LINEAR MULTIUSER PRECODING WITH TRANSMIT ANTENNA DIVERSITY FOR DS/CDMA SYSTEMS

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ABSTRACT

For the downlink of Direct-Sequence (DS) Code-Division Multiple Access (CDMA) systems, the demand for high data rate transmission and the constraint of low-complexity mobile stations have led to the investigation of transmitter (Tx)-based multiple access interference (MAI) cancellation techniques. In this paper, we propose two linear MAI cancellation techniques, PreRAKE Linear Decorrelating Precoding (PreRAKELDP) and Multipath Decorrelating Precoding (MDP). The multipath diversity and antenna diversity are utilized in these transmitter designs. In the PreRAKELDP method, the PreRAKE combiner is preceded by a linear precoding filter. We illustrate two different transmit power control strategies for PreRAKELDP, the zero-forcing (ZF)-based with individual user power scaling and the minimum-mean-square-error (MMSE)-based under total power constraint. The MMSE solution achieves lower error probability while the ZF solution has lower computational complexity. Compared to the PreRAKELDP, the MDP method employs a simpler precoding filter but has modestly higher error rate. By theoretical analysis and numerical experiments, it is shown that the PreRAKELDP and MDP outperform the existing linear transmitter precoding methods with similar complexity. In addition, the two proposed schemes result in very simple channel-independent receiver structure. Thus, they are suitable for the downlink of CDMA systems.

I. INTRODUCTION

The next-generation CDMA systems are required to meet the increasing demands for wireless multimedia services, such as Internet and video conferencing. For these applications, the improvement of downlink capacity becomes more important than that of uplink. However, for the downlink (from base station (BS) to mobile station (MS)), the size and power consumption limitations of the MSs restrict the complexity of the receiver structures; therefore the well-developed multiuser detection techniques [1] are not suitable for downlink CDMA. In recent years, various transmitter (Tx)-based techniques have been proposed to increase the downlink capacity. Diversity techniques such as the Pre-RAKE multipath diversity combining [5] and transmit antenna diversity techniques [8, 9, 16] can improve data rate without expanding bandwidth; however, they cannot reduce the multipath-induced multiple access interference (MAI), which is a major limitation to the capacity of CDMA systems.

The purpose of multiuser precoding is to cancel the MAI while keeping the mobile user receiver as simple as possible. The class of linear precoding methods is efficient and easy to implement. Several Zero-Forcing (ZF)-based linear precoders have been proposed for frequency-selective fading channels. The transmitter precoding with RAKE receiver in [2] and the decorrelating prefilters in [3] can completely cancel the MAI, but their performance is degraded by transmit power scaling. The Pre-RAKE Decorrelating Precoder (Pre-RDD) in [4] achieves the performance of the receiver (Rx)-based RAKE Decorrelating Detector (RDD), the optimal linear detector with unknown user energy [10], and outperforms the methods in [2, 3] by jointly using the linear MAI pre-decorrelation and Pre-RAKE multipath combiner. A common problem of the precoding schemes in [2-4] is that the precoding filter requires the inverse operation of CSI-dependent matrices, which has to be performed continuously as the channel fading coefficients vary. Thus these methods require high computation load, especially when the channel fading changes fast. A simpler precoder proposed in [6] employs a CSI-independent precoding filter; but its performance is poorer than that of the Pre-RDD. The hybrid design which combines this precoder with transmit antenna array, named Space-Time PreRAKE multiuser precoder (STPR MUP), was described in [7].

In this paper, we propose two novel linear precoding methods, PreRAKE Linear Decorrelating Precoding (PreRAKELDP) and Multipath Decorrelating Precoding (MDP). In the PreRAKELDP transmitter, the preRAKE combiner is preceded by a linear precoding filter. Based on two different transmit power control strategies, we propose

This work was supported by NSF grant CCR-0312294 and ARO grants DAAD 19-01-1-0638 and W911NF-05-1-0311.

two optimal solutions for the precoding filter. One is the zero forcing (ZF)-based filter with *individual user power scaling*, in which each user's average transmit power is normalized by a power scaling factor. The other is the minimum mean square error (MMSE)-based filter under the *total transmit power constraint*. (Note that if the individual-user transmit power constraint rather than the total transmit power constraint is enforced, the MMSE solution is identical to the ZF solution [2, 3].) When these two PreRAKELDP solutions are compared, the ZF precoder is much simpler while the MMSE precoder has better performance.

In the MDP transmitter, the multipath diversity combining is incorporated into the MAI cancellation process. Its decorrelating filter is independent of the channel fading coefficients and only determined by users' signature sequences. As a result, the computational complexity of the MDP is lower than that of the PreRAKELDP, Pre-RDD and the precoders in [2, 3].

We demonstrate that the proposed precoding methods compare favorably to previously investigated linear multiuser precoding and detection designs, while offering the desired performance-complexity trade-off. We only consider the synchronous system in the following discussion. The synchronous assumption is usually justified for the downlink channel since users employ orthogonal signature sequences, the chip interval is much smaller than the symbol interval and the multipath delay spread is on the order of a few chip intervals [3]. However, by employing the spectral factorization [17], the derivations in this paper can be easily extended to the asynchronous case. Another crucial assumption for Tx-based interference cancellation methods is that the transmitter has the knowledge of channel conditions. In practice, this information can be obtained via feedback channels, and requires long range prediction for rapidly varying fading channels [18].

In the following section, we briefly describe the mathematical model of frequency-selective fading channels for downlink CDMA with multiple transmit antennas at BS. In section III and IV, the PreRAKELDP and MDP schemes are explained respectively and compared with other methods. The simulation results in section V provide more performance comparisons among different linear precoding techniques and power control methods. We present the concluding remarks in section VI.

II. SYSTEM MODEL

Consider a synchronous downlink DS/CDMA system with K active mobile users in a cell. The BS employs L transmit antennas, and the MS for each user employs single receive antenna. The BPSK data symbol and signature sequence for the ith user in the bit interval $[0, T)$ are denoted by b_i and $s_i(t)$, respectively, $i = 1, 2, \ldots, K$. The amplitude of b_i is A_i. The channels from the antennas at the BS to the MSs are each subject to frequency-selective Rayleigh fading with N resolvable paths. If the baseband-equivalent signal transmitted from the lth antenna in the data interval of interest is denoted by $x^{(l)}(t)$, $l = 1, 2, \ldots, L$, then the signal received by the ith user can be expressed as $r_i(t) = \sum_{n=0}^{N-1}\sum_{l=1}^{L} c_{i,n}^{(l)} x^{(l)}(t - nT_c) + n_i(t)$, where T_c is the chip duration, $c_{i,n}^{(l)}$ is the channel gain coefficient of the nth path from the lth transmit antenna to the ith user's receiver and $n_i(t)$ is the complex AWGN with power spectral density N_0, for $\forall\ i = 1, 2, \ldots, K$ and $n = 0, 1, \ldots, N-1$. The channel coefficients $c_{i,n}^{(l)}$ are modeled as independent and identically distributed (i.i.d) Rayleigh fading random variables with $E\{\sum_{n=0}^{N-1} | c_{i,n}^{(l)} |^2\} = 1$. ($E\{\cdot\}$ is the expectation.) Without any Tx-based processing, the signal received by the ith user is given by

$$r_i(t) = \sum_{n=0}^{N-1}\sum_{l=1}^{L}\sum_{k=1}^{K} c_{i,n}^{(l)} b_k s_k(t - nT_c) + n_i(t). \qquad (1)$$

Obviously, even if the spreading sequences for different users are orthogonal, in each user's received signal, there is interference from other users and self-interference due to multipath fading.

For the convenience of derivation, define the channel gain row vector for the ith user corresponding to the lth transmit antenna as $\mathbf{c}_i^{(l)} = [c_{i,0}^{(l)}, c_{i,1}^{(l)}, \ldots, c_{i,N-1}^{(l)}]$; and for all K users, define the K-row $(N\times K)$-column channel gain matrix $\mathbf{C}^{(l)} = diag\{\mathbf{c}_1^{(l)}, \mathbf{c}_2^{(l)}, \ldots, \mathbf{c}_K^{(l)}\}$, $\forall l = 1, 2, \ldots, L$. The data vector and signature sequence vector for all K users are defined as $\mathbf{b} = [b_1, b_2, \ldots, b_K]^T$ and $\mathbf{s}(t) = [s_1(t), s_2(t), \ldots, s_K(t)]^T$, respectively. To avoid the transmit power increase incurred by channel gain coefficients, we use the normalized values of channel gain coefficients in precoding. For the ith user, the channel gain normalization factor is defined as

$$S_{ci} = \left(\sum_{l=1}^{L}\sum_{n=0}^{N-1} \left| c_{i,n}^{(l)} \right|^2 \right)^{-1/2}. \qquad (2)$$

Then the normalized channel gain coefficient is $\hat{c}_{i,n}^{(l)} = S_{ci} c_{i,n}^{(l)}$. Define the normalized vector $\hat{\mathbf{c}}_i^{(l)} = S_{ci} \cdot \mathbf{c}_i^{(l)}$ and matrix $\hat{\mathbf{C}}^{(l)} = diag\{\hat{\mathbf{c}}_1^{(l)}, \hat{\mathbf{c}}_2^{(l)}, \ldots, \hat{\mathbf{c}}_K^{(l)}\}$, respectively. By representing the normalization factors with a diagonal $K\times K$ matrix $\mathbf{S_c} = diag\{S_{c1}, S_{c2}, \ldots, S_{cK}\}$, we obtain the relation

$\hat{\mathbf{C}}^{(l)} = \mathbf{S_c}\mathbf{C}^{(l)}$.

The correlation between the delayed signature waveforms for user i and user k is denoted by

$$R_{i,k}^{m} = \int_0^T s_i(t)s_k(t+mT_c)dt, \quad (3)$$

where $m = -(N-1),\ldots, N-1$. For $\forall$ i, $k = 1, 2, \ldots, K$, define the $N\times N$ matrix $\mathbf{R}_{i,k}$ with the jth row equal $[R_{i,k}^{1-j}, R_{i,k}^{2-j},\ldots, R_{i,k}^{N-j}]$, $j = 1, 2, \ldots, N$. Then for all K users, the ($K\times N$)-row and ($K\times N$)-column correlation matrix $\mathbf{R}$ is constructed by superimposing the $K\times K$ non-overlapping submatrices $\mathbf{R}_{i,k}$, i, $k = 1, 2, \ldots, K$.

III. PRERAKELDP WITH MULTIPLE TX ANTENNAS

A. ZF Solution with Individual Power Scaling

The transmitter diagram of the PreRAKELDP with multiple Tx antennas is shown in Fig. 1. The linear MAI pre-cancellation process actually increases the required transmit power. To offset the power increase, we scale the amplitude of the transmit signal b_i by a factor S_{fi}, $\forall$ $i = 1, 2, \ldots, K$. For all users, define the diagonal scaling factor matrix $\mathbf{S_f} = diag\{S_{f1}, S_{f2}, \ldots, S_{fK}\}$. Suppose the decorrelating (ZF) filter is a $K\times K$ matrix $\mathbf{G}$, and the result of power scaling and decorrelating is denoted by vector $\mathbf{w}$. Then $\mathbf{w} = [w_1, w_2,\ldots, w_K]^T = \mathbf{G}\mathbf{S_f}\mathbf{b}$. The transmitted signal from the lth antenna, $x^{(l)}(t)$, is generated by passing the decorrelating filter output through the lth branch of the preRAKE combiner

$$x^{(l)}(t) = \sum_{k=1}^{K}\sum_{n=0}^{N-1} w_k \hat{c}_{k,n}^{(l)*} s_k(t-(N-1-n)T_c). \quad (4)$$

At the ith user receiver, the equivalent baseband received signal is given by

$$r_i(t) = \sum_{p=0}^{N-1}\sum_{l=1}^{L} c_{i,p}^{(l)} x^{(l)}(t-pT_c) + n_i(t). \quad (5)$$

A channel-gain-independent matched filter (MF) is used at the front end of each user's receiver. To obtain the largest received signal energy, the output of the matched filters should be sampled at the moment $t = (N-1)T_c$ [5]. The output of the ith user's MF is given by

$$y_i = \int_{(N-1)T_c}^{T+(N-1)T_c} r_i(t)s_i(t-(N-1)T_c)dt. \quad (6)$$

Substituting (3) and (5) into (6), we obtain

$$y_i = \sum_{k=1}^{K}\sum_{n=0}^{N-1}\sum_{m=-n}^{N-1-n}\sum_{l=1}^{L} c_{i,m+n}^{(l)} \hat{c}_{k,n}^{(l)*} w_k R_{i,k}^{m} + n_i, \quad (7)$$

where n_i is the filtered noise component with power $E\{n_i n_i^*\} = N_0$. Define vectors $\mathbf{y} = [y_1, y_2, \ldots, y_K]^T$

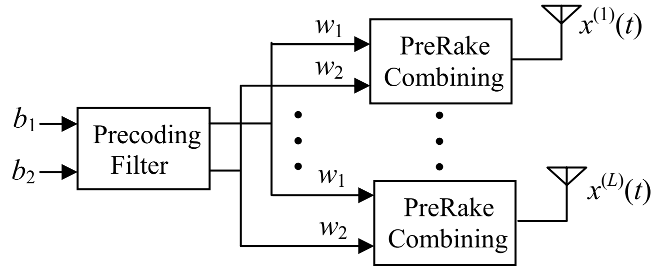

Fig. 1 PreRAKELDP Transmitter for a 2-user, *L*-Tx antenna System

and $\mathbf{n} = [n_1, n_2, \ldots, n_K]^T$. Then the vector of MF outputs of all K users

$$\mathbf{y} = \mathbf{S_c}^{-1}\left(\sum_{l=1}^{L}\hat{\mathbf{C}}^{(l)}\mathbf{R}\hat{\mathbf{C}}^{(l)H}\right)\mathbf{G}\mathbf{S_f}\mathbf{b} + \mathbf{n}. \quad (8)$$

Let

$$\hat{\mathbf{R}} = \sum_{l=1}^{L}\hat{\mathbf{C}}^{(l)}\mathbf{R}\hat{\mathbf{C}}^{(l)H}. \quad (9)$$

To recover the desired user signal, define the decorrelating precoding filter

$$\mathbf{G} = \hat{\mathbf{R}}^{-1}. \quad (10)$$

Then equation (8) can be simplified as

$$\mathbf{y} = \mathbf{S_c}^{-1}\mathbf{S_f}\mathbf{b} + \mathbf{n}. \quad (11)$$

It is obvious that the MF output for each user, $y_i = (S_{fi}/S_{ci})b_i + n_i$, does not contain MAI.

Now we determine the values of the transmit power scaling factors. The total average energy of the transmit signal during one symbol interval should satisfy $E_{\mathbf{b}}\left\{\sum_{l=1}^{L}\int_0^T \left|x^{(l)}(t)\right|^2 dt\right\} = E_{\mathbf{b}}\{\mathbf{b}^T\mathbf{b}\}$, where $E_{\mathbf{b}}\{\cdot\}$ is the expected value with respect to the data symbols. From equation (4), $E_{\mathbf{b}}\left\{\sum_{l=1}^{L}\int_0^T \left|x^{(l)}(t)\right|^2 dt\right\} = E_{\mathbf{b}}\{\mathbf{b}^T\mathbf{G}^H\mathbf{S_f}^2\mathbf{b}\}$; hence, it is sufficient to require user i to satisfy $G_{ii}S_{fi}^2 = 1$, $i = 1, 2, \ldots, K$. From (9) and (10), the diagonal elements of $\mathbf{G}$ are real. Therefore, the power scaling factor for user i is

$$S_{fi}^2 = \frac{1}{G_{ii}} = \frac{1}{[\hat{\mathbf{R}}^{-1}]_{ii}}. \quad (12)$$

Consequently, for user i, the bit error rate (BER) of the PreRAKELDP with individual power scaling in terms of the average transmit signal-to-noise ratio (SNR) γ_{bi} is

$$Pe_i(\gamma_{bi}) = Q\left(\sqrt{\frac{2\sum_{n=0}^{N-1}\sum_{l=1}^{L}\left|c_{i,n}^{(l)}\right|^2}{[\hat{\mathbf{R}}^{-1}]_{ii}}\gamma_{bi}}\right), \quad (13)$$

where $\gamma_{bi} = E_{bi}/N_0 = A_i^2/(2N_0)$. Comparing equation (13) with the BER formula of RDD [10], we find that in the absence of antenna diversity, i.e., when L=1, the performance of the PreRAKELDP with power scaling is identical to that of RDD. The Pre-RDD [4] is equivalent to the precoder derived in this section for a single antenna, but a global power scaling factor instead of individual power scaling factors is utilized. As a result, all users have the same BER given by the average of all users' BERs of the PreRAKELDP method. In practice, the PreRAKELDP is more flexible because it allows different users to satisfy their individual reliability criteria.

B. MMSE Solution with Total Power Constraint

From (13), we observe that the power scaling factors degrade precoding performance. With the same transmitter structure as in Fig.1, in this section our goal is to design the optimum linear precoding filter $\mathbf{G}$ in the sense of MMSE criterion under the total average transmit power constraint. Similarly to equation (8), with linear precoding, the sampled MF output vector at the receiver is given by

$$\mathbf{y} = \sum_{l=1}^{L} \mathbf{C}^{(l)}\hat{\mathbf{R}}\hat{\mathbf{C}}^{(l)H}\mathbf{G}\mathbf{b} + \mathbf{n}. \tag{14}$$

In equation (14), the filter $\mathbf{G}$ should be the solution of

$$\min_{\mathbf{G}\in C^{K\times K}} \quad E_{\mathbf{b}}\{||\mathbf{b}\text{-}\mathbf{S_c y}||^2\}$$

subject to

$$E_{\mathbf{b}}\left\{\sum_{l=1}^{L}\int_0^T \left|x^{(l)}(t)\right|^2 dt\right\} = E_{\mathbf{b}}\{\mathbf{b}^T\mathbf{b}\}\,. \tag{15}$$

For the convenience of derivation, define the amplitude matrix of the K users as $\mathbf{A_m} = diag\{A_1, A_2, \ldots, A_K,\}$. It can be easily shown that $E_{\mathbf{b}}\left\{\sum_{l=1}^{L}\int_0^T \left|x^{(l)}(t)\right|^2 dt\right\} = tr\left\{\mathbf{G}^H\hat{\mathbf{R}}\mathbf{G}\mathbf{A_m}^2\right\}$. Using the Lagrange multiplier method, it is desired to find the solution $\mathbf{G}$ that minimizes

$$\varepsilon \equiv E_{\mathbf{b}}\{||\mathbf{b}\text{-}\mathbf{S_c y}||^2\} + \lambda\, tr\{\mathbf{G}^H\hat{\mathbf{R}}\mathbf{G}\mathbf{A_m}^2\}. \tag{16}$$

It is shown in [19] that the solution to (15) is

$$\mathbf{G} = (\hat{\mathbf{R}} + \lambda\mathbf{I})^{-1}, \tag{17}$$

where the value of λ can be determined from the constraint (15), which is equivalent to $\mathrm{tr}\{\mathbf{G}^H\hat{\mathbf{R}}\mathbf{G}\mathbf{A_m}^2\} = \mathrm{tr}\{\mathbf{A_m}^2\}$. Note that $\hat{\mathbf{R}}$ is positive definite. Suppose its eigenvalues are $a_1, a_2, \ldots, a_K$. Define the unitary matrix $\mathbf{F}=\{F_{ij}\}_{K\times K}$ with jth column given by the eigenvector of $\hat{\mathbf{R}}$ corresponding to a_j, $j = 1, 2, \ldots, K$. Therefore, $\hat{\mathbf{R}} = \mathbf{F}diag\{a_j\}\mathbf{F}^H$ and $\mathbf{G} = \mathbf{F}diag\{(a_j+\lambda)^{-1}\}\mathbf{F}^H$, where $diag\{a_j\}$ represents the $K\times K$ diagonal matrix with a_j at the jth diagonal position. The total power constraint can be expressed as

$$\sum_{i=1}^{K}\left(A_i^2\sum_{j=1}^{K}\frac{|F_{ij}|^2 a_j}{(\lambda+a_j)^2}\right) = \sum_{i=1}^{K} A_i^2. \tag{18}$$

By solving this equation we can obtain the value of λ.

The total power constraint optimization for single-path AWGN channel was studied in [2] and [12]. It was shown that for the single-path channel with highly-correlated signature sequences, the total power constraint optimization does not have the BER performance advantage over the transmit power scaling method, due to the severe residual MAI in the received signals. However, we consider a different channel model in this paper corresponding to the orthogonal downlink CDMA over multipath fading channels. For this model, the residual MAI is very small for the total power constraint approach. Therefore, the total power constraint optimization results in better performance than the individual power scaling for the PreRAKELDP. This conclusion will be verified by simulation results. On the other hand, optimization under the total power constraint has much higher complexity than the transmit power scaling.

It should be noted that to achieve a simple and practical transmitter design for a system with large number of users, the allocation of individual user transmit powers is not taken into account in the PreRAKELDP designs. Since the pre-RAKE combiner is equivalent to a matched filter which is matched to the multipath fading channels, the PreRAKELDP with individual user power scaling and that with total transmit power constraint are the optimum linear ZF and MMSE precoders, respectively.

IV. MDP WITH MULTIPLE TX ANTENNAS

In the PreRAKELDP, the calculation of precoding filter coefficients requires the operation of matrix inversion of a CSI-dependent matrix, which results in high computational complexity. To simplify the transmitter, we present an alternative precoding algorithm, MDP. As in the PreRAKELDP, in the MDP design each MS only needs to use a simple and CSI-independent matched filter for data detection.

As shown in Fig. 2, for a system with L transmit antennas, the precoding process consists of L parallel and independent branches. Fig. 3 demonstrates the detailed structure of each branch. In the lth branch, the data symbols for the K users are first weighed by the normalized channel gains, with the output given by the ($K\times N$)-element vector $\hat{\mathbf{C}}^{(l)H}\mathbf{b}$. As shown in Fig. 2, thechannel gain weighed signals are filtered by the power control filter and decorrelating filter. The decorrelating filter is defined as $\mathbf{G} = \mathbf{R}^{-1}$. The power control filter for the

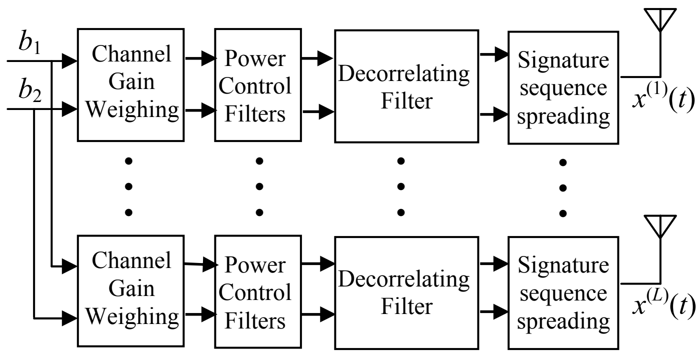

Fig. 2 Transmitter Diagram of MDP for a 2-user, *L*-antenna, N-channel paths/user system

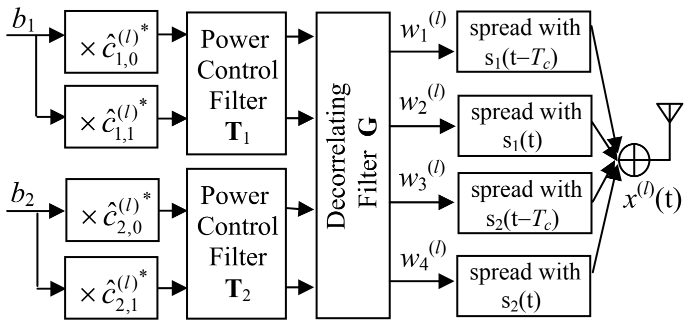

Fig. 3 The Structure of the *l*th branch in the MDP transmitter (a 2-user, 2-path/user system)

ith user is defined as $\mathbf{T}_i = \widetilde{S}_i([\mathbf{G}]_{ii})^{-1}$ ($[\mathbf{G}]_{ij}$ represents the (i, j)th $N\times N$ block of $\mathbf{G}$); for all K users, define the compact form as $\mathbf{T}\equiv diag\{\mathbf{T}_1,\mathbf{T}_2,\ldots,\mathbf{T}_K\}$, which is a ($K\times N$)-row ($K\times N$)-column matrix with $\mathbf{T}_i$ as the ith diagonal $N\times N$ block, $i = 1, 2, \ldots, K$. Note that the notation for the power control filters and decorrelating filter are not identified with an antenna index, which means these filters have exactly the same definitions for all the L branches. The output of the decorrelating filter for the lth branch is given by $\mathbf{w}^{(l)} = [w_1^{(l)}, w_2^{(l)},\ldots,w_{K\times N}^{(l)}]^T = \mathbf{G}\mathbf{T}\hat{\mathbf{C}}^{(l)H}\mathbf{b}$. Following the signature sequence spreading, the transmitted signal from the lth antenna is

$$x^{(l)}(t) = \sum_{k=1}^{K}\sum_{n=0}^{N-1} w_{k\times N-n}^{(l)} s_k(t-nT_c). \qquad (19)$$

It is sufficient to require the individual users to satisfy $\sum_{l=1}^{L}\hat{\mathbf{c}}_i^{(l)}\mathbf{T}_i^{T}[\mathbf{G}]_{ii}^{T}\mathbf{T}_i\hat{\mathbf{c}}_i^{(l)H} = 1, \quad i = 1, 2, \ldots, K$. It can be calculated that $\widetilde{S}_i$ is

$$\widetilde{S}_i = \left(\sum_{l=1}^{L}\hat{\mathbf{c}}_i^{(l)}([\mathbf{G}]_{ii})^{-1}\hat{\mathbf{c}}_i^{(l)H}\right)^{-1/2}. \qquad (20)$$

At the ith user's receiver, the baseband-equivalent received signal is given by

$$r_i(t) = \sum_{p=0}^{N-1}\sum_{l=1}^{L} c_{i,p}^{(l)} x^{(l)}(t-pT_c) + n_i(t). \qquad (21)$$

With the same MF as in the PreRAKELDP, the MF result for the ith user is

$$y_i = \sum_{k=1}^{K}\sum_{n=0}^{N-1}\sum_{p=0}^{N-1}\sum_{l=1}^{L} c_{i,p}^{(l)} w_{k\times N-n} R_{i,k}^{N-1-(n+p)} + n_i, \qquad (22)$$

where n_i is the filtered noise component with power $E\{n_i n_i^*\} = N_0$. For all K users, the compact form of the MF output is given by

$$\mathbf{y} = \mathbf{S_c}^{-1}\left(\sum_{l=1}^{L}\hat{\mathbf{C}}^{(l)}\mathbf{T}\hat{\mathbf{C}}^{(l)H}\right)\mathbf{b} + \mathbf{n}. \qquad (23)$$

Therefore, the simplified result of y_i is

$$y_i = \left(\sum_{l=1}^{L}\mathbf{c}_i^{(l)}([\mathbf{R}^{-1}]_{ii})^{-1}\mathbf{c}_i^{(l)H}\right)b_i + n_i. \qquad (24)$$

Note that the MAI is completely cancelled by transmitter precoding. The BER in terms of γ_{bi} is given by

$$Pe_i(\gamma_{bi}) = Q\left(\sqrt{2\sum_{l=1}^{L}\mathbf{c}_i^{(l)}([\mathbf{R}^{-1}]_{ii})^{-1}\mathbf{c}_i^{(l)H}\gamma_{bi}}\right). \qquad (25)$$

The BER of single-antenna MDP is identical to that of the Rx-based MDD [11, 15]. The MDD is a suboptimal linear decorrelating MUD. It is shown in [10] that the RDD outperforms the MDD. Since the PreRAKELDP with individual power scaling has identical performance to the RDD, it has better performance than the MDD and MDP. However, as the precoding process in MDP does not involve the inverse operation of CSI-related matrix, the MDP is computationally simpler than the PreRAKELDP and the RDD.

The STPR MUP in [7] employs the same decorrelating filter $\mathbf{R}^{-1}$ as the MDP does; thus it has similar complexity to the MDP. The analysis in [19] indicates that the average performance of MDP is better than that of the STPR MUP. More accurate comparison among linear precoders is given by the simulation experiments in the next section.

V. NUMERICAL RESULTS

In this section, we observe the performance of linear precoding techniques by a few numerical experiments. For all simulation examples, BPSK modulation, orthogonal signature sequence spreading and equal transmit power for all users are assumed. There are three resolvable channel paths from each Tx antenna at BS to each user's Rx

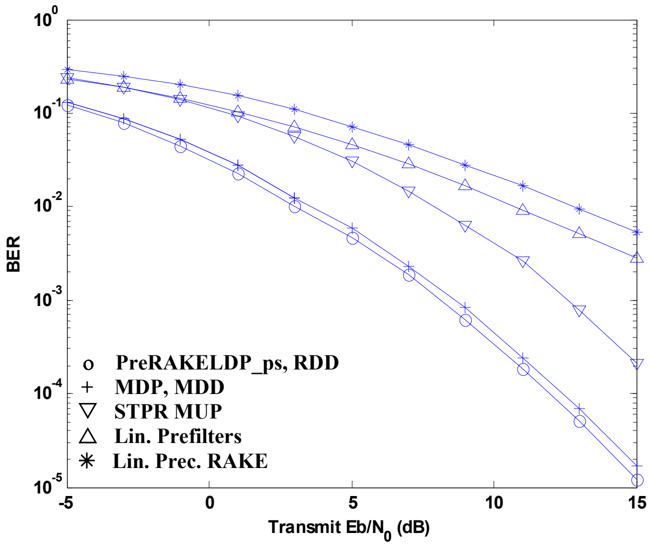

Fig. 4 Performance comparison of linear precoding techniques in multipath fading channels, 4 users with equal Tx powers, 3 channel-paths/user, single Tx antenna.

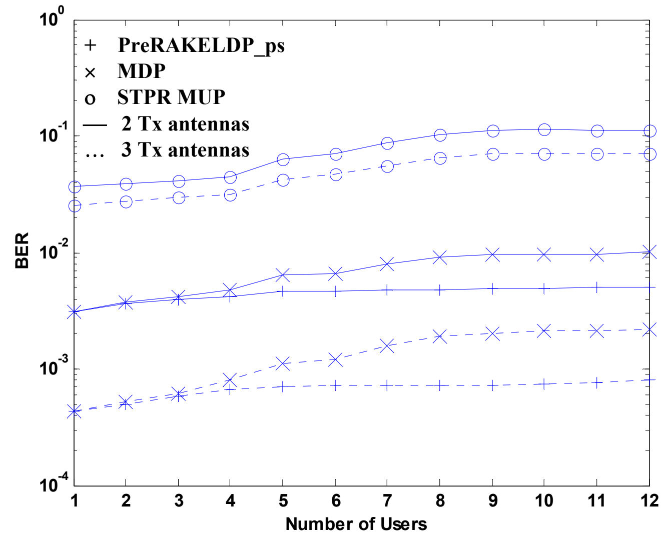

Fig. 6 Performance comparison of three space-time precoding methods, 1 to 12 users with equal Tx powers, 3 channel-paths/user, Tx Eb/N$_0$ = 0dB.

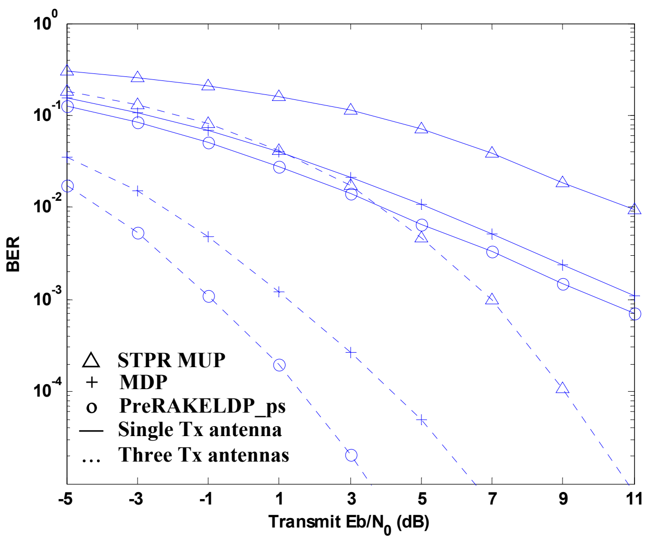

Fig. 5 Performance comparison of three space-time precoding methods, 8 users with equal Tx powers, 3 channel-paths/user.

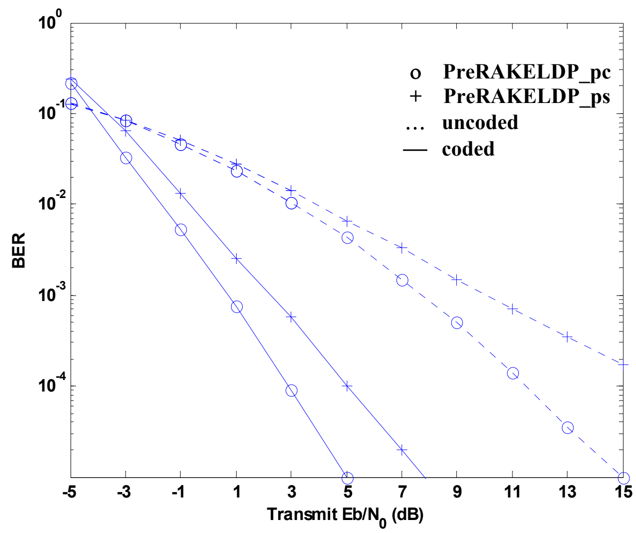

Fig. 7 Performance comparison of individual power scaling and total power constraint for the PreRAKELDP, 8 users with equal Tx powers, 3 channel-paths/user, single antenna.

antenna. In the first example, consider a 4-user uncoded system with single transmit antenna. Fig.4 shows the BER averaged over all users for the ZF-based approaches: PreRAKELDP with individual power scaling (labeled as"PreRAKELDP_ps"), MDP, linear precoding with RAKE receiver (labeled as "Prec. with RAKE") [2], linear decorrelating prefilters (labeled as "Prefilters") [3], STPR MUP [7], Rx-based RDD [10] and MDD [11]. As analyzed before, the proposed PreRAKELDP_ps and MDP have identical BER to RDD and MDD, respectively. Observe that the two new precoders significantly outperform other methods. The MDP has modestly higher BER than the PreRAKELDP_ps.

Next, we compare the performance of the three ZF precoders, PreRAKELDP_ps, MDP and STPR MUP, with multiple Tx antennas. An 8-user uncoded system is considered. Fig. 5 shows the average BER of these methods with one and three transmit antennas, respectively. Obviously, the antenna diversity results in significant performance improvement. Fig. 6 shows the

BER averaged over all users versus the number of users. The transmit SNR is fixed at 0dB. Observe that the BER saturates as the number of users becomes greater than eight. This observation is consistent with the result in [1, p255] that for decorrelating (ZF) MAI cancellation methods, the BER converges as the number of users $K \rightarrow \infty$.

In Fig. 7, we compare the performance of the PreRAKELDP with individual power scaling (PreRAKELDP_ps) and with total power constraint (PreRAKELDP_pc). An 8-user single-transmit antenna system is considered. Both coded and uncoded cases are simulated. For the coded system, we employ the rate ½ convolutional code with generator vectors 753 and 561, based on WCDMA standard [20]. At the receiver end, the soft decision decoding is implemented by the standard Viterbi decoder at the MF output. It is shown that for both coded and uncoded cases, the PreRAKELDP_pc has lower BER than PreRAKELDP_ps.

VI. CONCLUSIONS

In this paper, we proposed two Tx-based linear MAI cancellation techniques with multiple transmitter antennas, the PreRAKELDP and the MDP, for the downlink of DS-CDMA systems. Using theoretical and numerical analysis, it is shown that these novel schemes outperform previously investigated linear transmitter precoding methods. The optimal PreRAKELDP has better performance than MDP, at the cost of greater complexity. Two different transmit power control strategies are explored for the PreRAKELDP. It was demonstrated that the MMSE-based method with total transmit power constraint has lower error rate than computationally simpler ZF-based precoder with power scaling.

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