signals from two BSs are equally strong.

Fig. 3 shows the BER performance for different SNR values with $K = L = 128$ (full loading per cell). Note that the SNR is equivalent to E_b/N_0 where E_b denotes the bit energy and $N_0 = \sigma^2_n$. For the iterative detector, both hard-decision and soft-decision are considered. As shown in Fig. 3, the iterative detector can improve the performance and soft-decision can provide a better performance than hard-decision. Four iterations are run in the iterative detector.

Although the iterative detector can outperform the conventional non-iterative detector, its performance is still worse than the ideal performance obtained without intercell interference, as shown in Fig. 3. Since the results in Fig. 3 are obtained when the system is fully loaded, it would be interesting to see

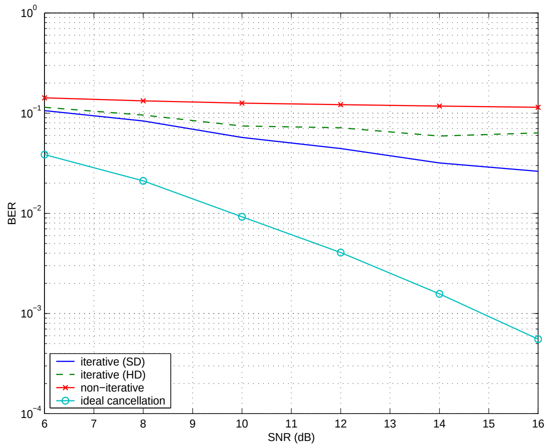


Fig. 3. BER versus SNR (in dB); $L = K = 128$ (full loading) and $Q = 2$.

the performance when the system is under-loaded. Fig. 4 shows the BER performance for different values of K. For the iterative detector, 4 iterations are considered. The results show that when the system is lightly loaded (up to 50%, i.e., $K = 64$), the performance of the iterative detector can approach the ideal performance. As shown in Fig. 4, however, the difference between the BERs of the iterative receiver and the receiver with ideal cancellation increases with K. Thus, we can see that the proposed iterative detector cannot fully remove the interference, especially for a large K. This result is predicted by the EXIT chart in Fig. 2 (b).

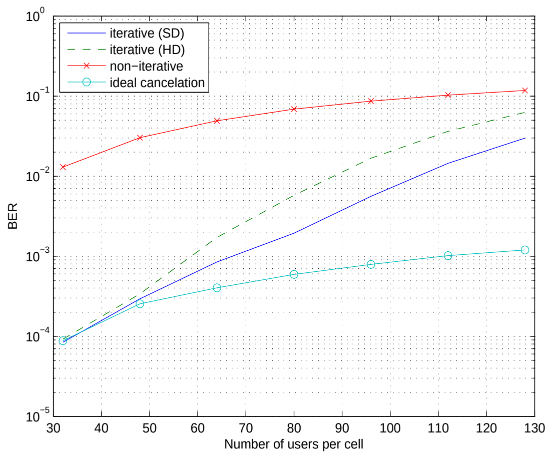


Fig. 4. BER versus K; $L = 128$, SNR = 14 dB, and $Q = 2$.

In general, the performance of the iterative detector is improved for more iterations. Fig. 5 shows the BER for different numbers of iterations. It is shown that more iterations are required for higher SNR. For example, when SNR = 6 dB, 4 iterations would be enough, while when SNR = 18 dB, more than 8 iterations are required for convergence.

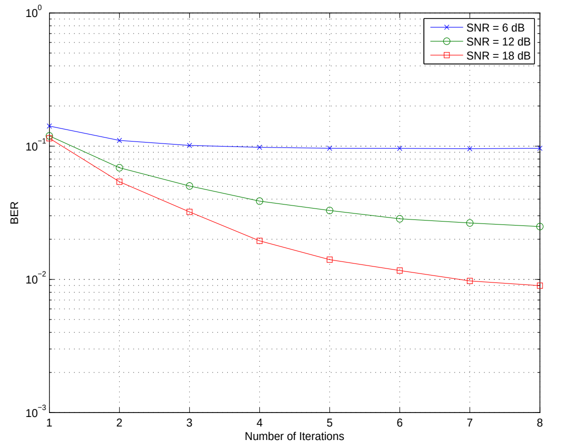


Fig. 5. BER versus the number of iterations; $L = 128$, $K = 128$, and $Q = 2$.

4. Iterative Receivers for Coded Signals

The performance of the iterative detector depends on the reliability of soft-decision. If the transmitted signals are coded, a better soft-decision is available after decoding. In this section, we derive iterative receivers based on the proposed iterative detector in Section 3 for coded signals, which are also called turbo receivers [22].

4.1. Coded Signals

In MC-CDMA systems, as coded OFDM systems, a codeword can be transmitted over L subcarriers such as $\{b_{q,k}, k = 1, 2, \ldots, K_q\}$ is a coded sequence from the qth BS. Then, the coded signal is transmitted after spreading as in (1). Thus, the bandwidth expansion by spreading can be given as $B_{\mathrm{sp},q} = L/K_q$. The bandwidth expansion by coding is the inverse of the code rate, denoted by R_q. Thus, the total bandwidth expansion becomes

$$B_q = \frac{L}{K_q}\frac{1}{R_q} = \frac{L}{K_q R_q}.$$

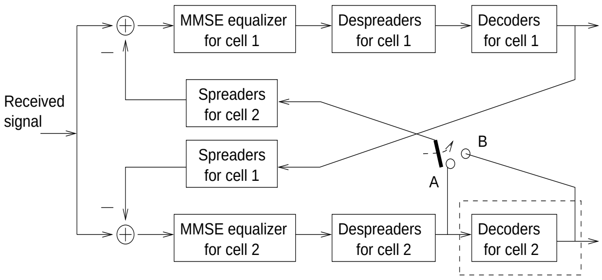


Fig. 6. Block diagram for the proposed iterative receiver for coded signals when $Q = 2$.

Given a fixed B_q, there can be the coding-spreading trade-off [23, 24]. However, in this paper, we focus on the design of the iterative receiver including channel decoders. For convenience, we assume that a convolutional code is used for coded signals $\{b_{q,k}\}$. For decoding, we consider the maximum a posteriori probability (MAP) decoder to obtain soft-decision.

4.2. Proposed Iterative Receiver and EXIT Charts

Based on the iterative detector in Section 3, an iterative receiver including channel decoding within iterations can be proposed as a generalization of the iterative detector. The structure of the proposed iterative receiver is shown in Fig. 6 when $Q = 2$. There can be two possible cases in deriving iterative receivers depending on the availability of the information of channel encoder. In the first case, only the information of channel encoder (and interleaver) for the desired signals is available. The channel decoder for the signals from BS 1 can be included within the iteration and the resulting iterative receiver is the case that the switch is connected to position "A" in Fig. 6. In this case, the extrinsic bit information of the signals from BS 1 is available from the decoder, while that from BS 2 is available from the detector. Thus, the extrinsic bit information of the signals from BS 2 would be less reliable than that from BS 1. For convenience, this receiver is referred to as Type-A iterative receiver.

In the second case, the information of channel encoder and interleaver for both the desired and interfering signals is available. In this case, the channel decoders for the signals form BSs 1 and 2 are included in the iterative receiver, and the extrinsic bit information would be available from channel decoders for both desired and interfering signals, and the resulting iterative receiver has the switch connected position "B" in Fig. 6. This receiver is referred to as Type-B iterative receiver. The performance of Type-B iterative receiver should be better than that of Type-A iterative receiver, because a better cancellation performance of the interfering signal is expected.

Fig. 7 shows the EXIT charts for Type-B iterative receiver when the SNR is 9 dB. The EXIT chart for a rate-half convolutional code with generator polynomial (5,7) is presented with the EXIT chart for the detector consisting of the MMSE equalizer, despreaders, and group interference canceller. The multipath fading channel described in Subsection 3.4 is used. Since the mutual information for the detector depends on the channel and the channel is random, the average and standard deviation of the mutual information are obtained and shown in the EXIT chart. From the EXIT charts in Fig. 7, we can see that the iterative receiver can converge although a worst case (average − standard deviation) is considered. It is noteworthy that the EXIT charts can provide an accurate performance prediction when the length of codeword is sufficiently large [19].

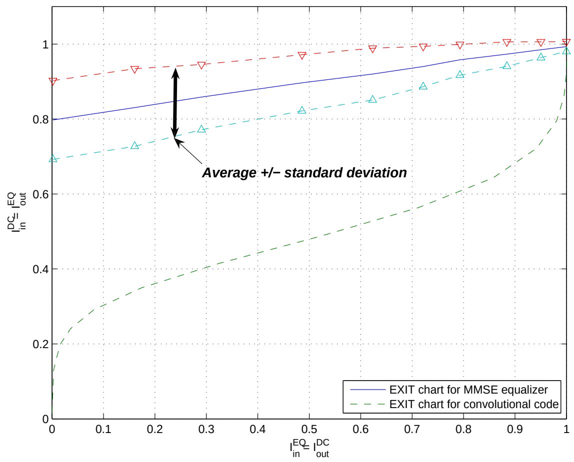


Fig. 7. EXIT charts for Type-B iterative receiver when the SNR = 9 dB.

4.3. Simulation Results

A rate-half convolutional code with generator polynomial (5,7) is used to generate coded signals. Two cells ($Q = 2$) are considered, and we assumed that the desired and interfering signals are equally strong, i.e., the SIR is 0 dB. The same frequency selective fading channels as in Subsection 3.4 are used.

Fig. 8 shows simulation results with various SNRs and full loading in each cell (i.e., $L = K$ and the

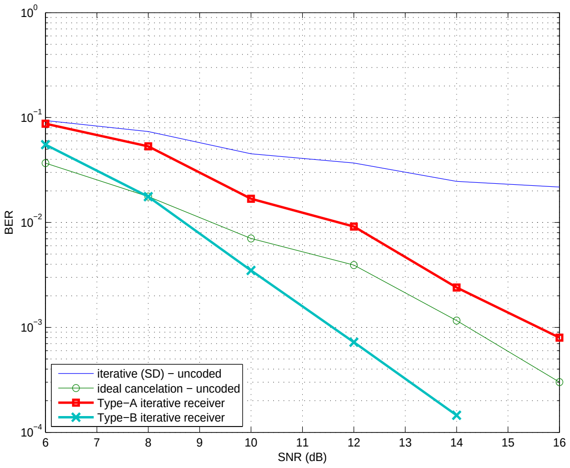


Fig. 8. Coded and uncoded BER versus SNR (in dB) with $Q = 2$, $L = K = 1024$ (full loading).

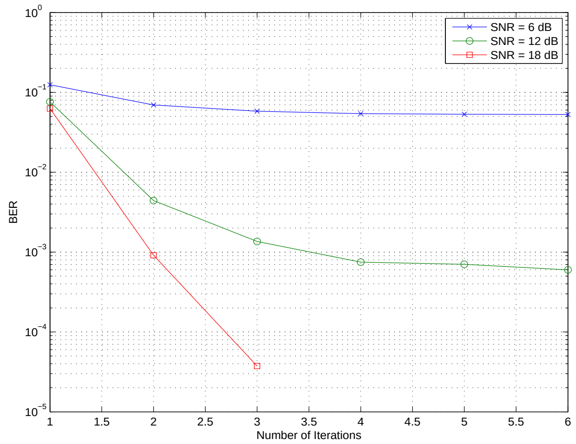


Fig. 9. Coded BER versus the number of iterations; $L = K = 1024$ and $Q = 2$.

total bandwidth expansion factor becomes $B = 2$). The number of iterations is set to 4. The fact that the SNR is equivalent to E_{b}/N_0 leads to a fair comparison with uncoded cases. Generally, it is shown that the performance of the iterative receivers for coded signals performs much better than that of the iterative detector (for uncoded signals). In addition, Type-B iterative receiver outperforms Type-A iterative receiver as expected.

Fig. 9 shows coded BER results for different numbers of iterations with $L = K = 1024$. We can see that the convergence can be faster if the gap between two EXIT charts is larger. Thus, the higher SNR, the faster convergence the iterative receiver can achieve.

From the results in Fig. 8 and Fig. 9, we can see that the iterative receiver requires a high SNR to properly mitigate the intercell interference and a fast convergence. Based on this observation, we can conclude that the iterative receiver can be effective if the cell size is sufficiently small, where a high SNR can be expected at cell edge.

With SNR = 14 dB, simulations are carried out for different system loading and results are shown in Fig. 10. It is shown that the iterative receiver can generally provide better performance.

5. Concluding Remarks

In this paper, we have derived iterative detector and receiver using SAIC techniques for MC-CDMA cellular systems. Through iterations, the intercell interference is cancelled with improved reliability, while the MMSE equalizer mitigates the intracell interference. Since the complexity of the MMSE equalizer and the intercell interference canceller is low, the resulting iterative detector and receiver have low complexity and are suitable for mobile terminals. From simulation results, we observed that the performance of the iterative receiver (which has channel decoders within the iteration) is improved as the SNR increases in the presence of the intercell interference (with an SIR of 0 dB). From this, we found that the iterative receiver can be effective when the cell size is sufficiently small where the SNR can be reasonable high at cell edge.

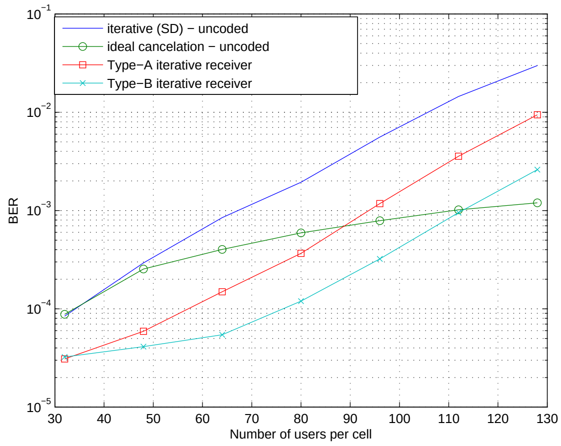


Fig. 10. Coded and uncoded BER versus K; $L = 128$, SNR = 14 dB, and $Q = 2$.

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