Nonlinear Multiuser Precoding for Downlink DS-CDMA Systems over Multipath Fading Channels

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Abstract — **We propose a Transmitter (Tx)-based nonlinear decorrelating interference cancellation method, Tomlinson-Harashima precoding (THP), to combat multiple access interference (MAI) for the downlink of Direct Sequence Code Division Multiple Access (DS-CDMA) systems in multipath fading channels. Since diversity combining and MAI cancellation are employed in the transmitter, the mobile user receivers remain as simple as in single-path single user channels. Two THP designs, PreRakeTHP and Multipath Decorrelating THP (MDTHP), are derived and compared with previously investigated decorrelating techniques. It is shown that the proposed methods improve upon linear precoding and detection techniques and on the decision-feedback detectors for multipath CDMA channels, while retaining low complexity. Moreover, the MDTHP precoder is attractive for rapidly varying mobile radio channels since its filters do not need to be updated as fading coefficients vary.**

Key Words — **multiple access interference cancellation, transmitter precoding, multiuser detection.**

I. Introduction

MAI presents a major limitation on the performance of DS-CDMA systems. In recent years, the requirement of small-size low-power mobile user receivers for the downlink CDMA channels motivated the development of Tx-based MAI cancellation techniques, termed multiuser transmitter precoding. Similarly to the receiver (Rx)-based decorrelating multiuser detection (MUD) [1], decorrelating precoding techniques based on the zero forcing (ZF) criterion are simple and efficient [2, 3]. These precoding methods also satisfy the minimum mean square error (MMSE) criterion [2, 3]. The linear decorrelating precoding [2] can completely pre-cancel MAI, but its performance is inherently degraded by transmit power scaling.

To improve performance but retain low complexity, we develop two nonlinear precoding techniques based on the THP principle [5]. The basic idea of THP for Multiple Input Multiple Output (MIMO) additive white Gaussian noise (AWGN) channels was developed in [6]. For multipath fading CDMA channels, we incorporate diversity combining techniques into the THP transmitter. Thus, the mobile station requires only a single path single user receiver. Two different designs, PreRakeTHP and MDTHP, are proposed and provide desired performance-complexity trade-off in precoder implementation.

As discussed in [6,13-15], many techniques used in transmitter for the downlink of CDMA systems have analogous structure and similar performance to those employed in the receiver for the uplink. In this paper, we address this duality and show that THP methods outperform previously proposed Tx- and Rx-based linear and decision-feedback methods.

In the next section, we introduce the mathematical model for the downlink of the multipath fading CDMA channel and describe the PreRakeTHP and MDTHP methods. The numerical analysis and final conclusions are presented in sections III and IV, respectively.

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II. Transmitter Precoding With Multipath Diversity

Consider the downlink channel of a K-user DS-CDMA system. Suppose the transmitted signals are subject to frequency selective slow fading with N resolvable multipath components for every user. By assuming that the multipath spread is small relative to the symbol duration, the inter-symbol interference (ISI) is minor and ignored in the following discussions. For user i, $c_{i,n} = \alpha_{i,n}e^{j\varphi_{i,n}}$ represents the gain of the nth path component, $\forall i = 1,2,\ldots,K$ and $n = 0,1,\ldots, N\text{-}1$. The received equivalent baseband signal at the ith mobile user receiver site can be expressed by $r_i(t) = \sum_{k=1}^{K}\sum_{n=0}^{N-1} c_{i,n}b_k s_k(t-nT_c) + n_i(t)$, where b_k is the data symbol for the kth user in the symbol interval $[0,T)$, $s_k(t)$ is the signature sequence for the kth user, T_c is the chip duration, and $n_i(t)$ is complex white Gaussian noise with zero mean and variance N_0. Note that the data symbols can be either PAM or QAM modulated, but we only consider M-PAM data as an example in the following discussion, and QAM modulation example is given in simulation results. For user i, denote the minimum Euclidean distance for a data symbol as $2A_i$, i.e., $b_i \in \{-(M-1)A_i, -(M-3)A_i, \ldots, (M-1)A_i\}$. Define the channel gain vector for the ith user $\mathbf{c}_i=[c_{i,0},c_{i,1},\ldots,c_{i,N-1}]$, the column vector of data

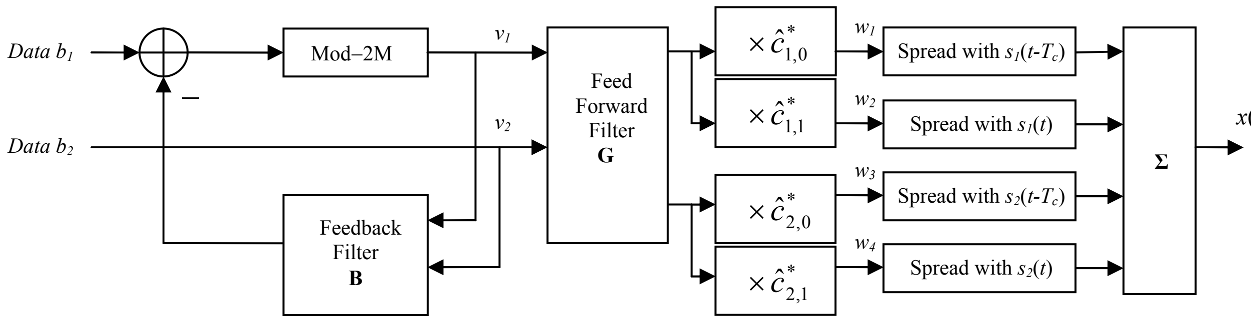

Fig. 1 PreRakeTHP Transmitter for Multipath Channels (a 2-user 2-channel path/user system)

symbols for all K users $\mathbf{b}=[b_1,b_2,\ldots,b_K]^T$, and the diagonal matrix $\mathbf{A}=diag\{A_1,A_2,\ldots,A_K\}$.

In this paper, we assume that the transmitter has the perfect knowledge of channel coefficients. In practice, this information can be obtained via feedback channels, and requires long range prediction for rapidly varying fading channels [11].

Fig.1 shows the transmitter diagram of the proposed ***PreRakeTHP*** scheme. In this method, pre-RAKE combining is employed at the transmitter to achieve frequency diversity [7]. We use the THP precoder to cancel multipath-induced MAI prior to feeding the input signals to the pre-RAKE filters. The output of the feedback (FB) filter is fed to a bank of mod-2M operators to limit the transmit power. For user i, given an arbitrary real input β, the output of the mod-2M operator $\tilde{\beta}$ satisfies $\tilde{\beta}/A_i=\beta/A_i+2Md_i$, where d_i is the integer to render $\tilde{\beta}/A_i$ within $(-M, M]$. (For complex input data, mod-2M operation is applied to the real and imaginary parts of the input.) The output vector of the mod-2M operator bank $\mathbf{v}=[v_1, v_2, \ldots, v_K]^T$ satisfies

$$\mathbf{v} = \mathbf{b} - \mathbf{B}\mathbf{v} + 2M\mathbf{A}\mathbf{d}, \qquad (1)$$

where $\mathbf{d}\equiv[d_1,d_2,\ldots,d_K]^T$. Equivalently, $\mathbf{v}=(\mathbf{B}+\mathbf{I})^{-1}(\mathbf{b}+2M\mathbf{A}\mathbf{d})$. The FB filter matrix $\mathbf{B}=\{B_{ij}\}_{K\times K}$ is upper triangular with all-zero diagonal. Therefore, for the last user, $v_K=b_K$ and $d_K=0$ (mod-2M operation is not needed for the last user); for $i=K-1, K-2, \ldots, 1$, (1) becomes $v_i=b_i-\sum_{j=i+1}^{K}B_{ij}v_j+2MA_id_i$. The outputs of the FB filter are fed to the feed forward (FF) filter $\mathbf{G}$, which is a $K\times K$ matrix. The outputs of $\mathbf{G}$ undergo pre-RAKE weighing, and the resulting vector $\mathbf{w}$ has $K\times N$ elements.

For the ith user, denote the normalized values of channel gains for pre-RAKE combining as $\hat{c}_{i,n}\equiv S_{fi}\,c_{i,n}$, and the normalized channel gain vector $\hat{\mathbf{c}}_i=[\hat{c}_{i,1},\hat{c}_{i,2},\ldots,\hat{c}_{i,K}]$, where the normalization factor is $S_{fi}=\left(\sum_{n=0}^{N-1}|c_{i,n}|^2\right)^{-1/2}$. For all K users, define a K-row $(K\times N)$-column channel gain matrix $\mathbf{C}=diag\{\mathbf{c}_1,\mathbf{c}_2,\ldots,\mathbf{c}_K\}$ and the corresponding normalized channel matrix $\hat{\mathbf{C}}=diag\{\hat{\mathbf{c}}_1,\hat{\mathbf{c}}_2,\ldots,\hat{\mathbf{c}}_K\}$. If we express the normalization factors for K users in terms of a diagonal matrix $\mathbf{S_f}=diag\{S_{f1},S_{f2},\ldots,S_{fK}\}$, then $\hat{\mathbf{C}}=\mathbf{S_f}\mathbf{C}$. With these definitions, $\mathbf{w}=\hat{\mathbf{C}}^H\mathbf{G}\mathbf{v}$. Following spreading with the signature sequences, the transmitted signal is given by

$$x(t) = \sum_{k=1}^{K}\sum_{l=0}^{N-1} w_{kN-l}\,s_k(t-lT_c). \qquad (2)$$

The equivalent baseband received signal for the ith user is

$$r_i(t) = \sum_{k=1}^{K}\sum_{n=0}^{N-1}\sum_{l=0}^{N-1} c_{i,n}w_{kN-l}\,s_k(t-lT_c-nT_c) + n_i(t). \qquad (3)$$

The outputs of matched filters sampled at $t=(N-1)T_c$ are

$$y_i = \int_{(N-1)T_c}^{T+(N-1)T_c} r_i(t)s_i(t-(N-1)T_c)dt. \qquad (4)$$

Represent the cross-correlations between the delayed signature sequences as

$$R^m_{i,k}=\int_0^T s_i(t)s_k(t+mT_c)dt,\ m\in\{-(N-1),-(N-2),\ldots,(N-1)\} \qquad (5)$$

Substituting (3) and (5) into (4), we obtain

$$y_i = \sum_{k=1}^{K}\sum_{n=0}^{N-1}\sum_{m=-(N-1)}^{N-1} c_{i,n}R^m_{i,k}w_{(k-1)N+m+n+1} + n_i, \qquad (6)$$

where n_i is the filtered noise component with power $E\{n_in_i^*\}=N_0$.

For any $i,k\in\{1,2,\ldots K\}$, define the $N\times N$ correlation matrix $\mathbf{R}_{i,k}$ with the jth row $[R^{1-j}_{i,k}, R^{2-j}_{i,k}, \ldots, R^{N-j}_{i,k}]$, $j=1,2,\ldots,N$. Then construct a $(K\times N)\times(K\times N)$ matrix $\mathbf{R}$ by superimposing $K\times K$ non-overlapping $N\times N$ submatrices $\mathbf{R}_{i,k}$, $i,k\in\{1,2,\ldots,K\}$. Then (6) reduces to

$$\mathbf{y} = \mathbf{C}\mathbf{R}\mathbf{w} + \mathbf{n} = \mathbf{S_f}^{-1}\hat{\mathbf{C}}\mathbf{R}\hat{\mathbf{C}}^H\mathbf{G}(\mathbf{B}+\mathbf{I})^{-1}(\mathbf{b}+2M\mathbf{A}\mathbf{d}) + \mathbf{n}, \quad (7)$$

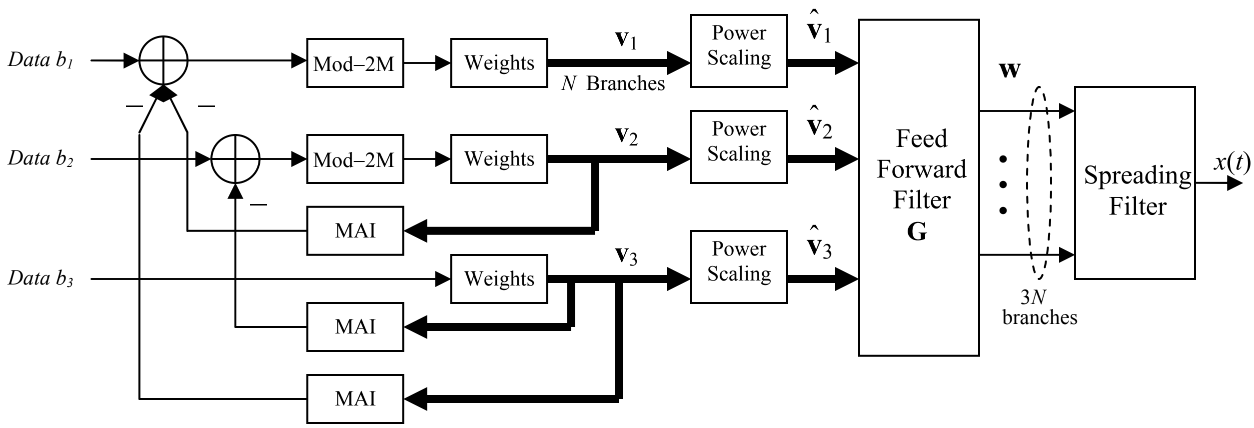

Fig. 2 MDTHP Transmitter for Multipath Channels (a 3-user *N*-channel path/user system)

where $\mathbf{y}=[y_1, y_2, \ldots, y_K]^T$ and $\mathbf{n}=[n_1, n_2, \ldots, n_K]^T$.

Define $\mathbf{R}_p \equiv \mathbf{C}\mathbf{R}\mathbf{C}^H$ and $\hat{\mathbf{R}}_p \equiv \mathbf{S}_f^2\mathbf{R}_p = \hat{\mathbf{C}}\mathbf{R}\hat{\mathbf{C}}^H$. The matrix $\mathbf{R}_p$ is Hermitian symmetric with real diagonal elements, and in practice $\mathbf{R}_p$ is usually positive definite. Consequently, it can be factored as $\mathbf{F}_p^H\mathbf{F}_p$ using the Cholesky factorization, where $\mathbf{F}_p=\{f_{ij}\}_{K\times K}$ is complex lower triangular matrix. The corresponding normalized matrix is $\hat{\mathbf{F}}=\{\hat{f}_{ij}\}_{K\times K}=\mathbf{S}_f\mathbf{F}_p$. To cancel MAI in (7), the filters $\mathbf{B}$ and $\mathbf{G}$ are defined as

$$\mathbf{B} \equiv diag(\hat{\mathbf{F}})^{-1}\times\hat{\mathbf{F}}^H-\mathbf{I}, \tag{8}$$

$$\mathbf{G} \equiv \hat{\mathbf{F}}^{-1}. \tag{9}$$

Using (8) and (9), equation (7) can be simplified as

$$\mathbf{y} = \mathbf{S}_f^{-1} diag(\hat{\mathbf{F}})(\mathbf{b}+2M\mathbf{A}\mathbf{d}) + \mathbf{n}. \tag{10}$$

Equivalently, for the ith user, $\forall\, i = 1, 2, \ldots, K$,

$$y_i = f_{ii}(b_i+2MA_i d_i) + n_i. \tag{11}$$

After scaling by f_{ii}^{-1} and mod-2M reduction at the receiver, the input to the decision device is given by

$$f_{ii}^{-1}y_i = b_i + f_{ii}^{-1}n_i, \tag{12}$$

where the noise component has power $f_{ii}^{-2}N_0$. For the M-PAM system, the average transmit signal-to-noise ratio (SNR) per bit is [12]

$$\gamma_{bi} \equiv \frac{E_{bi}}{N_0} = \frac{(M^2-1)A_i^2}{6N_0\log_2 M}. \tag{13}$$

The ideal instantaneous Symbol Error Rate (SER) for the ith user is obtained from equation (12) as

$$Pe_i\,(\gamma_{bi}) = \frac{2(M\text{-}1)}{M} Q\left(\sqrt{\frac{6(\log_2 M)f_{ii}^2}{M^2\text{-}1}\,\gamma_{bi}}\right). \tag{14}$$

The PreRakeTHP described above is related to several precoders and MUDs proposed previously. In general, it represents the Tx precoding version of the DF MUD [4] and is based on MIMO THP schemes [6]. On the other hand, it can be viewed as a non-linear (DF) Tx precoding implementation of the optimal Rake Decorrelating Detector (RDD) [10], a linear MUD that employs the RAKE combiner followed by the decorrelating matrix (or, equivalently, of the pre-RDD precoder [13, 14]).

To compare the PreRakeTHP with the RDD, first we note that the correlation matrix $\mathbf{R}$ in [10] corresponds to $\mathbf{R}_p^H$ above. Thus, $\mathbf{R}^{-1}=\mathbf{R}_p^{-H}=\mathbf{F}_p^{-H}\mathbf{F}_p^{-1}$, and $(\mathbf{R}^{-1})_{ii} > [(\mathbf{F}_p^{-1})_{ii}]^2 = f_{ii}^{-2}$. Comparing the error rate formulas for PreRakeTHP (equation (14)) and RDD (equation (7) in [10]), we observe that PreRakeTHP outperforms the RDD (the optimal decorrelating linear MUD for frequency selective channels) for all users $i>1$. For the first user, these methods have the same SER. Similarly, it can be demonstrated that the PreRakeTHP has better performance than the pre-RDD proposed in [13]. Since it has been proved in [10] that the RDD outperforms the Multipath Decorrelating Dectector (MDD) [9], PreRakeTHP also outperforms MDD.

One drawback of the PreRakeTHP is that the coefficients of the FB and FF filters depend on the channel gains. Consequently, the matrix factorization and inversion required for the computation of these coefficients have to be performed frequently, especially for rapidly varying fading channels. We present an alternative THP design termed the ***Multipath Decorrelating Tomlinson-Harashima Precoding (MDTHP)*** that alleviates this problem. The transmitter structure of the MDTHP is shown in Fig.2.

First, note that the matrix $\mathbf{R}$ above (see (5-7)) is determined by the spreading sequences and is not related to channel gains. Since $\mathbf{R}$ is symmetric and positive definite in practice, we can decompose $\mathbf{R} = \mathbf{F}^T\mathbf{F}$ by the Cholesky factorization, where $\mathbf{F}$ is a lower triangular $KN\times KN$ matrix. We divide $\mathbf{F}$ into $K\times K$ non-overlapping $N\times N$ submatrices $[\mathbf{F}]_{ij}$, $i,j\in\{1,2,\ldots,K\}$,where $[\mathbf{F}]_{i,i}$ are lower triangular. The feedback loop in the transmitter is implemented as follows. First, the $N\times 1$ vector $\mathbf{v}_K$ is computed by applying N weights for the last user, $\mathbf{v}_K = b_K \cdot ([\mathbf{F}]_{KK}\hat{\mathbf{c}}_k^H)$. The interference caused by user K is calculated from $\mathbf{v}_K$, and fed back to be canceled from the signals of other users. This procedure is repeated consecutively for $k = K-1, K-2, \ldots, 2$, thus forming vectors $\mathbf{v}_k$. For user i, $\forall i=1,2,\ldots,K-1$, the feedback from user j, j=i+1,…,K, is calculated by $\frac{\hat{\mathbf{c}}_i[\mathbf{F}]_{ji}{}^T\mathbf{v}_j}{\sqrt{\beta_i\beta_j}}$, where $\beta_i \equiv \hat{\mathbf{c}}_i[\mathbf{F}]_{ii}{}^T[\mathbf{F}]_{ii}\hat{\mathbf{c}}_i{}^H$. Thus, for users 1 through K−1, the output of the feedback loop is

$$\mathbf{v}_i = [\mathbf{F}]_{ii}\hat{\mathbf{c}}_i{}^H\left(b_i - \sum_{j=i+1}^{K}\frac{\hat{\mathbf{c}}_i[\mathbf{F}]_{ji}{}^T\mathbf{v}_j}{\sqrt{\beta_i\beta_j}} + 2MA_id_i\right),\ i=1,2,\ldots,K\text{-}1. \quad (15)$$

The output power is normalized by multiplying $\mathbf{v}_i$ by the scaling factor

$$S_{vi} = \left(\hat{\mathbf{c}}_i[\mathbf{F}]_{ii}{}^T[\mathbf{F}]_{ii}\hat{\mathbf{c}}_i{}^H\right)^{1/2} = \beta_i^{-1/2}. \quad (16)$$

Let $\hat{\mathbf{v}}_i = S_{vi}\cdot\mathbf{v}_i$, i = 1,2,…,K. Represent the input of the FF filter by the vector $\hat{\mathbf{v}} = [\hat{\mathbf{v}}_1{}^T,\hat{\mathbf{v}}_2{}^T,\ldots,\hat{\mathbf{v}}_K{}^T]^T$. Then its output is $\mathbf{w} = \mathbf{G}\hat{\mathbf{v}}$, where the FF filter is defined as $\mathbf{G} = \mathbf{F}^{-1}$. Following the derivation similar to that for the PreRakeTHP above, we obtain the matched filter bank output in the receivers as $\mathbf{y} = \mathbf{CRw} + \mathbf{n}$. For user i, this output is

$$y_i = (\sqrt{\beta_i}\,/S_{fi})(b_i+2MA_id_i) + n_i,\ \ i = 1,2,\ldots,K, \quad (17)$$

where $S_{fi} = \left(\sum_{n=0}^{N\text{-}1}|c_{i,n}|^2\right)^{-1/2}$. Consequently, the ideal instantaneous SER for the ith user is

$$Pe_i\,(\gamma_{bi}) = \frac{2(M\text{-}1)}{M}Q\left(\sqrt{\frac{6(\log_2 M)\ (\mathbf{c}_i[\mathbf{F}]_{ii}{}^T[\mathbf{F}]_{ii}\mathbf{c}_i{}^H)}{M^2\text{-}1}\,\gamma_{bi}}\right). (18)$$

The MDTHP method significantly simplifies Tx precoding relative to the PreRakeTHP, since it employs the factorization of the channel gain-independent matrix $\mathbf{R}$. Thus even for rapidly varying mobile radio channels the matrix factorization and inversion operations do not have to be performed frequently. The MDTHP is related to several other simplified MUD and Tx precoding methods that also employ $\mathbf{R}$. First, equation (18) is identical to the theoretical performance of Multipath Decorrelating Decision Feedback Receiver (MDDFR) (equation (16) in [8]). However, since this DF MUD is degraded by the error propagation, MDTHP has better practical performance than MDDFR. Next, we compare performance of the MDTHP with two simplified linear methods: the MDD [9] and the linear Tx precoding method equivalent to the MDD (a slightly modified version of the precoder in [15]). Compare the decision statistic of MDTHP, $\mathbf{c}_i[\mathbf{F}]_{ii}{}^T[\mathbf{F}]_{ii}\mathbf{c}_i{}^H$, with that of MDD (see equation (1) in [10]), $\mathbf{c}_i([\mathbf{R}^{-1}]_{ii})^{-1}\mathbf{c}_i{}^H$. Based on the theorem for the inverse of a partitioned symmetric matrix, it is easy to prove that $\mathbf{c}_i[\mathbf{F}]_{ii}{}^T[\mathbf{F}]_{ii}\mathbf{c}_i{}^H \geq \mathbf{c}_i([\mathbf{R}^{-1}]_{ii})^{-1}\mathbf{c}_i{}^H$ (the equality is satisfied for i=1). Therefore, for user 1 MDTHP and MDD have the same performance, and for other users MDTHP outperforms MDD (and the equivalent linear precoder).

The PreRakeTHP and MDTHP result in the same performance for the last user, which agrees with the SER of the isolated RAKE receiver for that user. We can obtain some insight into the performance comparison for other users by considering their decision statistics averaged over the channel fading. For practical spreading codes such as Walsh-Hadamard sequences, the cross-correlations usually satisfy $|R^m_{i,j}|$ << $|R^m_{i,i}|$, and $R^0_{i,j}$=0; $R^0_{i,i}$=1, $\forall$ $i\neq j$; $i,j\in\{1,2,\ldots,K\}$ and $m\in\{\text{-}(N\text{-}1),\text{-}(N\text{-}2),\ldots 0,\ldots,(N\text{-}1)\}$. With this assumption, it is easy to show that $E\{f_{ii}^2\} \approx E\{\mathbf{c}_i[\mathbf{R}]_{ii}\mathbf{c}_i{}^H\} \geq E\{\mathbf{c}_i[\mathbf{F}]_{ii}{}^T[\mathbf{F}]_{ii}\mathbf{c}_i{}^H\}$ ($E\{\cdot\}$ is the expectation over the channel gains), which indicates better performance of PreRakeTHP. This conclusion is further verified by simulations in the next section. On the other hand, MDTHP is easier to implement as discussed above. In summary, there is a performance-complexity tradeoff between the PreRakeTHP and the MDTHP.

For M-PAM/QAM systems, the practical performance of THP methods is degraded when M is small due to the power penalty and end effect of mod-2M operation [5]. Usually the first user suffers the worst degradation and the last user is not influenced at all. Therefore, to achieve more balanced performance, we sort users in the order of decreasing received powers in Section III. For larger values of M, since THP methods are not affected by error propagation, different user ordering might be desirable in practice [6].

III. Numerical Results And Analysis

Consider an 8-user, 4-channel path/user DS-CDMA system. All channel paths experience independent and identically distributed (i.i.d) Rayleigh fading, and the total average channel power is normalized to one for each user. The ideal performance of the THP methods is evaluated above. Fig.3 and Fig.4 show the SER for BPSK and 16-QAM, respectively. The SER is averaged over all users. Orthogonal Hadamard codes with length 32 chips are employed as the signature sequences. The transmit powers of all users are equal. In both figures, the THP methods significantly outperform the conventional RAKE receiver and linear decorrelating precoding [2]. For BPSK, when BER is lower than 10^{-3}, PreRakeTHP outperforms the RDD, and MDTHP is better than MDDFR and MDD, as expected from previous analysis. This confirms that the inherent advantage of nonlinear THP methods overcomes the adverse influence of the modulo operation even if M is very small. Linear precoding is seriously degraded by transmit power scaling in this case. The single user bound (SUB) is also given for reference, which is the performance of isolated single user with Rake receiver. In Fig.5, the best and poorest user performances of the three precoding methods and the optimum linear decorrelating detector RDD are further compared with 8-PAM modulation. We observe that the SER of linear precoding is higher than those of other methods, and the lowest SER is achieved by PreRakeTHP. As we have analyzed, PreRakeTHP and MDTHP result in the same SER for the last user. Due to the bit-by-bit user ordering, the last user experiences the worst instantaneous fading, and thus its average SER is the highest among all users in this example.

IV. Conclusions

Two nonlinear Tx precoding methods, PreRakeTHP and MDTHP, were proposed for CDMA systems in multipath fading channels. These schemes simplify the mobile user receiver by shifting the diversity combining and MAI cancellation to the Tx. Moreover, MDTHP is simple to implement since its filters do not depend on the rapidly varying fading coefficients. The proposed methods outperform linear decorrelating precoding and MUD methods, as well as the decision-feedback MUD for multipath fading channels.

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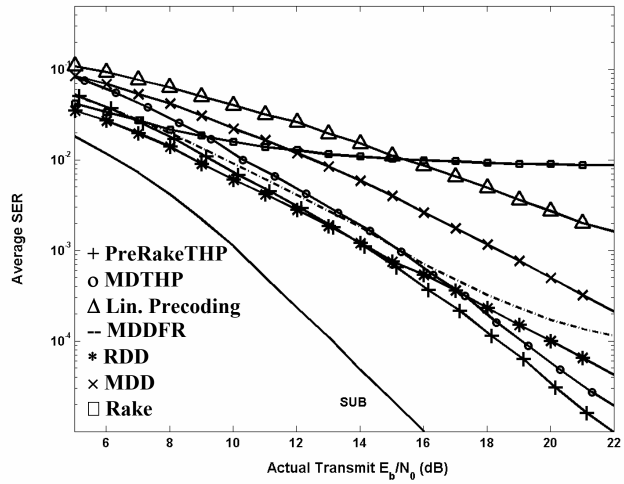

Fig. 3 Average SER comparison for 8 users, 4 channel-paths/user, BPSK

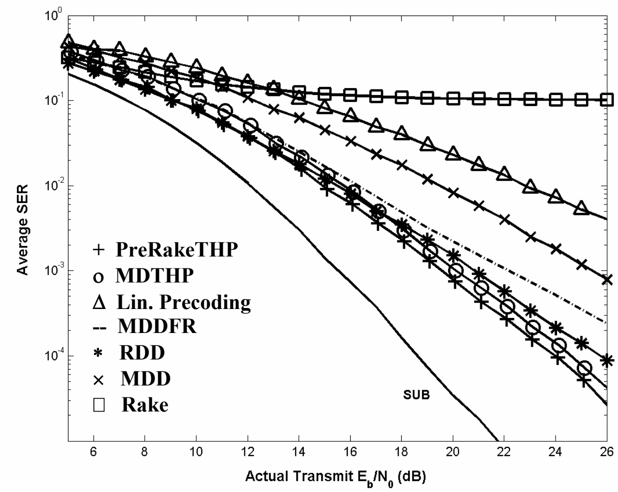

Fig. 4 Average SER comparison for 8 users, 4 channel-paths/user, 16-QAM

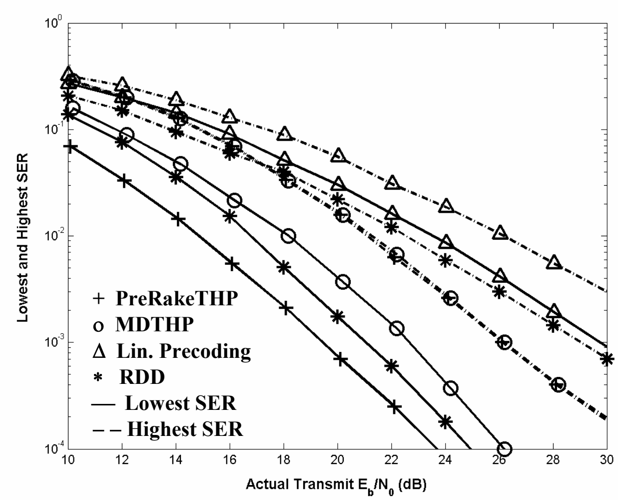

Fig. 5 Best and poorest user SER for 8 users, 4 channel-paths/user, 8-PAM