Transmitter-based Joint Nonlinear Multiuser Interference Cancellation and Multipath Diversity Combining for DS/CDMA Systems[1]

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Abstract — **A nonlinear transmitter-based multiple access interference (MAI) cancellation technique, Tomlinson-Harashima transmitter precoding (THP), is proposed for Direct Sequence (DS) Code-Division Multiple Access (CDMA) systems with or without multipath fading. This zero forcing (ZF)-based transmitter precoding method minimizes the mean square error at the receiver. It is suitable for the downlink since the receiver is as simple as the single user matched filter receiver. For frequency-selective fading channels, by combining THP with diversity techniques, we develop two specific designs, PreRAKETHP and multipath decorrelating THP (MDTHP), which provide desired performance-complexity trade-off. It is shown that THP designs outperform linear transmitter (Tx)-based methods and both linear and nonlinear receiver (Rx)-based decorrelating approaches. The nonlinear precoders have better performance for the weaker users. Therefore, they reduce total transmit power relative to linear precoders. In addition, in rapidly varying multipath fading channels, THP aided by the long range channel prediction (LRP) method achieves almost the same performance as when the channel state information is perfectly known at the transmitter.**

Index Terms — **Code-division multiple access, diversity methods, multiuser detection, Tomlinson-Harashima precoding, transmitter precoding.**

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I. INTRODUCTION

In practical wireless channels, multiple access interference is a major limitation to the performance of DS-CDMA systems. Over the past decade, various MAI rejection techniques have been developed. This research has primarily focused on Rx-based multiuser detection (MUD) [1] that results in a complex receiver while the transmitter remains simple. Thus, MUD techniques are mostly suitable for the uplink. For the downlink CDMA channel, the requirements of small-size low-power mobile station (MS) have motivated the development of Tx-based MAI pre-rejection techniques in base station (BS), termed transmitter precoding [2, 3]. The class of decorrelating precoding techniques is simple, efficient and satisfies the minimum mean square error (MMSE) criterion [2, 4], thus is most promising in practical applications. The linear decorrelating precoder proposed in [2, 5] can completely pre-cancel MAI by inserting a linear filter in the transmitter, but the performance is inherently degraded by transmit power scaling. For multipath fading channels, the RAKE receiver is employed for the method in [2]. An alternative method with simpler receiver is the decorrelating prefilters [4], which pre-cancels MAI by designing transmit waveforms rather than filtering the transmit data symbols. The simulation results in [4] show that there is little difference in performance between these two methods when user number is large. Two recently proposed linear techniques, Pre-RDD [6] and PreRAKE precoding [7, 8] simplify the receiver structure by replacing Rx-based RAKE combining with Tx-based diversity combining. Inspired by Tomlinson-Harashima equalization [9], the nonlinear method of Tomlinson Harashima Precoding (THP) proposed in [10] utilizes a feedback (FB) loop and a feed-forward (FF) filter to jointly cancel MAI. An independent work on THP [11] presented the significant duality between THP and the Decision-Feedback (DF) MUD [17]. Both [10, 11] assume that the non-orthogonal effective spreading codes on the downlink arise due to multipath fading. However, the channel models employed in these papers do not reflect the frequency-selective fading channel environment, and can only be viewed as flat fading non-orthogonal CDMA channels. Moreover, the issues and trade-offs of joint precoding and diversity combining are not addressed in these references.

In this paper, we first illustrate the principle of THP for the simple case of single-path channels with additive white Gaussian noise (AWGN) and flat Rayleigh fading channels. Then we develop two novel THP designs for frequency-selective fading channels, PreRAKETHP and Multipath Decorrelating THP (MDTHP) (see also [12]). PreRAKETHP and MDTHP incorporate Tx-based diversity combining techniques in different ways. In PreRAKETHP, the MAI cancellation is followed by the pre-RAKE combining [13]. While this precoder is the optimal THP design for multipath channels, it requires high computational complexity, since its MAI cancellation filters depend on the rapidly time variant mobile radio channel coefficients and need to be updated frequently. In MDTHP, the diversity combining is incorporated into the MAI cancellation, and the precoding filter is independent of the channel. Thus, MDTHP is simpler than PreRAKETHP, and results in moderate bit error rate (BER) loss. The essential distinction between the precoders in this paper and the methods in [10, 11] is the extension to multipath fading channels, by employing Tx-based diversity combining jointly with interference cancellation.

Several MAI cancellation 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 outperforms previously proposed Tx- and Rx-based linear and nonlinear decorrelating methods in [2, 4-8, 14-17].

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 [4]. However, by employing the spectral factorization [18], 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. For rapidly time varying fading channels, the LRP is required to enable these techniques [19]. In this paper, we investigate transmitter precoding aided by the LRP.

In addition to the CDMA systems, THP is also an efficient interference rejection approach for various multi-input/multi-output (MIMO) systems, such as orthogonal frequency division multiplexing (OFDM) and multiple-antenna channels [10]. THP is especially beneficial as an alternative to decision-feedback receivers in coded systems [9, 20].

In section II, we first present the system model for the downlink CDMA over single-path channels; then describe the principle of THP approach. In section III, the two proposed THP schemes for multipath fading channels are described. Their performance is analyzed and compared with that of related linear and nonlinear MAI rejection approaches. The numerical analysis and final conclusions are presented in sections IV and V, respectively.

II. NONLINEAR PRECODING IN SINGLE-PATH CHANNELS

In this section, we describe THP precoding for AWGN and frequency-non-selective fading downlink CDMA channels. Consider a K-user DS/CDMA system and a set of pre-assigned signature sequences $s_i(t)$, $i = 1, 2, \ldots, K$, where each sequence is restricted to a symbol interval of duration T. All the signature sequences are normalized, i.e., $\int_0^T s_i^2(t)\mathrm{d}t = 1$. In the symbol interval of interest $[0, T)$, the data symbol for user i is denoted by b_i. For high speed downlink CDMA channels, the higher-order modulation is favorable because the data transmission efficiency can be improved without increasing MAI [21, 22]. In the following discussion, we use pulse amplitude modulation (PAM) as an example in derivation, while quadrature amplitude modulation (QAM) is also considered in numerical experiments. For an M-PAM system, $b_i \in \{-(M-1)A_i, -(M-3)A_i, \ldots, (M-3)A_i, (M-1)A_i\}$, where A_i is half of the minimum Euclidean distance for the data symbols of user i. Thus the average bit energy of b_i is $E_{bi} = (M^2 - 1)A_i^2 / 6\log_2 M$ (see [20]). For the CDMA downlink over synchronous AWGN channels, the equivalent low-pass received signal at the ith user receiver is $r_i(t) = \sum_{i=1}^{K} b_i s_i(t) + n_i(t)$, where $n_i(t)$ are independent identical distributed (i.i.d.) white Gaussian noise processes with zero mean and power spectrum density N_0. In the ith user's receiver, the output of matched filter (MF) is $y_i = \int_0^T r_i(t)s_i(t)\mathrm{d}t$, i=1, 2,…, K. Define the vectors $\mathbf{b}=[b_1, b_2,\ldots, b_K]^T$, $\mathbf{s}(t)=[s_1(t), s_2(t),\ldots, s_K(t)]^T$, $\mathbf{y} = [y_1, y_2, \ldots, y_K]^T$ and the $K\times K$ correlation matrix $\mathbf{R} = \int_0^T \mathbf{s}(t)\mathbf{s}^T(t)\mathrm{d}t$. The MF output vector $\mathbf{y}$ satisfies

$$\mathbf{y} = \mathbf{R}\mathbf{b} + \mathbf{n}, \tag{1}$$

where $\mathbf{n} = [n_1, n_2, \ldots, n_K]^T$ is a zero-mean Gaussian noise vector with elements $n_i = \int_0^T n_i(t)s_i(t)\mathrm{d}t$ and the autocorrelation matrix $N_0\mathbf{I}$. ($\mathbf{I}$ is the $K\times K$ identity matrix.) Note that this model is similar to the CDMA synchronous model for the uplink, except that in the uplink model the autocorrelation matrix of the noise vector is $N_0\mathbf{R}$ [1, 8, 11]. In practice, these channels are not degraded by MAI due to the orthogonal signature sequences and synchronous transmission, and, thus, does not require Tx precoding. However, as in [2, 5], we employ this simple model with non-orthogonal users' codes to illustrate the general principle of multiuser THP precoding. Moreover, the same synchronous model arises in other MIMO systems (e.g., space-time or multicarrier transmission), and the proposed THP can be utilized in those systems (see also [10, 11] for a derivation of joint Tx/Rx THP implementation appropriate for MIMO channels with centralized receivers.)

The transmitter and receiver structures of the proposed nonlinear approach are illustrated in Fig.1. From equation (1), we observe that the multiuser interference is caused by the nondiagonal elements of the correlation matrix $\mathbf{R}$. Note that $\mathbf{R}$ is symmetric and usually positive definite, thus it can be decomposed by Cholesky factorization, $\mathbf{R} = \mathbf{F}^T\mathbf{F}$, where $\mathbf{F}$ is a lower triangular matrix. In Fig.1, the FB filter $\mathbf{B}$ and FF filter $\mathbf{G}$ are defined as

$$\mathbf{B} = diag(\mathbf{F})^{-1}\mathbf{F}^T - \mathbf{I}, \tag{2}$$

$$\mathbf{G} = \mathbf{F}^{-1}, \tag{3}$$

where $diag(\mathbf{F})^{-1} = diag(\mathbf{F}^{-1})$ is the diagonal matrix that contains the diagonal elements of $\mathbf{F}^{-1}$. Thus $\mathbf{B}$ is an upper triangular matrix with zeros along the diagonal. A bank of mod-2M operators is used to limit the transmit power. For user i, given an arbitrary real number 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 $(-M, M]$. The output vector of the mod-2M operator bank $\mathbf{v} = [v_1, v_2, \ldots, v_K]^T$ satisfies

$$\mathbf{v} = \mathbf{b} - \mathbf{B}\mathbf{v} + 2M\mathbf{A}\mathbf{d} = (\mathbf{B}+\mathbf{I})^{-1}(\mathbf{b} + 2M\mathbf{A}\mathbf{d}), \tag{4}$$

where $\mathbf{d} = [d_1, d_2, \ldots, d_K]^T$ is an integer vector. For the ith user, $i = 1, 2, \ldots, K$, d_i is chosen to guarantee v_i in the range $(-A_iM, A_iM]$. Equation (4) is equivalent to $(\mathbf{B}+\mathbf{I})\mathbf{v} = \mathbf{b} + 2M\mathbf{A}\mathbf{d}$. Since $(\mathbf{B}+\mathbf{I})$ is upper triangular with ones along the diagonal, $v_K = b_K$ and $d_K = 0$, i.e., mod-2M is not required for user K; for the ith user, $i = 1, 2, \ldots, K-1$, we can successively determine d_i and v_i based on the values of $v_{i+1}, v_{i+2}, \ldots, v_K$, i.e., $v_i = \left(b_i - \sum_{j=i+1}^{K} B_{ij} v_j\right) + 2MA_i d_i$.

The feedback structure is similar to that of DF-MUD, but error propagation is avoided here because the feedback in the transmitter is based on the actual values of user data instead of past decisions. Following the FF filtering and signature sequence spreading, the final transmitted signal at BS is $x(t) = \mathbf{s}^T(t)\mathbf{G}\mathbf{v}$. The average transmit energy per bit is given by

$$E = E_{\mathbf{b}}\left\{\int_0^T x^2(t)\mathrm{d}t\right\} = E_{\mathbf{b}}\left\{\int_0^T \mathbf{v}^T\mathbf{G}^T\mathbf{s}(t)\mathbf{s}^T(t)\mathbf{G}\mathbf{v}\mathrm{d}t\right\} = E_{\mathbf{b}}\{\mathbf{v}^T\mathbf{G}^T\mathbf{R}\mathbf{G}\mathbf{v}\} = E_{\mathbf{b}}\{\mathbf{v}^T\mathbf{v}\} = E_{\mathbf{b}}\{\mathbf{v}^T\mathbf{R}\mathbf{v}\}, \tag{5}$$

where $E_{\mathbf{b}}\{\cdot\}$ represents the expectation over the data symbol vector $\mathbf{b}$. The last step in (5) is satisfied by assuming $b_1, b_2, \ldots, b_K$, and $v_1, v_2, \ldots, v_K$ are independent and approximately uniformly distributed [9]. The result of (5) demonstrates that the filter $\mathbf{G}$ does not affect the Tx signal power, and, thus, power scaling is not required.

As shown in Fig. 1, the received signal is passed through a scaled MF, where f_{ii} is the ith diagonal element of $\mathbf{F}$, $i = 1, 2, \ldots, K$. Since f_{ii} is completely determined by the signature sequences, complexity of the THP receiver is comparable to that of the conventional single user receiver. The output vector of the scaled MF bank is

$$\mathbf{y} = diag(\mathbf{F})^{-1}(\mathbf{RGv} + \mathbf{n}) = diag(\mathbf{F})^{-1}[\mathbf{RG}(\mathbf{B}+\mathbf{I})^{-1}(\mathbf{b}+2M\mathbf{Ad}) + \mathbf{n}]. \tag{6}$$

Substituting the definitions of the FB and FF filters into (6), we obtain

$$\mathbf{y} = \mathbf{b}+2M\mathbf{Ad}+\mathbf{z}. \tag{7}$$

Mod-2M operations are then applied for users 1 through $K-1$ to eliminate the term $2M\mathbf{Ad}$. The noise component $\mathbf{z} = diag(\mathbf{F})^{-1}\mathbf{n}$ is white with the autocorrelation matrix $N_0 diag(\mathbf{F})^{-2}$. For the ith user, $i = 1, 2, \ldots, K$, the average transmit signal to noise ratio (SNR) per bit, γ_{bi}, is given by [20]

$$\gamma_{bi} \equiv \frac{E_{bi}}{N_0} = \frac{(M^2-1)A_i^2}{6N_0\log_2 M}, \tag{8}$$

then the theoretical symbol error rate (SER) is

$$Pe_i(\gamma_{bi}) = \frac{2(M-1)}{M} Q\left(\sqrt{\frac{6(\log_2 M) f_{ii}^2}{M^2-1}\gamma_{bi}}\right). \tag{9}$$

For flat Rayleigh fading channels, in the symbol interval of interest, the channel gain for the ith user is denoted as $c_i = \alpha_i \mathrm{e}^{\mathrm{j}\varphi_i}$, where the envelope α_i has Rayleigh distribution and the phase φ_i is uniformly distributed over $(-\pi, \pi]$, $i = 1, 2, \ldots, K$. Define the channel gain matrix as $\mathbf{C} = diag\{c_i\}_{K\times K}$. The signal received at the ith user receiver is $r_i(t) = c_i\mathbf{s}^T(t)\mathbf{v}+n_i(t)$. With THP precoding derived as for the AWGN channel, the output vector of the MF bank equals

$$\mathbf{y} = \mathbf{CRv}+\mathbf{n} = \mathbf{C}diag(\mathbf{F})(\mathbf{b}+2M\mathbf{Ad})+\mathbf{n}. \tag{10}$$

The received signals are detected coherently, scaled and passed through the mod-2M operators. Given the average transmit SNR per bit as (8), the average SER can be expressed as [20]

$$Pe_i(\gamma_{bi}) = \frac{M-1}{M}\left(1-\sqrt{\frac{3(\log_2 M) f_{ii}^2 \gamma_{bi} E\{\alpha_i^2\}}{M^2-1+3(\log_2 M) f_{ii}^2 \gamma_{bi} E\{\alpha_i^2\}}}\right). \tag{11}$$

The SER given by (9) and (11) are identical to the ideal SER of DF-MUD [17] (assuming no error propagation) for the AWGN and flat fading channels, respectively. Since DF-MUD improves upon the linear decorrelating MUD for all but the first user, the proposed THP method also has better performance than the linear decorrelating MUD. Furthermore, if equal transmit powers are employed for all users and all the transmit signals have the same cross-correlation, the linear decorrelating precoding for the downlink [2] yields the same performance as the linear decorrelating MUD for the uplink, and is outperformed by THP. The advantage of THP over linear schemes will be further demonstrated in section IV. The THP in (2, 3) satisfies decorrelating (zero-forcing) criterion. Moreover, it is proven in [2] that the ZF solution of the linear precoding filter is identical to the MMSE solution. Using similar derivation, it can be shown that the method defined by (2, 3) provides the optimal THP solution in the MMSE sense.

It is observed from (9) and (11) that for THP approach, the order of users affects individual performance. In particular, for the first user, the ideal performance of the THP method equals that of the linear decorrelating MUD; for the last user, the ideal SER achieves the single user bound (SUB) since $f_{KK} = 1$.

Although (9) and (11) give the theoretical SER, the actual performance of THP systems is influenced by the mod-2M operation. Note that in the transmitter the mod-2M operator output v_i is in the range $(-A_iM, A_iM]$, which is larger than the range of original PAM symbol b_i, $[-A_i(M-1), A_i(M-1)]$, i.e., the signal power reduced by modulo operation is still larger than the original signal power. This power penalty is not related to the original transmit SNR. It diminishes as M increases and can be ignored for large M ($M \geq 8$) [9]. It also should be noted the actual performance is slightly worse than the ideal performance because of the "end effect" of modulo operations in the receiver. The property that the detection errors for outer signals are less likely than for other signals is lost due to the mod-2M operations in the receiver, because any outer symbol will be re-assigned a bounded magnitude. The end effect also diminishes as M increases [9]. The influence of mod-2M operation is further observed in the simulation examples in Section IV.

III. NONLINEAR PRECODING IN MULTIPATH CHANNELS

In this section, we assume the transmitted signals are subject to frequency-selective multipath fading. Suppose there are N resolvable multipath components for every user, and $c_{i,n} = \alpha_{i,n} \mathrm{e}^{\mathrm{j}\varphi_{i,n}}$ represents the gain of the nth path component over the channel between the transmitter and the ith user receiver, $\forall\, i = 1,2,\ldots,K$ and $n = 0,1,\ldots, N$-1. For each user, define the vector

$\mathbf{c}_i = [c_{i,1}, c_{i,2},\ldots, c_{i,K}]$. By assuming the multipath spread is small relative to the symbol duration, the inter-symbol interference (ISI) is minor and therefore ignored in the remaining discussion. Without transmitter precoding, the equivalent baseband received signal at the ith mobile user receiver site can be expressed as $r_i(t) = \sum_{k=1}^{K}\sum_{n=0}^{N-1} c_{i,n} b_k s_k(t - nT_c) + n_i(t),$ where T_c is the chip duration and $n_i(t)$ is complex AWGN with power spectral density N_0. For the downlink of CDMA systems, in an attempt to minimize the complexity of mobile user receivers, we combine THP pre-decorrelating and diversity techniques in the transmitter at the BS instead of using RAKE receiver at the MS, so that the receivers remain as simple as those for single-user single-path channels.

A. PreRAKE Tomlinson-Harashima Precoding (PreRAKETHP)

The transmitter structure of PreRAKETHP for a 2-user 2-channel path/user system is shown in Fig.2. The basic THP decorrelating structure is similar to that for the single-path case, and is followed by a pre-RAKE combiner. The FB and FF filters for PreRAKETHP are denoted by $\mathbf{B}_\mathrm{P}$ and $\mathbf{G}_\mathrm{P}$, respectively. To avoid the transmit power increase due to multipath pre-RAKE combining, we use the normalized fading channel coefficients in the transmitter. For the ith user, define the normalization factor $S_i \equiv \left(\sum_{n=0}^{N-1} \left|c_{i,n}\right|^2\right)^{-1/2}$, then the normalized fading coefficient is obtained by $\hat{c}_{i,n} \equiv S_i\, c_{i,n}$, and the normalized vector corresponding to $\mathbf{c}_i$ is $\hat{\mathbf{c}}_i = [\hat{c}_{i,1}, \hat{c}_{i,2},\ldots, \hat{c}_{i,K}]$. With the same derivation as for single-path channels, the output vector of the mod-2M operator bank satisfies $\mathbf{v} = (\mathbf{B}_\mathrm{P}+\mathbf{I})^{-1}(\mathbf{b}+2M\mathbf{A}\mathbf{d})$. As we will see later, $\mathbf{B}_\mathrm{P}$ is a complex matrix related to the channel fading; as a result, the inputs of the mod-2M operators are complex numbers. The mod-2M operations are performed to constrain the magnitudes of the real part and imaginary part of the input number respectively. Thus for the ith user, d_i is a complex number with integral real and imaginary parts. As in Fig. 2, the outputs of the FF filter $\mathbf{G}_\mathrm{P}$ are multiplied with the conjugates of normalized fading coefficients. 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 matrix $\hat{\mathbf{C}} = diag\{\hat{\mathbf{c}}_1, \hat{\mathbf{c}}_2, \ldots, \hat{\mathbf{c}}_K\}$. Let $\mathbf{S} = diag\{S_1, S_2, \ldots, S_K\}$, then $\hat{\mathbf{C}} = \mathbf{SC}$. The output vector of the pre-RAKE combiner, $\mathbf{w} = \hat{\mathbf{C}}^H \mathbf{G}_\mathrm{P} \mathbf{v}$, has $K \times N$ elements. Following the signature sequence spreading, the ultimate transmitted signal is given by

$$x(t) = \sum_{k=1}^{K} \sum_{l=0}^{N-1} w_{kN-l} s_k (t - lT_c) . \tag{12}$$

The equivalent baseband received signal at the ith user receiver site is given by

$$r_i(t) = \sum_{k=1}^{K} \sum_{n=0}^{N-1} \sum_{l=0}^{N-1} c_{i,n} w_{kN-l} s_k (t - lT_c - nT_c) + n_i(t) . \tag{13}$$

The output of the MF is sampled at $t = (N-1)T_c$, and for the ith user is given by

$$y_i = \int_{(N-1)T_c}^{T+(N-1)T_c} r_i(t) s_i (t - (N-1)T_c) dt . \tag{14}$$

The cross-correlations between delayed signature sequences can be represented as

$$M_{i,k}^{m} = \int_0^T s_i(t) s_k(t + mT_c) dt, \qquad m \in \{-(N-1), \ldots, (N-1)\}. \tag{15}$$

From (13), (14) and (15), we have

$$y_i = \sum_{k=1}^{K} \sum_{n=0}^{N-1} \sum_{m=-(N-1)}^{N-1} c_{i,n} M_{i,k}^{m} w_{(k-1)N+m+n+1} + n_i , \tag{16}$$

where n_i is the filtered noise component with power N_0. $\forall\, i, k \in \{1,2,\ldots K\}$, define the $N \times N$ correlation matrix $\mathbf{M}_{i,k}$ with the jth row equal $[M_{i,k}^{1-j}, M_{i,k}^{2-j}, \ldots, M_{i,k}^{N-j}]$, j=1, 2,…, N. Construct the ($K \times N$)-row and ($K \times N$)-column matrix $\mathbf{M}$ by superimposing the $K \times K$ non-overlapping submatrices $\mathbf{M}_{i,k}$, $i,k \in \{1,2,\ldots,K\}$. Then (16) reduces to

$$\mathbf{y} = \mathbf{CMw} + \mathbf{n} = \mathbf{S}^{-1} \hat{\mathbf{C}} \mathbf{M} \hat{\mathbf{C}}^H \mathbf{G}_\mathrm{P} (\mathbf{B}_\mathrm{P} + \mathbf{I})^{-1} (\mathbf{b} + 2M\mathbf{Ad}) + \mathbf{n}, \tag{17}$$

where $\mathbf{y} = [y_1, y_2, \ldots, y_K]^T$ and $\mathbf{n} = [n_1, n_2, \ldots, n_K]^T$. Define $\mathbf{R}_\mathrm{P} \equiv \mathbf{CMC}^H$ and $\hat{\mathbf{R}}_\mathrm{P} \equiv \mathbf{S}^2 \mathbf{R}_\mathrm{P} = \hat{\mathbf{C}} \mathbf{M} \hat{\mathbf{C}}^H$. The matrix $\mathbf{R}_\mathrm{P}$ is obviously positive definite. Consequently, it can be factorized as $\mathbf{F}_\mathrm{P}^H \mathbf{F}_\mathrm{P}$ using the Cholesky factorization, where $\mathbf{F}_\mathrm{P} = \{f_{\mathrm{P}ij}\}_{K \times K}$ is complex lower triangular matrix with real diagonal elements. The normalized matrix corresponding to $\mathbf{F}_\mathrm{P}$ is $\hat{\mathbf{F}} = \{\hat{f}_{ij}\}_{K \times K} = \mathbf{SF}_\mathrm{P}$. To cancel the MAI, the FB and FF filters are defined as

$$\mathbf{B}_\mathrm{P} \equiv diag(\hat{\mathbf{F}})^{-1} \hat{\mathbf{F}}^H - \mathbf{I} , \tag{18}$$

$$\mathbf{G}_\mathrm{P} \equiv \hat{\mathbf{F}}^{-1}. \tag{19}$$

Then equation (17) can be simplified as

$$\mathbf{y} = diag(\mathbf{F}_{\mathrm{P}})(\mathbf{b}+2M\mathbf{A}\mathbf{d}) + \mathbf{n}. \tag{20}$$

We assume the channel gains and, therefore, $f_{\mathrm{P}ii}$ are estimated in the receivers. Then for the *i*th user, we have $f_{\mathrm{P}ii}^{-1}y_i = b_i + 2MA_id_i + f_{\mathrm{P}ii}^{-1}n_i$. The term $2MA_id_i$ will then be cancelled by mod-2M operation. The instantaneous ideal SER of the *i*th user is given by

$$Pe_i(\gamma_{bi}) = \frac{2(M-1)}{M} Q\left(\sqrt{\frac{6(\log_2 M) f_{\mathrm{P}ii}^{\;2}}{M^2-1}\gamma_{bi}}\right). \tag{21}$$

The PreRAKETHP described above is related to several precoders and MUDs for frequency-selective CDMA channels. It represents the transmitter precoding version of the nonlinear multiuser detector that consists of the RAKE receiver followed by the DF MUD. This detector can be viewed as a nonlinear (DF) version of the RAKE Decorrelating Detector (RDD) [14], the optimum linear MUD for multipath channels (or, equivalently, of the pre-RDD precoder in [6]). In this detector, the decorrelating DF processing is applied at the output of the filter bank matched to the signature sequences convolved with the channel responses for all users (the RAKE receiver), while in the RDD, the matched filter bank is followed by the linear decorrelating detection. Therefore, this DF detector is the optimal decorrelating DF MUD for the frequency selective CDMA channel. Since PreRAKETHP is the equivalent Tx-based implementation of this DF detector, it is the optimal THP method according to the decorrelating (ZF) and MMSE criteria.

To compare the nonlinear PreRAKETHP with the linear RDD, first we note that the correlation matrix $\mathbf{R}$ in [14] corresponds to $\mathbf{R}_{\mathrm{P}}^{T}$ above. Since $\mathbf{R}_{\mathrm{P}}^{-T} = (\mathbf{F}_{\mathrm{P}}^{-T})^H\mathbf{F}_{\mathrm{P}}^{-T}$, $(\mathbf{R}^{-1})_{ii}$ in equation (7) in [14] is equal to or larger than $f_{\mathrm{P}ii}^{-2}$ (equality is satisfied for i=1). Comparing the error rate formulas for PreRAKETHP (equation (21)) and RDD (equation (7) in [14]), we observe that PreRAKETHP outperforms the RDD for all users i>1. For the first user, these methods have the same SER. For the last user, since $f_{\mathrm{P}KK}^{\;2} = [\mathbf{R}_{\mathrm{P}}]_{KK} = \mathbf{c}_K[\mathbf{M}]_{KK}\mathbf{c}_K^{\;H}$, the performance of PreRAKETHP given by equation (21) achieves that of the RAKE receiver in the absence of

MAI. Similarly, it can be demonstrated that the PreRAKETHP has better performance than the pre-RDD in [6]. Since it has been proved in [14] that the RDD outperforms the Multipath Decorrelating Detector (MDD) [16], the PreRAKETHP also outperforms the MDD.

B. Multipath Decorrelating Tomlinson-Harashima Precoding (MDTHP)

One drawback of 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. In this section, we present an alternative THP design, MDTHP, to alleviate this problem.

The transmitter diagram for the MDTHP is shown in Fig. 3. Because $\mathbf{M}$ is symmetric and positive definite in practice, we can decompose $\mathbf{M} = \mathbf{F}_\mathrm{M}{}^T\mathbf{F}_\mathrm{M}$ by Cholesky factorization, where $\mathbf{F}_\mathrm{M}$ is a lower triangular matrix. Divide $\mathbf{F}_\mathrm{M}$ into $K\times K$ blocks, where each block is an $N\times N$ submatrix; and represent the block on the ith row and jth column as $[\mathbf{F}_\mathrm{M}]_{ij}$, $\forall\ i, j = 1, 2,\ldots, K$. The FB loop in the transmitter works in the following way. First, the N-dimension vector $\mathbf{v}_K$ is computed by applying N weights to the input symbol of the last user, $\mathbf{v}_K = b_K \cdot ([\mathbf{F}_\mathrm{M}]_{KK}\hat{\mathbf{c}}_K{}^H)$. 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, $j = i+1, i+2, \ldots, K$, is given by $\dfrac{\hat{\mathbf{c}}_i[\mathbf{F}_\mathrm{M}]_{ji}{}^T\mathbf{v}_j}{\sqrt{\beta_i\beta_j}}$, where $\beta_i \equiv \hat{\mathbf{c}}_i[\mathbf{F}_\mathrm{M}]_{ii}{}^T[\mathbf{F}_\mathrm{M}]_{ii}\hat{\mathbf{c}}_i{}^H$. Thus, for users 1 through $K–1$, the output of the FB loop is

$$\mathbf{v}_i = \left(b_i - \sum_{j=i+1}^{K}\frac{\hat{\mathbf{c}}_i[\mathbf{F}_\mathrm{M}]_{ji}{}^T\mathbf{v}_j}{\sqrt{\beta_i\beta_j}} + 2MA_i d_i\right)\cdot[\mathbf{F}_\mathrm{M}]_{ii}\hat{\mathbf{c}}_i{}^H, \tag{22}$$

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

$$S_{vi} = (\hat{\mathbf{c}}_i[\mathbf{F}_\mathrm{M}]_{ii}{}^T[\mathbf{F}_\mathrm{M}]_{ii}\hat{\mathbf{c}}_i{}^H)^{-1/2} = \beta_i^{-1/2}, \tag{23}$$

Let $\hat{\mathbf{v}}_i = S_{vi} \cdot \mathbf{v}_i$, $i = 1, 2, \ldots, K$. 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 $\mathbf{w} = \mathbf{G}_{\mathrm{M}}\hat{\mathbf{v}}$, where the FF filter is defined as $\mathbf{G}_{\mathrm{M}} = \mathbf{F}_{\mathrm{M}}^{-1}$. Following the derivation similar to that for the PreRAKETHP above, we obtain the matched filter bank output in the receivers as $\mathbf{y} = \mathbf{CMw} + \mathbf{n}$. For user i, this output is

$$y_i = \sqrt{\beta_i} \big/ S_i \, (b_i + 2MA_i d_i) + n_i, \qquad i = 1, 2, \ldots, K, \tag{24}$$

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

$$Pe_i(\gamma_{bi}) = \frac{2(M-1)}{M} Q\left(\sqrt{\frac{6(\log_2 M)\mathbf{c}_i [\mathbf{F}_{\mathrm{M}}]_{ii}^T [\mathbf{F}_{\mathrm{M}}]_{ii} \mathbf{c}_i^H}{M^2 - 1} \gamma_{bi}} \right) \tag{25}$$

The MDTHP method significantly simplifies transmitter precoding compared to the PreRAKETHP, since it employs the factorization of the channel gain-independent matrix **M**. Thus even for rapidly varying mobile radio channels the matrix factorization and inversion operations do not have to be performed frequently. These filters depend only on the signature sequences and the order of the users. The MDTHP is related to several other simplified MUD and Tx precoding methods that also employ the matrix **M**. First, equation (25) is identical to the theoretical performance of the Multipath Decorrelating Decision Feedback Receiver (MDDFR) (equation (16) in [15]). However, since DF MUD is degraded by error propagation, the MDTHP has better actual performance than the MDDFR. For the purpose of comparing the MDTHP with linear decorrelating precoders, in Appendix we briefly describe a linear multipath decorrelating precoder closely related to the MDTHP. This linear precoder can be regarded as the Tx-based counterpart of the Multipath Decorrelating Detector (MDD) [16]. Therefore we call it Pre-MDD. Compare the decision statistic of the MDTHP, $\mathbf{c}_i[\mathbf{F}_{\mathrm{M}}]_{ii}^T[\mathbf{F}_{\mathrm{M}}]_{ii}\mathbf{c}_i^H$, with that of the Pre-MDD (c.f. equation (A.1)), $\mathbf{c}_i([\mathbf{M}^{-1}]_{ii})^{-1}\mathbf{c}_i^H$. Based on the theorem for the inverse of a partitioned symmetric matrix [23, p.18], it is easy to prove that $\mathbf{c}_i[\mathbf{F}_{\mathrm{M}}]_{ii}^T[\mathbf{F}_{\mathrm{M}}]_{ii}\mathbf{c}_i^H \geq \mathbf{c}_i([\mathbf{M}^{-1}]_{ii})^{-1}\mathbf{c}_i^H$ (the equality is satisfied for $i = 1$). Therefore, for user 1, the MDTHP, Pre-MDD as well as MDD have the same performance; and for other users the MDTHP outperforms the Pre-MDD and MDD.

Since $f_{\mathrm{P}KK}^2 = \mathbf{c}_K[\mathbf{M}]_{KK}\mathbf{c}_K^H = \mathbf{c}_K[\mathbf{F}_\mathrm{M}]_{KK}^T[\mathbf{F}_\mathrm{M}]_{KK}\mathbf{c}_K^H$, by equations (21) and (25), the PreRAKETHP and MDTHP have the same performance for the last user. Their SER is the same as for an isolated single user with RAKE receiver. We can obtain some insight into the performance comparison of the two THP schemes for other users by considering their decision statistics averaged over the channel fading. Suppose the channel fading along the paths towards different user receivers are independent processes, it is easy to show that for $i = 1, 2, \ldots, K$, $E\{f_{\mathrm{P}ii}^2\} = E\{\mathbf{c}_i[\mathbf{M}]_{ii}\mathbf{c}_i^H\} \geq E\{\mathbf{c}_i[\mathbf{F}_\mathrm{M}]_{ii}^T[\mathbf{F}_\mathrm{M}]_{ii}\mathbf{c}_i^H\}$ ($E\{\cdot\}$ is the expectation over the channel gains), which indicates better performance of the PreRAKETHP [24]. This conclusion is further verified by simulations in the next section. On the other hand, the MDTHP is easier to implement as discussed above. In summary, there is a performance-complexity tradeoff between the PreRAKETHP and MDTHP.

IV. NUMERICAL RESULTS AND ANALYSIS

In this section, we provide the numerical evaluations of the performance of THP designs for single-path channels (Fig. 5-6) and multipath fading channels (Fig. 7-10), respectively.

First, consider the downlink of a 2-user M-PAM CDMA system over an AWGN channel. Suppose the transmit powers of the two users are equal, i.e., $A_1 = A_2$, and the two signature sequences have cross-correlation $R_{12} = 0.5$. For user 1, the theoretical SER of THP, linear decorrelating MUD [1] and linear decorrelating precoding [2] are equal in this case. Fig. 5 demonstrates the difference between the simulated and the theoretical SER for THP. There are two reasons for the gap between the theoretical and simulated SERs. One reason is the “end effect” of mod-2M operation in the receivers. The “end effect” diminishes as M increases. If M is fixed, the “end effect” is weakened as the SNR increases since the probability that the received symbols fall outside $(-A_1M, A_1M]$ decreases. The second reason is the transmit power increase due to mod-2M operation in the transmitter. This effect is reduced as M increases and is not related to the original transmit SNR. For M = 2, 4, 8, 16, the corresponding power penalty are

0.67dB, 0.08dB, 0.03dB and 0.006dB, respectively. It is observed that in this case the performance degradation due to mod-2M operation is negligible for 8-PAM systems with transmit SNR larger than 20dB, and for 16-PAM systems.

In Fig. 6, we consider a heavily loaded 3-user 8-PAM system over an AWGN channel. The cross-correlation between any two users is 0.8. All users have equal powers. We compare the performance of four methods: THP, linear decorrelating transmitter precoding (labeled as Dp.), linear decorrelating MUD receiver (labeled as Dr.) and DF-MUD. The single user bound (SUB) and the conventional system performance is also shown for comparison purposes. We notice that the linear decorrelating methods have the same performance for all users because of equal transmit powers and equal signal correlations. For THP the simulation results are very close to the theoretical performance which indicates the negative influence of mod-2M is slight. For THP, the best performance is for user 3, which achieves the SUB and has much lower error rate than the DF-MUD and linear methods; the worst performance is for user 1 and is theoretically the same as that of DF-MUD for user 1 and the linear methods for all users. For user 2, THP outperforms the linear methods and DF-MUD. Clearly, the THP scheme improves on the linear schemes for most users. Although THP approach and DF-MUD achieve the same performance theoretically, the actual performance of THP is better than that of DF-MUD since the latter one suffers from severe error propagation when user powers are similar.

In the examples above, all users' powers were equal both at the Tx and the Rx. Next, we address the multipath fading channels that result in different received powers. Power control is often needed in these channels. Since the order of users influences the SER in THP systems, we propose to select the order that saves the total transmit power. As mentioned in section II, the practical performance of THP schemes for M-PAM/QAM systems is degraded when M is small due to the power penalty and end effect of mod-2M operation. Usually the first user suffers the worst degradation and the last user is not influenced at all. Moreover, similarly to DF-MUD, as the users' order increases, the benefit of the THP becomes more pronounced. Therefore, to achieve more balanced performance and to minimize transmit power, we sort users in the order of decreasing received powers in the following simulation experiments. In practice, it might be desirable to maintain constant user order for an extended period of time to minimize complexity. (Note that unlike DF-MUD, THP is not affected by error propagation, so ordering users according to their powers is not required for maintaining reliable performance.)

In Fig. 7-10, we investigate the performance of transmitter precoding in multipath fading environments. In these examples, we employ orthogonal 32-chip Hadamard codes as the signature sequences. All channel paths experience i.i.d. Rayleigh fading and the total average channel power is normalized to one for each user. First, consider an 8-user, 4-channel paths/user system. Fig. 7 and Fig. 8 show the SER averaged over all users for BPSK and 16-QAM, respectively. (We do not employ power control in this paper, although in practice it is required to maintain the target SER.) The transmit powers of all users are equal. In both figures, the THP methods significantly outperform the conventional RAKE receiver and linear decorrelating precoding with RAKE receiver in [2] (labeled as "Lin. Prec. RAKE"). When the bit error rate (BER) is lower than 10^{-3}, PreRAKETHP is the best among all the Tx-based and Rx-based, linear and nonlinear decorrelating methods, and MDTHP is better than the MDDFR, Pre-MDD and MDD, as expected from theoretical analysis. For higher SNR values, both THP methods outperform other techniques. This confirms that the inherent advantage of nonlinear THP methods overcomes the adverse influence of modulo operation even if M is very small. The linear precoding in [2] is seriously degraded by transmit power scaling. The SUB is also shown for reference. It is given by the performance of isolated single user with RAKE receiver.

Next, consider a 4-user 3-channel paths/user BPSK system, where the transmit powers for all users are equal. In addition to short-term multipath Rayleigh fading, the transmitted signal is also subject to long-term shadow fading modeled by log-normal distribution with variance 6dB [25].

Suppose the average received signal powers for the four users satisfy the ratio of 8:4:2:1. The instantaneous received power order is corresponding to the order of average power, so bit-by-bit reordering is not performed. Fig. 9 demonstrates the BER of the weakest user. The proposed two nonlinear techniques and the linear techniques of Pre-RDD, Pre-MDD and linear precoder with RAKE receiver in [2] (labeled as "Lin. Prec. RAKE") are compared. As expected, for the last user PreRAKETHP and MDTHP result in the same performance as for the isolated single user with RAKE receiver (SUB), while other methods have poorer performance. Equivalently, the users that experience greater propagation loss will require lower power to satisfy the target BER. Thus, in practice the overall transmit power can be reduced when THP methods are employed.

A critical assumption for the above simulation experiments is that the channel state information (CSI) is perfectly known at the transmitter. Next we observe the performance of the THP techniques under the condition that the CSI is estimated at the Rx and fed back to the Tx. Consider a BPSK modulated 4-user 3-channel paths/user wideband-CDMA (W-CDMA) system with the following parameters; the carrier frequency is 2GHz, the maximum Doppler frequency is 200 Hz and the transmit data rate is 128kbps [26]. The frequency selective Rayleigh fading is simulated by the Jakes model [27]. In Fig. 10, the average BER of the proposed precoding methods is shown for three different cases: CSI is perfectly known at the transmitter; CSI is predicted; CSI is fed back to the transmitter (no prediction is used). For CSI prediction, the 3-step LRP algorithm for W-CDMA [19] is utilized with the channel sampling frequency of 1600Hz and the prediction range of 0.625ms (the slot interval). Since the channel sampling and prediction frequency is less than the data rate, it is necessary to interpolate the intermediate channel gain coefficients. In the case of delayed fed back CSI (FBCSI), the channel gains are fed back without prediction with the feedback delays of 0.3125ms and 0.625ms. Note that transmitter precoding aided by channel prediction achieves almost the same performance as for the case when the channel is perfectly known at the Tx, while feeding back delayed CSI without prediction results in significant performance degradation. Thus, accurate LRP is required for reliable Tx precoders in practical mobile radio channels.

V. CONCLUSIONS

In this paper, we investigated Tomlinson-Harashima transmitter precoding methods for the mitigation of MAI in the downlink of CDMA systems. The THP designs for synchronous single-path AWGN channels, flat Rayleigh fading channels and multipath fading channels have been described. Using theoretical analysis and simulation examples, it was shown that THP has better performance than the previously proposed linear precoders, and linear and nonlinear MUDs with similar complexity. Furthermore, when the received signal powers of all users are unequal, THP can significantly improve the performance of weaker users compared to linear precoding methods, thus enabling reduction of total average transmit power. It is also shown that the LRP enables performance of THP in practical W-CDMA channels.

APPENDIX — LINEAR PRE-MDD

Note that the (K×N)-row (K×N)-column real matrix $\mathbf{M}$ defined in section III.A is only determined by the spreading sequences and not related to channel gains. Fig. 4 shows the transmitter structure of Pre-MDD for an example of 2-user 2-path/user system. The decorrelating filter $\mathbf{G} = \mathbf{M}^{-1}$ is preceded by power control subfilters $\mathbf{T}_i=\tilde{S}_i([\mathbf{G}]_{ii})^{-1}$ for each user, in which $[\mathbf{G}]_{ii}$ is the submatrix that contains the elements on the $[(i-1)\times N+1]$th to $[i\times N]$th rows and columns of $\mathbf{G}$, and $\tilde{S}_i$ is a scaling factor, $\forall i = 1, 2, \ldots, K$. The total average transmit power per symbol interval should satisfy $E_{\mathbf{b}}\left\{\int_0^T |x(t)|^2 dt\right\} = E_{\mathbf{b}}\left\{\mathbf{b}^T\mathbf{b}\right\}$. For all K users, define the compact form $\mathbf{T} \equiv diag\{\mathbf{T}_1, \mathbf{T}_2, \ldots, \mathbf{T}_K\}$ and decorrelating filter output $\mathbf{w} \equiv [w_1, w_2, \ldots, w_{K\times N}]^T$. Since $E_{\mathbf{b}}\left\{\int_0^T |x(t)|^2 dt\right\} = E_{\mathbf{b}}\left\{\mathbf{w}^H\mathbf{M}\mathbf{w}\right\} = E_{\mathbf{b}}\left\{\mathbf{b}^T\hat{\mathbf{C}}\mathbf{T}^T\mathbf{G}^T\mathbf{M}\mathbf{G}\mathbf{T}\hat{\mathbf{C}}^H\mathbf{b}\right\}$, it is equivalent to require the individual users to satisfy $E_{b_i}\left\{b_i^2\hat{\mathbf{c}}_i\mathbf{T}_i^T[\mathbf{G}]_{ii}^T\mathbf{T}_i\hat{\mathbf{c}}_i^H\right\} = E_{b_i}\left\{b_i^2\right\}$, $i = 1, 2, \ldots, K$. Therefore, the scaling factor can be obtained by $\hat{\mathbf{c}}_i \mathbf{T}_i^T [\mathbf{G}]_{ii}^T \mathbf{T}_i \hat{\mathbf{c}}_i^H = \tilde{S}_i^2 \cdot \hat{\mathbf{c}}_i [\mathbf{G}]_{ii} \hat{\mathbf{c}}_i^H = 1$, that is, $\tilde{S}_i = (\hat{\mathbf{c}}_i [\mathbf{G}]_{ii} \hat{\mathbf{c}}_i^H)^{-1/2}$.

With the same receiver structure as in PreRAKETHP, the MF bank output is given by $\mathbf{y} = \mathbf{CMw} + \mathbf{n} = \mathbf{CT}\hat{\mathbf{C}}^H\mathbf{b} + \mathbf{n}$. For user i, the input of the decision device is $y_i = \mathbf{c}_i\mathbf{T}_i\hat{\mathbf{c}}_i^H b_i + n_i$. Thus, the SER of the ith user is given by

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

Comparing (A.1) with the theoretical performance of MDD [16], we find that the Pre-MDD and the MDD have identical performance. Note that although the decorrelating filter $\mathbf{M}^{-1}$ can pre-cancel the correlations among all the $K\times N$ signal paths, $\mathbf{T}_i$ incurs the correlations among the N signal paths for user i. Therefore, the self interference is not canceled in this approach. A closely related linear multipath decorrelating precoder was proposed in [8], in which the transmit power control is implemented by global power scaling, and all paths of a given user are completely decorrelated. Compared with other linear decorrelating precoders [2, 4], a common advantage of Pre-MDD and [8] is that the transmitter has low complexity since the decorrelating filter is not related to channel gains.

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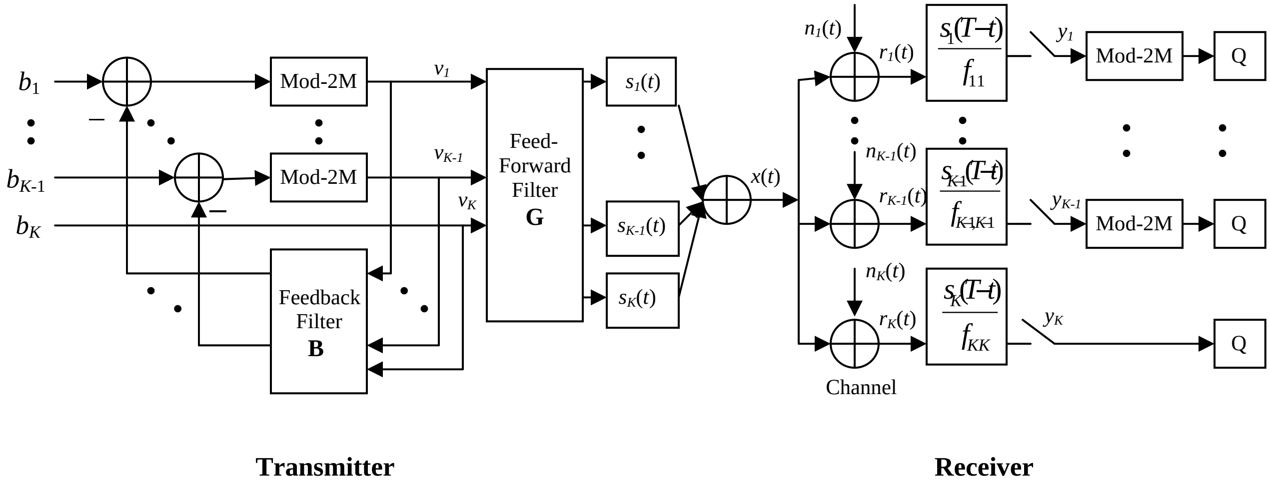

Fig. 1 The Diagram of THP for Single-path Channels

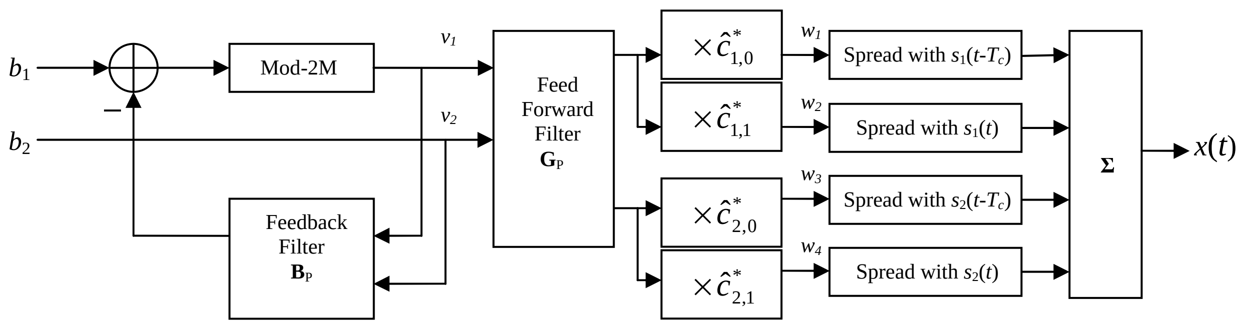

Fig. 2 The Transmitter of PreRAKETHP for Multipath Fading Channels (a 2-user 2-channel path/user system)

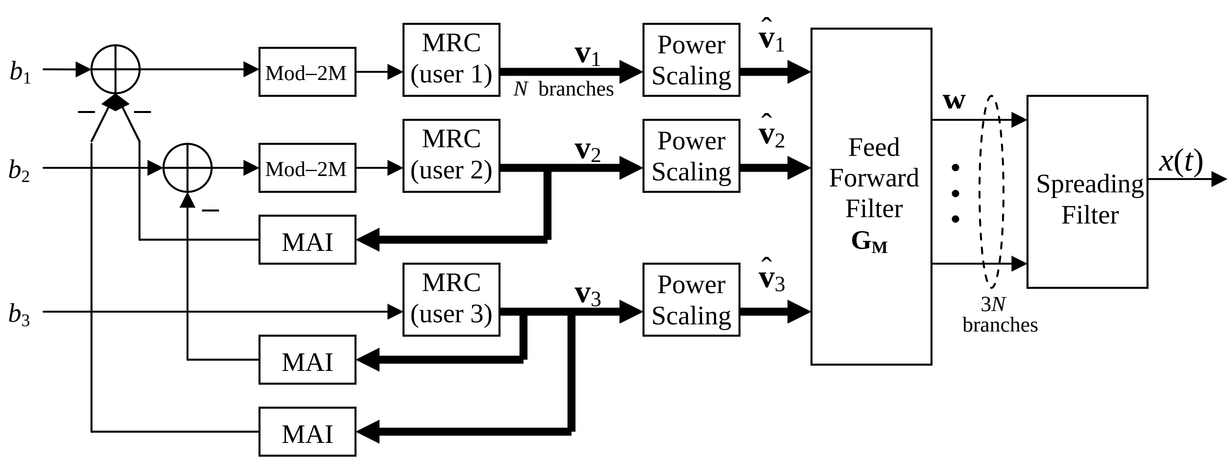

Fig. 3 The Transmitter of MDTHP for Multipath Channels (a 3-user *N*-channel path/user System)

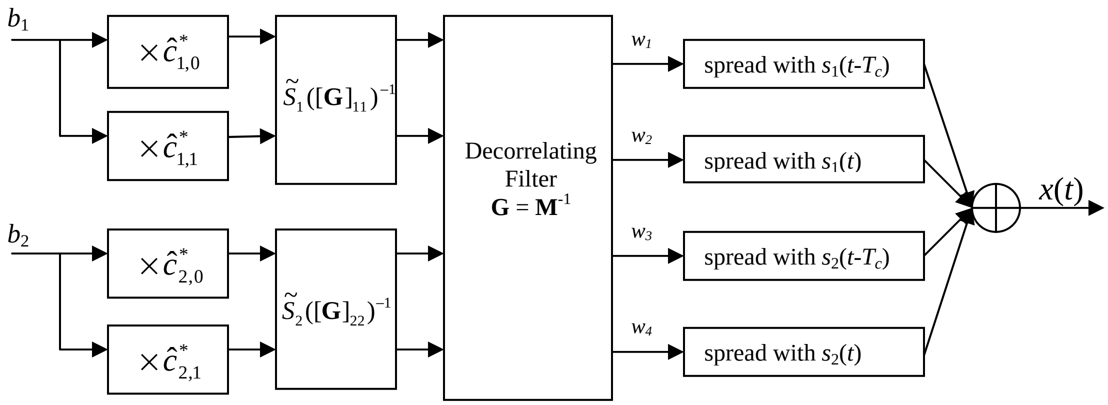

Fig. 4 The Transmitter of Pre-MDD for Multipath Fading Channels (a 2-user, 2-path/user system)

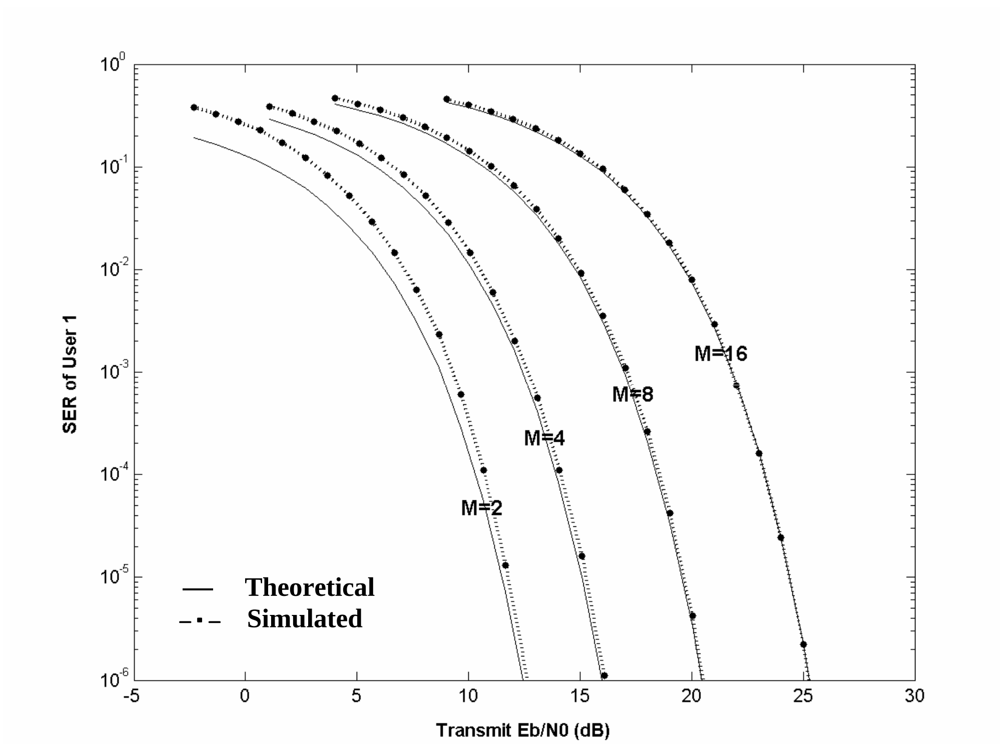

Fig. 5 Comparison of simulated and theoretical performance of THP with M = 2, 4, 8, 16, in AWGN channel, 2 users, A_1=A_2, R_{12} = 0.5.

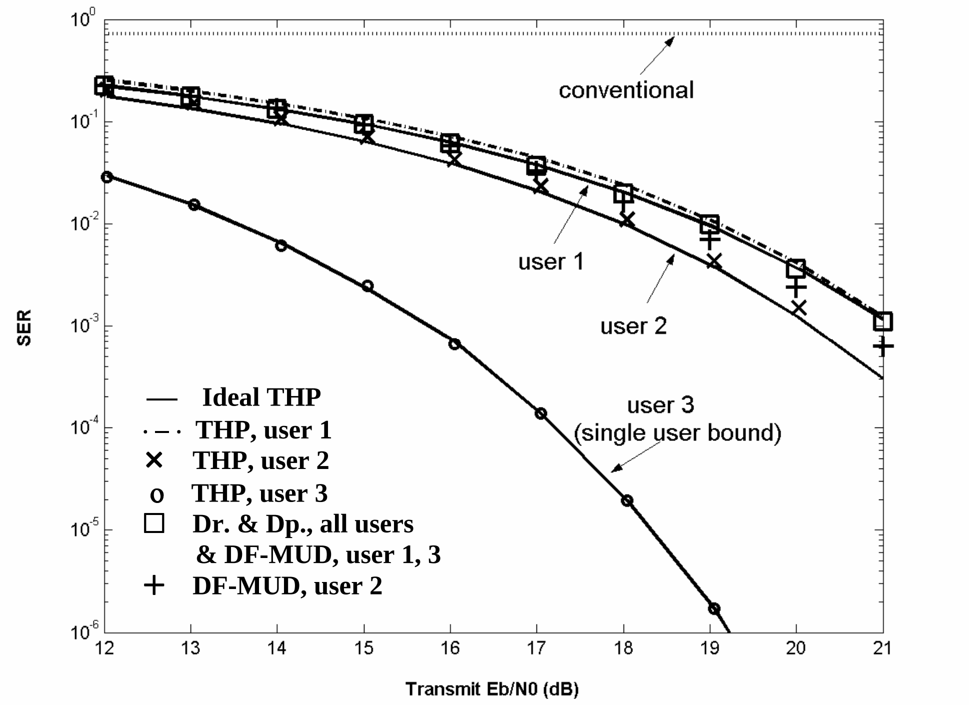

Fig. 6 Performance comparison of THP, DF-MUD, decorrelating precoding (Dp.) and decorrelating MUD (Dr.) in AWGN channel, 8-PAM, 3 users with equal transmit powers, $R_{12} = R_{13} = R_{23} = 0.8$.

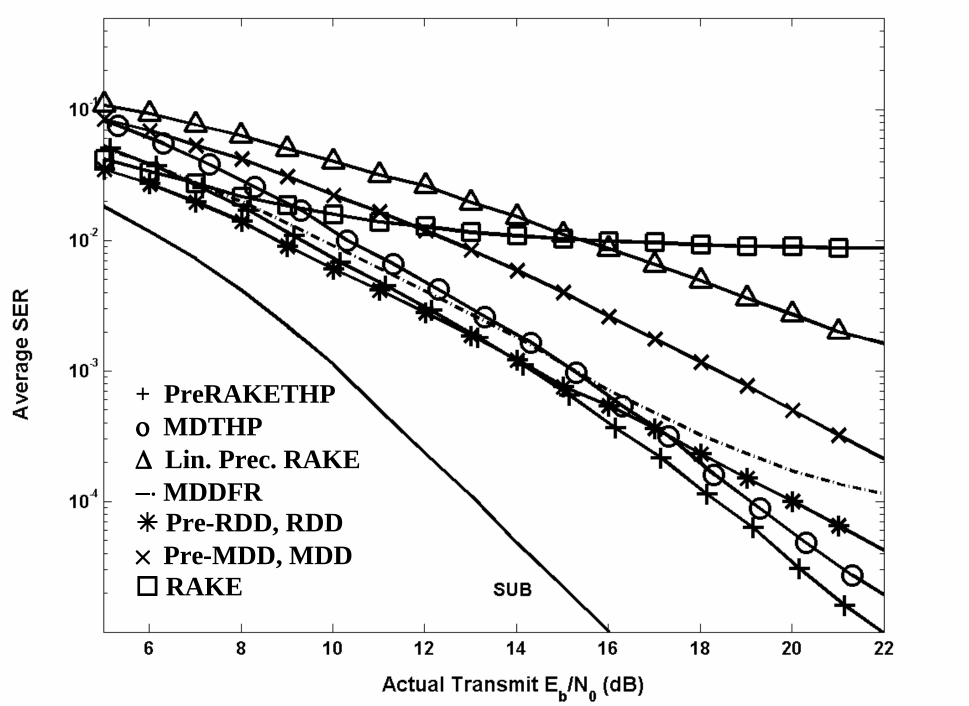

Fig. 7 Performance comparison of various techniques in multipath fading channels, 8 users with equal Tx powers, 4 channel-paths/user, BPSK.

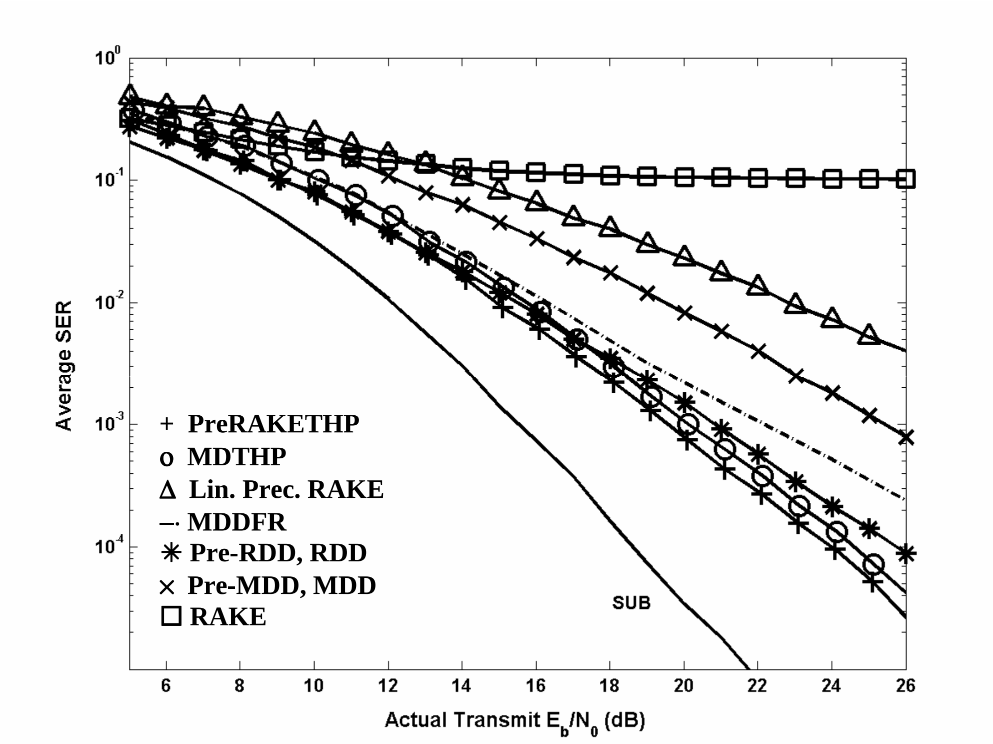

Fig. 8 Performance comparison of various techniques, 8 users with equal Tx powers, 4 channel-paths/user, 16-QAM.

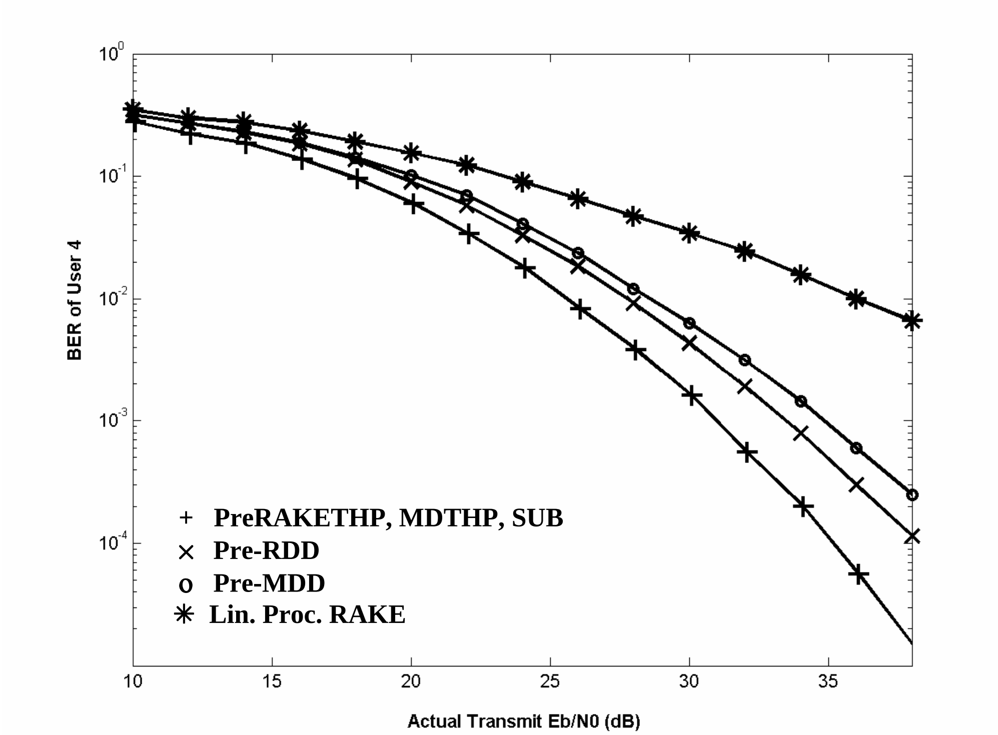

Fig. 9 Performance of the user with the weakest received signal power in large scale lognormal multipath fading channels, 4 users, 3 channel paths/user, equal transmit power for all users, the ratio of the received signal powers is 8:4:2:1, BPSK.

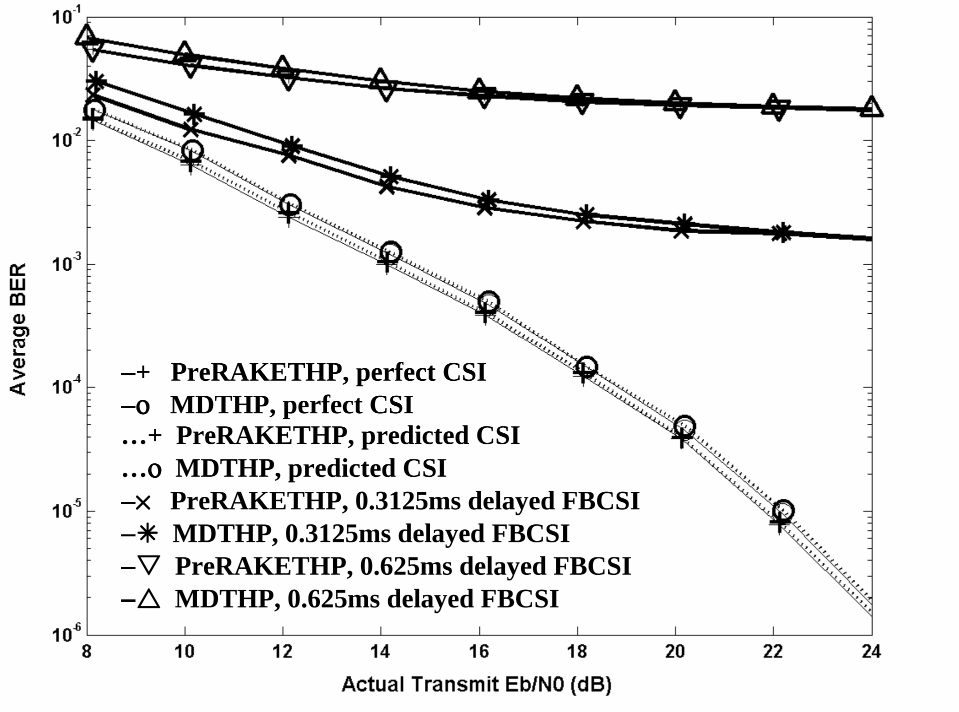

Fig. 10 Performance comparison of precoding aided by CSI prediction and by CSI feedback, 4 users with equal transmit powers, 3 channel paths/user, BPSK, data rate 128kbps, carrier frequency 2GHz, maximum Doppler frequency 200Hz, channel sampling frequency 1600Hz.