This full text paper was peer reviewed at the direction of IEEE Communications Society subject matter experts for publication in the IEEE GLOBECOM 2005 proceedings.

BER-Optimized Linear Parallel Interference Cancellers for Multicarrier DS-CDMA Systems

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***Abstract*— In this paper, we consider the design and bit error performance analysis of linear parallel interference cancellers (LPIC) for multicarrier (MC) DS-CDMA systems. We propose an LPIC scheme where we estimate (and cancel) the multiple access interference (MAI) based on the soft decision outputs on individual subcarriers, and the interference cancelled outputs on different subcarriers are combined to form the final decision statistic. We scale the MAI estimate on individual subcarriers by a weight before cancellation. In order to choose these weights optimally, we derive exact closed-form expressions for the bit error rate (BER) at the output of different stages of the LPIC, which we minimize to obtain the optimum weights for the different stages. We show that the proposed BER-optimized weighted LPIC scheme performs better than the MF detector and the conventional LPIC scheme (where the weights are taken to be unity), and close to the decorrelating detector.**

Keywords – *Linear PIC, multicarrier DS-CDMA, optimum weights*

I. Introduction

Multicarrier approach in CDMA offers several advantages including robustness in fading/interference, operation at lower chip rates, etc. Because of their potential to remove multiple access interference (MAI) and increase system capacity, interference cancellation techniques applied to multicarrier direct-sequence CDMA (MC DS-CDMA) are of interest [1]-[4]. Multistage successive/parallel interference cancellers are attractive owing to their implementation simplicity and good performance [5]. The MAI estimates for these cancellers can be obtained using the soft values of the decision statistics from the previous stages, in which case the cancellers are termed as linear cancellers [4]-[9]. Linear cancellers have the advantages of implementation simplicity, analytical tractability, and good performance. Here, we focus on linear parallel interference cancellers (LPIC) for MC DS-CDMA systems.

The conventional way to realize LPIC schemes is to use unscaled values of the soft outputs from different users for MAI estimation. A known problem with this conventional LPIC (CLPIC) approach is that it can perform even worse than the matched filter (MF) detector (where cancellation is not done), particularly at low SNRs [6],[9]. This is because the MAI estimates obtained using unscaled values of soft outputs can become quite inaccurate under poor channel conditions (e.g., low SNRs) to such an extent that it may be better not to do cancellation. This problem can be alleviated by properly weighing (scaling) the MAI estimates before cancellation [6],[9]. A key question in this regard is how to choose these weights (scaling factors) for different stages of the LPIC. For the case of single carrier CDMA, the issue of the choice of the weights in LPIC has been addressed in [10] for AWGN, and in [9] for Rayleigh fading and diversity channels.

Here, we propose a weighted LPIC (WLPIC) scheme for a MC DS-CDMA system, where we scale the MAI estimate on individual subcarriers by a weight before cancellation. One way to optimally choose the weights is to derive analytical expressions for the average SIR at the output of the IC stages as a function of the weights, and maximize these SIR expressions to obtain the optimum weights for different stages, as done in [9] for single carrier CDMA. However, for the MC DS-CDMA scheme we consider in this paper, the instantaneous SIR expression at the combined output from multiple carriers in the system is such that the unconditioning on the fade variables to obtain the average SIR in closed-form is difficult. An alternate approach to obtain the optimum weights can be to derive expressions for the average BER at the output of each IC stage of the MC DS-CDMA system in terms of the weights, and choose those weights that minimize this average BER. A new contribution in this paper, in this context, is that we are able to derive exact closed-form expressions for the average BER for different stages of the LPIC scheme for MC DS-CDMA, which we minimize and obtain the optimum weights for different stages. We point out that the BER analysis does not resort to Gaussian approximation of the interference. We show that the proposed BER-optimized weighted LPIC scheme for MC DS-CDMA performs better than the MF detector and the conventional LPIC scheme (where the weights are unity) and close to the decorrelating detector.

II. System Model

We consider a K-user synchronous multicarrier DS-CDMA system (an asynchronous system can be considered likewise). Figure 1 shows the transmitter of the kth user [1]. M is the number of subcarriers, and $c_{k,i}(t)$ is the spreading waveform of the kth user on the ith subcarrier. The number of chips per bit on each subcarrier is N. The channel coefficients $h_k^{(i)}$, $i = 1, 2, \cdots, M$, are assumed to be i.i.d. complex Gaussian r.v's (i.e., fade amplitudes are Rayleigh distributed) with zero mean and $E\left[\left(h_{kI}^{(i)}\right)^2\right] = E\left[\left(h_{kQ}^{(i)}\right)^2\right] = 1$, where $h_{kI}^{(i)}$ and $h_{kQ}^{(i)}$ are the real and imaginary parts of $h_k^{(i)}$. It is assumed that the channel is frequency non-selective on each subband and fades are independent from one subband to the other. The MC DS-CDMA receiver with the proposed weighted LPIC scheme is shown in Fig. 2.

This work was supported in part by the Swarnajayanti Fellowship, Department of Science and Technology, New Delhi, Government of India, under Project Ref: No.6/3/2002-S.F.

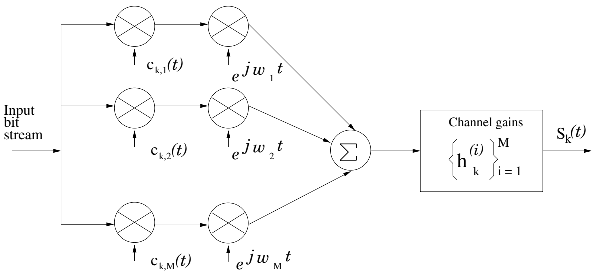

Fig. 1. Multicarrier DS-CDMA transmitter of kth user

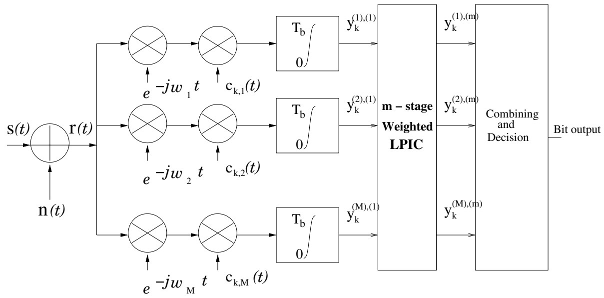

Fig. 2. Multicarrier DS-CDMA receiver with weighted LPIC

Let $\mathbf{y}^{(i)} = \left(y_1^{(i)}, y_2^{(i)}, ..., y_K^{(i)}\right)^T$, where T denotes the transpose operator, denote the K-length received signal vector on the ith subcarrier; i.e., $y_k^{(i)}$ is the output of the kth user's matched filter on the ith subcarrier. Assuming that the inter-carrier interference is negligible, the K-length received signal vector on the ith subcarrier $\mathbf{y}^{(i)}$ can be written in the form

$$\mathbf{y}^{(i)} = \mathbf{C}^{(i)}\mathbf{H}^{(i)}\mathbf{b} + \mathbf{n}^{(i)}, \tag{1}$$

where $\mathbf{C}^{(i)}$ is the $K \times K$ cross-correlation matrix on the ith subcarrier, given by

$$\mathbf{C}^{(i)} = \begin{bmatrix} 1 & \rho_{12}^{(i)} & \cdots & \rho_{1K}^{(i)} \\ \rho_{21}^{(i)} & 1 & \cdots & \rho_{2K}^{(i)} \\ \vdots & \vdots & \ddots & \vdots \\ \rho_{K1}^{(i)} & \rho_{K2}^{(i)} & \cdots & 1 \end{bmatrix}, \tag{2}$$

where $\rho_{lj}^{(i)}$ is the correlation coefficient between the signature waveforms of the lth and the jth users on the ith subcarrier. $\mathbf{H}^{(i)}$ represents the $K\times K$ channel coefficient matrix,

$$\mathbf{H}^{(i)} = diag\left\{h_1^{(i)}, h_2^{(i)}, \cdots, h_K^{(i)}\right\}. \tag{3}$$

The K-length data vector $\mathbf{b}$ is given by

$$\mathbf{b} = \begin{bmatrix} A_1b_1 & A_2b_2 & \cdots & A_Kb_K \end{bmatrix}^T, \tag{4}$$

where A_k denotes the transmit amplitude and $b_k \in \{+1, -1\}$ denotes the data bit of the kth user, and $[\cdot]^T$ denotes the transpose operator. The K-length noise vector $\mathbf{n}^{(i)}$ is given by

$$\mathbf{n}^{(i)} = \begin{bmatrix} \left(n_1^{(i)}\right)^* & \left(n_2^{(i)}\right)^* & \cdots & \left(n_K^{(i)}\right)^* \end{bmatrix}^H, \tag{5}$$

where $n_k^{(i)}$ denotes the additive noise component of the kth user on the ith subcarrier, which is assumed to be complex Gaussian with zero mean with $E[n_k^{(i)}(n_j^{(i)})^*] = 2\sigma^2$ when $j = k$ and $2\sigma^2\rho_{kj}^{(i)}$ when $j \neq k$. Here, $[\cdot]^H$ denotes the Hermitian operator and $(\cdot)^*$ denotes the complex conjugate.

III. Weighted LPIC Scheme for MC DS-CDMA

In the proposed weighted LPIC scheme, we cancel weighted estimates of the MAI on individual subcarriers, and the interference cancelled outputs from all the subcarriers are combined to form the combined decision statistic. The interference cancellation performed on the ith subcarrier in the mth stage is explained as follows.

A. Interference Cancellation on ith Subcarrier in Stage-m

The estimate of the MAI from the jth interfering user to the desired user k on the ith subcarrier in the mth cancellation stage is scaled by a factor $w_{jk}^{(i),(m)}$ before cancellation. Specifically, the estimate of the MAI from the jth interfering user to the desired user k on the ith subcarrier in stage-m, $m > 1$, is obtained by multiplying $y_j^{(i),(m-1)}$ with $\rho_{jk}^{(i)}$ for all $j \neq k$ and summing them up, where $y_j^{(i),(m-1)}$ is the jth interfering user's soft output at the $(m-1)$th stage. That is, $\sum_{j\neq k} w_{jk}^{(i),(m)} \rho_{jk}^{(i)} y_j^{(i),(m-1)}$ is the weighted MAI estimate on the ith subcarrier in stage-m for the desired user k. Accordingly, the mth stage interference cancelled output on the ith subcarrier for the desired user k, $y_k^{(i),(m)}$, is given by

$$y_k^{(i),(m)} = y_k^{(i),(1)} - \sum_{j=1,\, j\neq k}^{K} w_{jk}^{(i),(m)} \rho_{jk}^{(i)} y_j^{(i),(m-1)}. \tag{6}$$

Note that both the conventional LPIC as well as the MF detector become special cases of the above weighted LPIC for $w_{jk}^{(i),(m)} = 1, \forall\, i, j, m$ and $w_{jk}^{(i),(m)} = 0, \forall\, i, j, m$, respectively. All the subcarrier outputs of the desired user are then coherently combined to get the combined output, $y_k^{(m)}$, as

$$y_k^{(m)} = \sum_{i=1}^{M} \left(h_k^{(i)}\right)^* y_k^{(i),(m)}. \tag{7}$$

The bit decision at the m-th stage output is then obtained as

$$\hat{b}_k^{(m)} = \text{sgn}\left(\text{Re}\left(y_k^{(m)}\right)\right). \tag{8}$$

For this weighted LPIC scheme, choice of the weights $w_{jk}^{(i),(m)}$ can be made based on maximizing the average SIR at the combined output or minimizing the average BER of each stage. For the MC DS-CDMA system considered in the above, the instantaneous SIR expression at the combined output from the multiple carriers in the system is such that the unconditioning on the fade variables to obtain the average SIR in closed-form is difficult. However, we could derive exact closed-form expressions for the average BER of the system which when minimized can give optimum weights.

B. Derivation of BER Expressions

Here, we derive the average BER expressions for the 2nd and 3rd stage outputs of the weighted LPIC described above.

1) 2nd stage output statistics and BER: From (6) and (7), the weighted interference cancelled output of the second stage (i.e., $m = 2$) for the desired user k can be written as

$$y_k^{(2)} = A_k b_k \sum_{i=1}^{M} \left|h_k^{(i)}\right|^2 \left(1 - \sum_{j=1,\, j\neq k}^{K} w_{jk}^{(i),(2)} \left(\rho_{jk}^{(i)}\right)^2\right) + I_2 + N_2, \quad (9)$$

where

$$I_2 = \sum_{i=1}^{M} \left(h_k^{(i)}\right)^* \left[\sum_{j=1,\, j\neq k}^{K} \left(1 - w_{jk}^{(i),(2)}\right) A_j b_j h_j^{(i)} \rho_{jk}^{(i)} - \sum_{j=1,\, j\neq k}^{K} w_{jk}^{(i),(2)} \rho_{jk}^{(i)} \sum_{\substack{l=1 \\ l\neq j,k}}^{K} \rho_{lj}^{(i)} A_l b_l h_l^{(i)} \right], \quad (10)$$

$$N_2 = \sum_{i=1}^{M} \left(h_k^{(i)}\right)^* \underbrace{\left[n_k^{(i)} - \sum_{j=1,\, j\neq k}^{K} w_{jk}^{(i),(2)} \rho_{jk}^{(i)} n_j^{(i)} \right]}_{\Delta_n}. \quad (11)$$

The terms I_2 and N_2 in (9) represent the interference and noise terms introduced in the 2nd stage output due to imperfect cancellation in using the soft output values from the first (i.e., MF) stage. Conditioned on the desired user's channel coefficients, it can be seen that I_2 and N_2 are independent Gaussian r.v's, each with mean zero. The conditional variances of I_2 are N_2 are denoted by $\sigma_{I_2}^2$ and $\sigma_{N_2}^2$, respectively. Since N_2 in (11) is a Gaussian r.v. with zero mean, $\sigma_{N_2}^2$ is

$$\sigma_{N_2}^2 = E\left[N_2 N_2^* \middle| h_k^{(1)}, \cdots, h_k^{(M)}\right] = \sum_{i=1}^{M} |h_k^{(i)}|^2 \sigma_{N_{(i,2)}}^2, \quad (12)$$

where

$$\sigma_{N_{(i,2)}}^2 = E\left[\Delta_n \Delta_n^*\right] = 2\sigma^2 \left(1 - 2\sum_{\substack{j=1 \\ j\neq k}}^{K} w_{jk}^{(i),(2)} \left(\rho_{jk}^{(i)}\right)^2 + \sum_{\substack{l=1 \\ l\neq k}}^{K} w_{lk}^{(i),(2)} \rho_{lk}^{(i)} \sum_{\substack{j=1 \\ j\neq k}}^{K} w_{jk}^{(i),(2)} \rho_{jk}^{(i)} \rho_{jl}^{(i)} \right), \quad (13)$$

where we have used $E[n_k^{(i)} (n_j^{(i)})^*] = 2\sigma^2 \rho_{kj}^{(i)}$, for $j \neq k$ and $2\sigma^2$ for $j = k$. To derive $\sigma_{I_2}^2$, note that I_2 in (10) can be rearranged in the form $I_2 = \sum_{i=1}^{M} \left(h_k^{(i)}\right)^* \beta$, where

$$\beta = \sum_{l=1,\, l\neq k}^{K} A_l b_l h_l^{(i)} \left(\left(1 - w_{lk}^{(i),(2)}\right) \rho_{lk}^{(i)} - \sum_{\substack{j=1 \\ j\neq k,l}}^{K} w_{jk}^{(i),(2)} \rho_{jk}^{(i)} \rho_{jl}^{(i)} \right). \quad (14)$$

Since $h_l^{(i)}$'s are independent complex Gaussian with zero mean and b_l's do not affect the statistics of I_2, $\sigma_{I_2}^2$ is given by

$$\sigma_{I_2}^2 = E\left[I_2 I_2^* \middle| h_k^{(1)}, \cdots, h_k^{(M)}\right] = \sum_{i=1}^{M} |h_k^{(i)}|^2 \sigma_{I_{(i,2)}}^2 \quad (15)$$

where $\sigma_{I_{(i,2)}}^2 = E\left[\beta\beta^*\right]$, which can be obtained as

$$\sigma_{I_{(i,2)}}^2 = \sum_{l=1,\, l\neq k}^{K} 2A_l^2 \left(\left(1 - w_{lk}^{(i),(2)}\right) \rho_{lk}^{(i)} - \sum_{\substack{j=1 \\ j\neq k,l}}^{K} w_{jk}^{(i),(2)} \rho_{jk}^{(i)} \rho_{jl}^{(i)} \right)^2. \quad (16)$$

Now, the bit error analysis of the decision rule in (8) can be carried out by conditioning with respect to the transmitted bits and the channel coefficients. The binary coefficients corresponding to the transmitted other user bits can be dropped since they do not affect the distribution of the decision variable. Accordingly, the probability of error conditioned on the channel fade coefficients of the desired user k at the 2nd stage output is given by

$$P_{e,\mathbf{h}_k}^{(2)} = \Pr\left(\operatorname{sgn}\left(\operatorname{Re}\left(y_k^{(2)}\right)\right) < 0 \,\middle|\, b_k = 1, h_k^{(1)}, \ldots, h_k^{(M)} \right). \quad (17)$$

The above equation simplifies to

$$P_{e,\mathbf{h}_k}^{(2)} = Q(Y), \quad (18)$$

where Y is of the form

$$Y = \frac{\sum_{i=1}^{M} X_i}{\sqrt{\sum_{i=1}^{M} q_i X_i}}, \quad (19)$$

where, from the first term in (9) it can be seen that

$$X_i = A_k \left|h_k^{(i)}\right|^2 \left(1 - \sum_{j=1,\, j\neq k}^{K} w_{jk}^{(i),(2)} \left(\rho_{jk}^{(i)}\right)^2\right), \quad (20)$$

and that X_i's are exponential r.v's with mean $\overline{X}_i$, given by (since $E\left[\left|h_k^{(i)}\right|^2\right] = 2$)

$$\overline{X}_i = 2A_k \left(1 - \sum_{j=1,\, j\neq k}^{K} w_{jk}^{(i),(2)} \left(\rho_{jk}^{(i)}\right)^2\right), \quad (21)$$

and q_i's are given by

$$q_i = \frac{\sigma_{I_{(i,2)}}^2 + \sigma_{N_{(i,2)}}^2}{\overline{X}_i}. \quad (22)$$

Now, unconditioning (18) over X_i's the average bit error probability for user k at the 2nd stage output is

$$P_e^{(2)} = E[Q(Y)], \quad (23)$$

which can be derived by using the relation [5]

$$E\left[Q(Y)\right] = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{\infty} F_Y(y) e^{-y^2/2} dy, \quad (24)$$

where $F_Y(y)$ is the CDF of Y defined in (19). Hence we need the CDF of Y in order to derive an expression for $P_e^{(2)}$. We have derived the CDF of Y to be (derivation is given in [11]):

$$F_Y(y) = \sum_{j=1}^{M} \xi_j \sum_{l=1}^{M} \frac{\alpha_j(l)}{q_j \overline{X}_j} \left\{ \frac{q_j \overline{X}_j^2}{\zeta_j(l)} \left[1 - \exp\left(-\frac{y^2 q_j}{\overline{X}_j}\right)\right] - \left(\frac{\overline{X}_j - \zeta_j(l)}{\zeta_j^2(l)}\right) \mathcal{J}_1\left(y, q_j, G, B\right) \right\}, \quad (25)$$

where

$$G = \frac{1}{\zeta_j(l)}, \quad B = \frac{1}{q_j \overline{X}_j} - \frac{1}{q_j \zeta_j(l)}, \quad (26)$$

$$\xi_j = \prod_{i=1, i\neq j}^{M} \frac{q_j \overline{X}_j}{q_j \overline{X}_j - q_i \overline{X}_i}, \quad (27)$$

$$\zeta_j(i) = \begin{cases} \overline{X}_j & \text{if } i = j \\ \frac{\overline{X}_i \overline{X}_j (q_j - q_i)}{q_j \overline{X}_j - q_i \overline{X}_i} & \text{if } i \neq j, \end{cases} \quad (28)$$

$$\mathcal{J}_1\left(y, q_j, G, B\right) = \frac{q_j}{G(G+q_j B)}\left[1 - e^{-(q_j G y^2 + q_j^2 B y^2)}\right] - \frac{2}{G}\mathcal{J}_2\left(-Gy, -B, q_j y\right), \quad (29)$$

$$\mathcal{J}_2(P_1, P_2, P_3) = \frac{e^{-\frac{P_1^2}{4P_2}}}{4P_2^{\frac{3}{2}}}\left\{\left(2\sqrt{P_2} e^{\frac{P_1^2}{4P_2}}\left[-1 + e^{P_3(P_1+P_2P_3)}\right]\right) + P_1\sqrt{\pi}\left(\text{Erfi}\left[\frac{P_1}{2\sqrt{P_2}}\right] - \text{Erfi}\left[\frac{P_1 + 2P_2P_3}{2\sqrt{P_2}}\right]\right)\right\}, \quad (30)$$

where $\text{Erfi}(x) = \text{erf}(jx)/j$, $j = \sqrt{-1}$ and $\text{erf}(x)$ is the standard error function defined for the real-valued x.

Substituting the CDF expression (25) into (24) and carrying out the integration, we get the expression for $P_e^{(2)}$, in closed-form, as (derivation is given in [11])

$$P_e^{(2)} = \sum_{j=1}^{M}\xi_j \sum_{l=1}^{M}\alpha_j(l)\left\{\frac{\overline{X}_j}{\zeta_j(l)}\left(\frac{1}{2} - \frac{1}{2}\sqrt{\frac{\overline{X}_j}{\overline{X}_j + 2q_j}}\right) - \left(\frac{\overline{X}_j - \zeta_j(l)}{\zeta_j(l)}\right)\left(\frac{1}{2} - \frac{1}{2}\sqrt{\frac{\overline{X}_j}{\overline{X}_j + 2q_j}} - 2V_j(l)\right)\right\}, \quad (31)$$

where $V_j(l)$ is given by

$$V_j(l) = \frac{1}{2}\left[-\frac{q_j\zeta_j(l)}{\sqrt{1 + \frac{2q_j}{\zeta_j(l)}}} + \zeta_j^2(l)\sqrt{1 + \frac{2q_j}{\zeta_j(l)}} - \zeta_j^2(l)\right], \quad \text{if } B = 0, \quad (32)$$

$$V_j(l) = \frac{1}{4B}\left[1 - \sqrt{\frac{\overline{X}_j}{\overline{X}_j + 2q_j}} - \frac{1}{\zeta_j(l)}\left(\frac{1}{\sqrt{\left(\frac{2B\zeta_j(l)Z^3}{1+2B\zeta_j(l)q_j}\right)^2 + 1}} - \frac{1}{\sqrt{\left(2B\zeta_j(l)Z^3\right)^2 + 1}}\right)\right], \quad \text{if } B \neq 0. \quad (33)$$

where $Z = \sqrt{1 - \frac{1}{2B\zeta_j^2(l)}}$.

2) 3rd stage output statistics and BER: Following similar steps carried out for deriving the 2nd stage output statistics and BER in the above, we have derived the expressions for $\overline{X}_i$, $\sigma^2_{I_{(i,3)}}$ and $\sigma^2_{N_{(i,3)}}$ corresponding to the the third stage outputs. These expressions for $\overline{X}_i$, $\sigma^2_{I_{(i,3)}}$ and $\sigma^2_{N_{(i,3)}}$ for the 3rd stage output are given in [11]. Using these expressions the BER at the 3rd stage output can be derived similar to the 2nd stage BER derivation given before.

3) Optimum weights for 2nd and 3rd stages: As can be seen, (31) gives the bit error rate as a function of the weights used in the cancellation. The optimum weights for the 2nd and 3rd stages can be obtained by numerically minimizing their corresponding BER expressions. For a given stage, instead of obtaining different weights for different subcarriers and for different users which requires considerable complexity in the numerical optimization, we can consider the simplified case of using the same weight for all subcarriers and for all users (i.e., $w_{jk}^{(i),(m)} = w_k^{(m)}, \forall i, j$), the optimization of which requires much less complexity. As we will see in the next section, even this simplified scheme which uses the optimized weights $w_k^{(m)}$ gives better performance than other detectors.

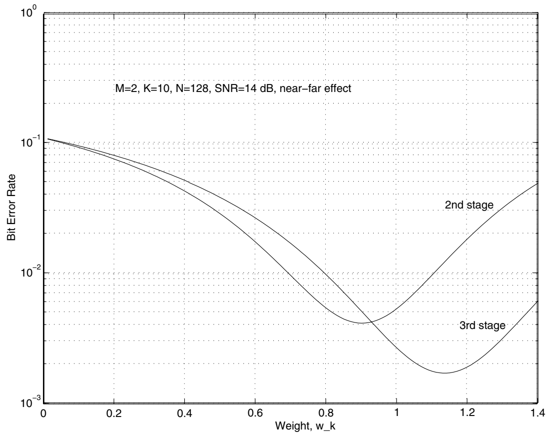

Fig. 3. BER at the 2nd and 3rd stage outputs as a function of the weights in the weighted LPIC scheme. $M = 2, K = 10, N = 128$, average SNR = 14 dB. Near-far effect: $A_2/A_1 = A_4/A_1 = A_5/A_1 = 10$.

In Fig. 3, we illustrate the variation of the BER performance at the 2nd and 3rd stage outputs as a function of the weights, $w_k^{(m)}$ for $m = 2, 3$, for the simplified scheme for $M = 2$, $K = 10$, $N = 128$, SNR=14 dB, and with near-far effect where users 2, 4, and 5 transmit with 10 times more amplitude than the desired user 1. From Fig. 3, it can be observed that $w_{k,opt}^{(m)}$ is about 0.9 for $m = 2$ and about 1.15 for $m = 3$. Further, it can be observed that the minimum achievable BER (corresponding to the optimum weights) is significantly better than that of the conventional LPIC (for which $w_k^{(m)} = 1$) and the MF detector (for which $w_k^{(m)} = 0$).

IV. Results and Discussions

In this section, we present numerical results of the BER performance of the proposed weighted LPIC for MC DS-CDMA. We computed the analytical BER performance for the 2nd and 3rd stages of the weighted LPIC for different number of subcarriers, M, and number of users K. We used random binary sequences of length N as the spreading sequences on each subcarrier. In all the performance plots, NM is taken to be 256 (i.e., the number of chips per bit on each subcarrier is chosen such that the total system bandwidth is fixed regardless of the number of subcarriers used). We take the number of subcarriers M to be 1, 2, and 4. We also keep the total transmit power to be the same irrespective of the number of subcarriers used. BER performance is computed in near-far scenarios where some users transmit with higher powers than the desired user. We take user 1 as the desired user.

In Fig. 4, we present the BER performance of the desired user at the 2nd stage output of the weighted LPIC schemes as a function of average SNR (given by A_k^2/σ^2), for $M = 2, K = 5, N = 128$ with near-far effect such that $A_2/A_1 = 15, A_3/A_1 = 10, A_4/A_1 = 20, A_5/A_1 = 25$. We show three plots for the weighted LPIC schemes, *viz.*, *i)* WLPIC-I, which corresponds to optimizing the same weight for all users and for all subcarriers (i.e., $w_{k,opt}^{(m)}$), *ii)* WLPIC-II, which corresponds to optimizing the same weight for all users but different weights for different subcarriers, (i.e., $w_{k,opt}^{(i),(m)}$), and *iii)* WLPIC-III, which corresponds to optimizing different weights for different users but same weights for all subcarriers (i.e., $w_{jk,opt}^{(m)}$). For the purpose of comparison, we also plot

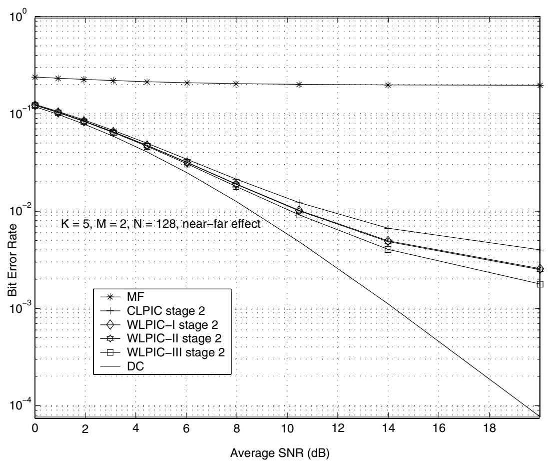

Fig. 4. BER vs average SNR performance at the 2nd stage output of the weighted LPIC schemes-I,II,III. $M = 2, K = 5, N = 128$.

the BER performance of other detectors including the MF detector, conventional LPIC and decorrelating detector.It can be noted that in terms of optimization complexity WLPIC-III is most complex and WLPIC-I is least complex, and in terms of BER performance WLPIC-III is expected to perform best. As expected, from Fig. 4, it can be observed that the WLPIC-III scheme performs better than the WLPIC-II and WLPIC-I schemes, and the performance of the WLPIC-I and WLPIC-II schemes are very close. Even the WLPIC-I scheme, which has the least optimization complexity among all, clearly performs better than the MF detector as well as the conventional LPIC. This is expected since in MF detector there is no cancellation, whereas, in conventional LPIC there is cancellation but the weights are not optimum.

In Fig. 5, we plot the BER performance of the WLPIC-I scheme at the 2nd and 3rd stage outputs for $M = 2, K = 16, N = 128$ with near-far effect such that $A_2/A_1 = A_4/A_1 = A_5/A_1 = 10$. It can be seen that the performance at the 3rd stage output of the WLPIC-I scheme is quite close to that of the decorrelating detector. We have also evaluated the BER performance through simulations and compared with the analytical results. The analytical and simulation results matched as there are no approximations involved in the analysis. Figure 6 shows the performance of the WLPIC-I scheme and the decorrelating detector for different number of subcarriers, $M = 1, 2, 4$ for $NM = 256$, average SNR=10 dB with near-far effect. The performance of $M = 4$ is better than $M = 2$ and $M = 1$ because of frequency diversity effect. Again, the performance of the WLPIC-I scheme is close to that of the decorrelating detector.

V. Conclusions

We presented the design and BER analysis of a weighted LPIC scheme for multicarrier DS-CDMA systems. In the proposed scheme, partial multiuser interference is cancelled at each stage, which is controlled by a weight that is optimized based on minimizing the bit error rate per stage. The bit error rate at each stage is computed based on an exact closed-form formula, which has relatively low complexity. We showed that the proposed BER-optimized wighted LPIC scheme performs significantly better than the MF detector and quite close to the decorrelating detector.

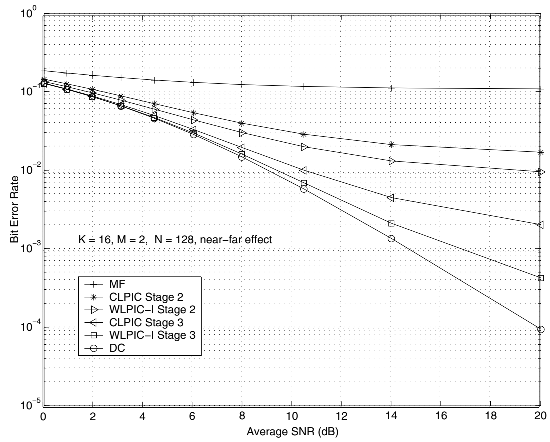

Fig. 5. BER vs average SNR performance at the 2nd and 3rd stage outputs of the weighted LPIC-I scheme. $K = 16, M = 2, N = 128$.

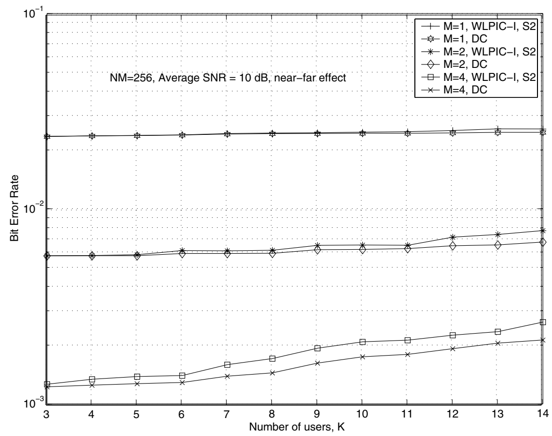

Fig. 6. BER vs number of users, K, performance at the 2nd stage output of the weighted LPIC-I scheme for different number of subcarriers, $M = 1, 2, 4$, $NM = 256$, average SNR = 10 dB.

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