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Sincerely,

Alexandra Duel-Hallen

Tomlinson – Harashima Transmitter Precoding for Synchronous Multiuser Communications[1]

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Abstract — In this paper, a nonlinear Tomlinson Harashima transmitter-precoding (THP) approach is developed to suppress multiple access interference (MAI) for multiuser systems. A feedback and a feed-forward filter are jointly used to cancel MAI, which is similar to the receiver-based decision feedback (DF) multiuser detector (MUD); modulo operation is used to constrain the transmit power. The THP design for decentralized channels (DC) is suitable for the downlink of code division multiple-access (CDMA) systems; the THP for centralized channels (CC) can be applied in multicarrier, space-time and other multi-input multi-output (MIMO) systems. It is shown that THP improves upon linear MAI rejection techniques. It has the same ideal performance as DF-MUD, but in practice THP has lower error rate because it avoids error propagation.

I. INTRODUCTION

For the downlink of CDMA systems, the high complexity of MUD receivers [1] is usually problematic for a resource-constrained mobile unit. As an alternative to receiver (Rx)-based optimization schemes, linear transmitter (Tx)-based MAI rejection methods have been recently proposed, e.g. [2, 3]. These techniques apply linear transformation to the transmit signals to reject MAI. The least complex approach is linear decorrelating transmitter precoding [2]. Although this method can completely eliminate MAI, the system performance is inherently degraded by transmit power scaling.

In this paper, we propose a non-linear tx-based technique with the complexity similar to that of linear decorrelating transmitter precoding. In this scheme, a feedback and a feed-forward filter are jointly used to cancel MAI. This method is inspired by the idea of Decision-Feedback Equalization [4] and Decision-Feedback Multiuser Detector (DF-MUD) [5]. Modulo operations are applied to constrain the transmit power, as in the Tomlinson-Harashima algorithm [6].

We propose two specific THP designs for centralized channels (CC) and decentralized channels (DC), respectively. In the DC model, which corresponds to the downlink of CDMA systems, the transmitter is centralized; the receivers of different users are separate, resource-constrained and do not have the knowledge of spreading codes of other users. Therefore, the processing burden is placed in the transmitter. In the CC model, both the transmitter and the receiver are centralized for all users (or all sub-channels in multi-channel systems), which makes it possible to perform the Tx-Rx joint optimization with balanced complexity for the transmitter and receiver. The CC model with MIMO DF detector has been applied in multicarrier [10], space-time [12, 13] and other MIMO systems [11]. Independent research has been reported recently on transmitter-based THP for MIMO systems [8, 9]. However, the system models used in these papers are quite different from our models; therefore the derivation of the feedback and feed-forward filters is more complex. Moreover, we address fundamental performance equivalence of the transmitter-based and receiver-based designs for MIMO channels that is not included in [8, 9].

In the following section, CC and DC system models are presented first, and then several existing MAI cancellation schemes are briefly reviewed. In Section III, we describe the THP schemes for additive white Gaussian noise (AWGN) and flat Rayleigh fading channels. Simulation results and conclusions are provided in Section IV and V, respectively.

II. CC/DC SYSTEM MODELS AND MAI CANCELLATION METHODS

For a K-user system, denote the transmit data symbols during the symbol interval of interest $[0, T)$ by a vector $\mathbf{b} = (b_1, b_2, \ldots, b_K)^T$. Suppose the symbols are M-ary PAM modulated with the equivalent baseband minimum Euclidean distance $2A_i$, i.e., $b_i \in \{-(M-1)A_i, -(M-3)A_i, \ldots, (M-3)A_i, (M-1)A_i\}$, $\forall i = 1, 2, \ldots, K$. The proposed schemes can be easily extended to quadrature

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amplitude modulated (QAM) systems. Let the vector of normalized signature waveforms for K users be $\mathbf{s}(t) = [s_1(t), s_2(t), \ldots, s_K(t)]^T$. For the DC model, e.g. the downlink of a CDMA system, the equivalent low-pass received signal at the ith receiver is $r_i(t)=\sum_{i=1}^{K} b_i s_i(t)+n_i(t)$, where $n_i(t)$ represents white Gaussian noise with zero mean and power spectrum density N_0. In the conventional receiver, $r_i(t)$ is fed to the filter matched to the ith signature waveform. The resulting output is $y_i=\int_0^T r_i(t)s_i(t)\mathrm{d}t$, $\forall i$, $j = 1,2,\ldots,K$. If we define the correlation between two signature sequences as $R_{i,j}=\int_0^T s_i(t)s_j(t)\mathrm{d}t$, $\forall$ i, $j = 1,2,\ldots,K$, and the $K\times K$ correlation matrix $\mathbf{R} = \{R_{i,j}\}$, the matched filter (MF) output vector $\mathbf{y} = (y_1, y_2, \ldots, y_K)^T$ satisfies

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

In equation (1), $\mathbf{n}=(n_1, n_2, \ldots, n_K)^T$ is a zero-mean Gaussian noise vector with the elements $n_i = \int_0^T n_i(t)s_i(t)\mathrm{d}t$, $i=1,2,\ldots,K$, and the autocorrelation matrix $N_0\mathbf{I}$. ($\mathbf{I}$ is the $K\times K$ identity matrix.)

For the CC systems, the equivalent low-pass received signal at the centralized receiver is $r(t) = \sum_{i=1}^{K} b_i s_i(t)+n(t)$. The MF output has the same form as equation (1), but for CC model the autocorrelation matrix of $\mathbf{n}$ is $N_0\mathbf{R}$. Note that we retain the terminology of CDMA systems for the CC model. Although its primary applications are in multicarrier, space-time and other systems, the mathematical models for these applications are equivalent to those used in this paper.

From equation (1), we observe that MAI is caused by the non-diagonal elements of matrix $\mathbf{R}$. For CC systems, as well as for the uplink of DS-CDMA, the interference can be conveniently canceled by the linear decorrelating MUD [1]. The output is $\mathbf{R}^{-1}\mathbf{y} = \mathbf{b}+\mathbf{z}$. Since the linear MUD enhances noise, the nonlinear DF-MUD was proposed in [5]. The positive definite symmetric matrix $\mathbf{R}$ is decomposed by Cholesky factorization [7],

$$\mathbf{R} = \mathbf{F}^{\mathrm{T}}\mathbf{F}, \tag{2}$$

where $\mathbf{F}$ is a lower triangular matrix. For DF-MUD, a nonlinear feedback loop and a feed-forward filter are used to cancel the partial MAI expressed by $\mathbf{F}$ and $\mathbf{F}^T$ respectively. This approach avoids noise enhancement and outperforms linear decorrelating method; however, because the feedback is based on past decisions, the performance is degraded by error propagation.

For the DC systems, the linear decorrelating transmitter precoding has been investigated in [2]. The precoding filter is $\mathbf{T} = S_f\mathbf{R}^{-1}$, where S_f is the power scaling factor used to normalize the average transmit power. The transmit signal is $S_f\mathbf{s}^T(t)\mathbf{R}^{-1}\mathbf{b}$, hence, the MF output in the receivers is $\mathbf{y} = S_f\mathbf{b}+\mathbf{n}$. While this approach does not suffer from noise enhancement, its performance is degraded by transmit power scaling.

III. TOMLINSON – HARASHIMA TRANSMITTER PRECODING (THP) SCHEMES

1. THP for Centralized Channels (*THP-CC*)

Fig.1 shows the system diagram of THP-CC. The feedback filter in the transmitter is defined as

$$\mathbf{B}_{CC} = \{B_{ij}\}_{K\times K} = diag(\mathbf{F})^{-1}\times\mathbf{F}-\mathbf{I}, \tag{3}$$

where $\mathbf{F}$ is as in equation (2), and $diag(\mathbf{F})^{-1}=diag(\mathbf{F}^{-1})$ is the diagonal matrix that contains the diagonal elements of $\mathbf{F}^{-1}$. Thus, $\mathbf{B}_{CC}$ is a lower triangular matrix with zeros along the diagonal. A bank of mod-2M operators are used 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 $(-M, M]$. In Fig.1(a), the output vector of the mod-2M operator bank $\mathbf{v} = [v_1, v_2, \ldots, v_K]^T$ satisfies

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

where $\mathbf{A}=diag\{A_i\}_{K\times K}$, and $\mathbf{d}=[d_1,d_2,\ldots,d_K]^T$ 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 $(\mathbf{B}_{CC}+\mathbf{I})\mathbf{v} = \mathbf{b}+2M\mathbf{A}\mathbf{d}$. Note that $(\mathbf{B}_{CC}+\mathbf{I})$ is a lower triangular matrix with ones along the diagonal. Thus, $v_1=b_1$ and $d_1=0$, i.e., mod-2M is not required for user 1; for the ith user, $i=2,3,\ldots K$, we can successively determine d_i and v_i based on the values of $v_1, v_2, \ldots, v_{i-1}$,

$$v_i = (b_i - \sum_{j=1}^{i-1} B_{ij}v_j) + 2MA_i d_i.$$

The feedback structure is similar to that of DF-MUD. However, because the feedback in the transmitter is based on the actual values of user data instead of past decisions, error propagation is avoided.

The received signal is $r(t) = \mathbf{s}^T(t)\mathbf{v} + n(t)$. In Fig.1(b), the output vector of the MF bank, $\mathbf{y} = [y_1, y_2, \ldots, y_K]^T$, is

$$\mathbf{y} = \mathbf{R}\mathbf{v} + \mathbf{n}. \tag{5}$$

Given (3) and (4), equation (5) is equivalent to

$$\mathbf{y} = \mathbf{F}^T diag(\mathbf{F})(\mathbf{b}+2M\mathbf{A}\mathbf{d}) + \mathbf{n}. \tag{6}$$

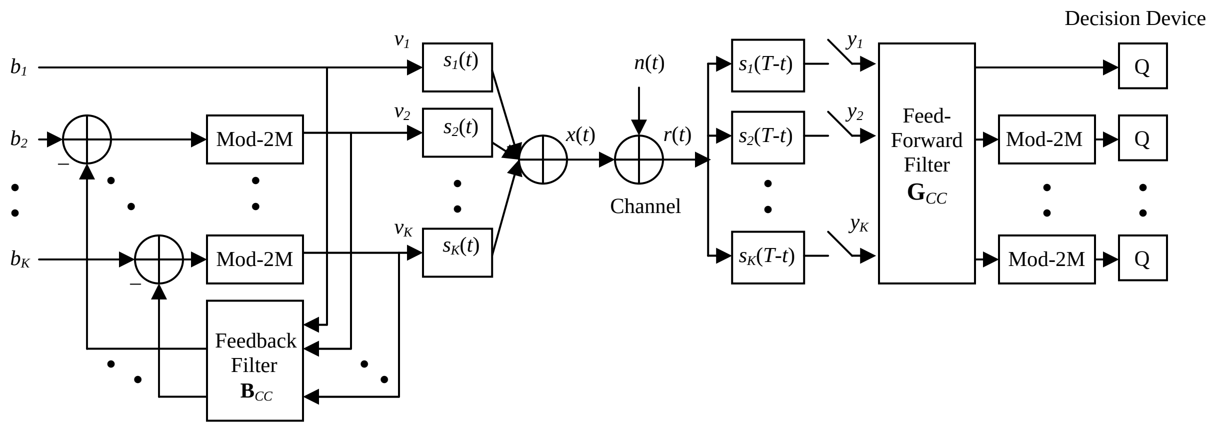
To recover the original data $\mathbf{b}$, the feed-forward filter should be defined as

$$\mathbf{G}_{CC} = diag(\mathbf{F})^{-1}\mathbf{F}^{-T}, \tag{7}$$

Where $\mathbf{F}^{-T} = (\mathbf{F}^{-1})^T = (\mathbf{F}^T)^{-1}$, thus the output of $\mathbf{G}_{CC}$ is

$$\mathbf{G}_{CC}\mathbf{y} = \mathbf{b} + 2M\mathbf{A}\mathbf{d} + \mathbf{z}_{CC}, \tag{8}$$

where $\mathbf{z}_{CC} = \mathbf{G}_{CC}\,\mathbf{n}$. Equation (8) shows that MAI has been completely eliminated at the output of the feed-forward filter. Finally, the mod-2M operations are



(a) Transmitter **(b) Receiver**

Fig. 1 Diagram of THP for a Centralized-Channel (CC) System

performed for users 2 through K (mod-2M is not required for user 1 because $d_1 = 0$). The input vector to the detector is $\mathbf{b}+\mathbf{z}_{CC}$. The autocorrelation matrix of $\mathbf{z}_{CC}$ is $N_0 diag(\mathbf{F})^{-2}$, that is, the noise is whitened. The average transmit signal to noise ratio (SNR) per bit for the ith user, γ_{bi}, equals [4]

$$\gamma_{bi} \equiv \frac{E_{bi}}{N_0} = \frac{(M^2-1)A_i^2}{6N_0 \log_2 M}, \quad i = 1,2,\ldots,K. \qquad (9)$$

Then the theoretical symbol error rate (SER) is

$$Pe_i = \frac{2(M-1)}{M} Q\left(\sqrt{\frac{6\log_2 M}{M^2-1} f_{ii}^2 \gamma_{bi}}\right), \quad i = 1,2,\ldots,K, \qquad (10)$$

which is identical to that of DF-MUD assuming no error propagation. (f_{ii} is the ith diagonal element of $\mathbf{F}$)

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 signal 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 diminishes as M increases and can be ignored for large M ($M \geq 8$) [6]. 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. However, as observed in the simulation examples, the end effect also diminishes as M increases [6].

***2. THP for Decentralized Channels* (*THP-DC*)**

The scheme proposed in previous section can be applied in DC systems by using adaptive linear processing in the receivers. However, it is often desirable to make the receivers of DC systems (e.g. the mobile stations in downlink CDMA) as simple as possible. Therefore, in this section we propose a different THP design in which the receivers are as simple as in a single-user system. The diagram of this design is shown in Fig.2. Although relocation of the feed-forward filter from the receiver to the transmitter appears to be the only difference relative to Fig.1, this alteration actually causes the changes in the structures of the feedback and the feed-forward filters denoted by $\mathbf{B}_{DC}$ and $\mathbf{G}_{DC}$, respectively. Similarly to equation (4), we obtain $\mathbf{v} = (\mathbf{B}_{DC}+\mathbf{I})^{-1}(\mathbf{b}+2M\mathbf{Ad})$. The output of the feed-forward filter is $\mathbf{G}_{DC}\mathbf{v}$. The signal received at the ith user's receiver site can be expressed as $r_i(t) = \mathbf{s}^T(t)\mathbf{G}_{DC}\mathbf{v}+n_i(t)$. With transmitter precoding, the output vector of the scaled MF bank is

$$\begin{aligned}\mathbf{y} &= diag(\mathbf{F})^{-1}(\mathbf{R}\mathbf{G}_{DC}\mathbf{v} + \mathbf{n}) \\ &= diag(\mathbf{F})^{-1}[\mathbf{F}^T\mathbf{F}\mathbf{G}_{DC}(\mathbf{B}_{DC}+\mathbf{I})^{-1}(\mathbf{b}+2M\mathbf{Ad}) + \mathbf{n}]. \qquad (11)\end{aligned}$$

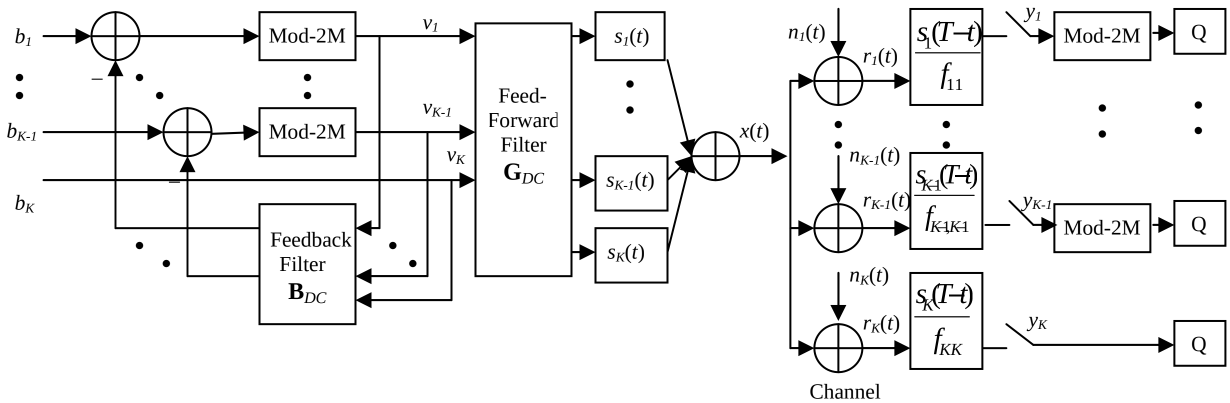
Although the ith user receiver needs the knowledge of f_{ii}, this requirement does not significantly increase receiver complexity. To cancel MAI, we define the feedback filter $\mathbf{B}_{DC}$ and the feed-forward filter $\mathbf{G}_{DC}$ as

$$\mathbf{B}_{DC} = diag(\mathbf{F})^{-1}\mathbf{F}^T\text{-}\mathbf{I}, \qquad (12)$$

$$\mathbf{G}_{DC} = \mathbf{F}^{-1}. \qquad (13)$$

From (12), the feedback filter $\mathbf{B}_{DC}$ is now upper triangular with zeros along the diagonal. Thus the computation of the elements of $\mathbf{v}$ is in the reverse order relative to CC systems, i.e., $v_K = b_K$ and $d_K = 0$ (the mod-2M operation is not required for the last user). For $i=K-1,K-2,\ldots,1$, $v_i = (b_i - \sum_{j=i+1}^{K} B_{ij}v_j)+2MA_id_i$. It can be shown that the feed-forward filter given by equation (13) does not increase the average transmit power and the power scaling factor is not required.

Substituting (12) and (13) into (11), we obtain $\mathbf{y}=\mathbf{b}+2M\mathbf{Ad}+\mathbf{z}_{DC}$. Mod-2M operations are then applied to eliminate the term $2M\mathbf{Ad}$. The noise component $\mathbf{z}_{DC}=diag(\mathbf{F})\mathbf{n}$ is whitened since its autocorrelation matrix is $N_0 diag(\mathbf{F})^{-2}$. THP-DC achieves the same ideal SER as THP-CC and DF-MUD (see equation (10)).



(a) Transmitter **(b) Receiver**

Fig. 2 Diagram of THP for a Decentralized-Channel (DC) System

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

In summary, THP-CC, THP-DC and DF-MUD have the same ideal performance provided that the effect of mod-2M operation in the THP schemes and the error propagation in DF-MUD are ignored. Since DF-MUD improves upon the linear decorrelating MUD for all but the first user [5], the proposed THP methods also have better performance than the linear decorrelating MUD. While this conclusion is valuable for CC systems, it is more meaningful to compare THP-DC with the linear transmitter precoding methods. For example, if equal transmit powers are employed for every user and all the transmit signals have the same cross-correlation, the linear decorrelating precoding yields the same performance as the linear decorrelating MUD, and is outperformed by THP. The advantage of THP over linear schemes will be further demonstrated in section IV.

3. THP for Flat Rayleigh Fading Channels

In this section, we extend the THP approach to flat Rayleigh fading channels. In the symbol interval of interest, the channel gain for the ith user is denoted as $c_i=\alpha_i e^{j\phi_i}$, 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 for K users as $\mathbf{C} = diag\{c_i\}_{K\times K}$.

For a DC system, 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-DC, the output vector of the MF bank equals

$$\mathbf{y} = \mathbf{CRv}+\mathbf{n} = \mathbf{C}diag(\mathbf{F})(\mathbf{b}+2M\mathbf{Ad})+\mathbf{n}. \qquad (14)$$

The received signals are detected coherently, scaled and passed through the mod-2M operators. The input of the detection device in the ith user receiver is $\alpha_i f_{ii} b_i+n_i$. The average SNR per bit at the input of the detection device γ_{bi} has the mean

$$\bar{\gamma}_{bi} = \frac{(M^2-1) f_{ii}^2 A_i^2}{6N_0 \log_2 M} E\{\alpha_i^2\}, \quad i = 1,2,...,K. \qquad (15)$$

The SER of the ith user is given by $Pe_i(\gamma_{bi})$ = $[2(M-1)/M]Q\left(\sqrt{6\log_2 M/(M^2-1)\gamma_{bi}}\right)$. The average SER can be expressed as [4]

$$Pe_i = \frac{M-1}{M}\left(1-\sqrt{\frac{3(\log_2 M)\bar{\gamma}_{bi}}{M^2-1+3(\log_2 M)\bar{\gamma}_{bi}}}\right). \qquad (16)$$

For the CC model over the Rayleigh fading channel, the received signal is $r(t) = \sum_{i=1}^{K} c_i v_i s_i(t)+n(t)$ The MF bank in the receiver is followed by the diagonal matrix $\mathbf{C}^H$, resulting in the output $\mathbf{y} = \mathbf{C}^H\mathbf{RCv} + \mathbf{C}^H\mathbf{n} = \mathbf{C}^H\mathbf{F}^T\mathbf{FCv} + \mathbf{C}^H\mathbf{n}$. Define $\hat{\mathbf{F}} = \mathbf{FC}$, which is also a lower triangular matrix, then the feedback and feed-forward filters are the same as (3) and (7) except that $\hat{\mathbf{F}}$ replaces $\mathbf{F}$. The output of the feed-forward filter is $\mathbf{b}+2M\mathbf{Ad}+\mathbf{z}_{CC}$, where $\mathbf{z}_{CC}=\mathbf{G}_{CC}\mathbf{C}^H\mathbf{n}$ is white noise vector with auto-correlation matrix $N_0 diag(\mathbf{F})^{-2}\mathbf{C}^{-2}$. Thus, the mean of average SNR per bit at the detector input and the average SER are given by (15) and (16), respectively. For transmitter precoding in fading channels, we assumed perfect knowledge of the channel coefficients at the transmitter. In practice, these coefficients need to be fed back and/or estimated. For rapidly varying fading channels, accurate long range fading prediction is also required [14]. The effect of quantized feedback, estimation and prediction errors need to be addressed in future performance investigation of various precoding methods.

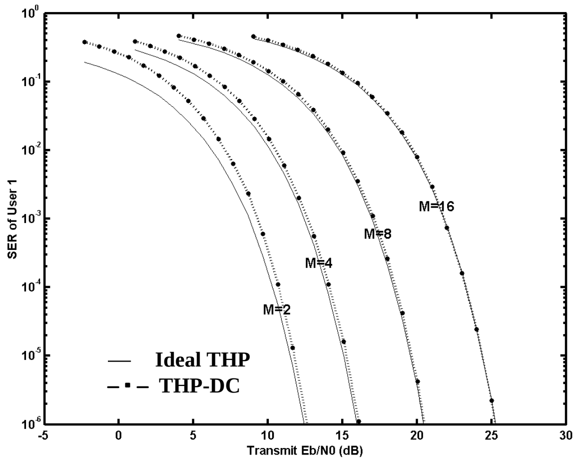


Fig. 3 Symbol error rate for user 1 of a 2-user system in AWGN channel, *M*-PAM, A_1=A_2, R_{12} = 0.5. (Example 1)

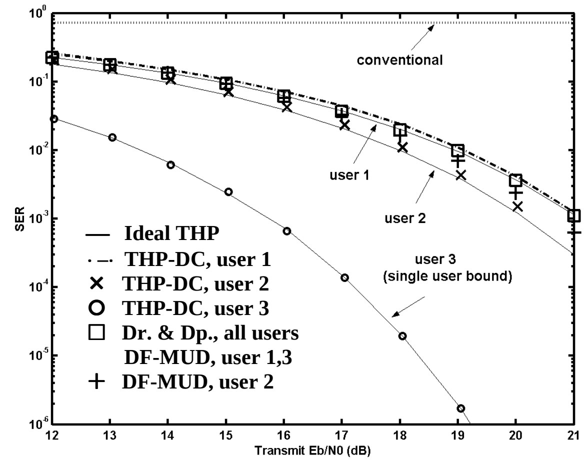


Fig.4 THP-DC, DF-MUD, decorrelating precoding (Dp.) and decorrelating MUD (Dr.) in AWGN channels, 3 users, 8-PAM, A_1=A_2=A_3, R_{12} = R_{13} = R_{23} = 0.8. (Example 2)

IV. NUMERICAL RESULTS AND ANALYSIS

Example 1 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 (both THP-DC and THP-CC), linear decorrelating MUD and linear decorrelating precoding are equal in this case. In Fig. 3, we demonstrate the difference between the simulation result for THP-DC and the theoretical SER. 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 $(-A_1\times M,\ A_1\times M]$ 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.

Example 2 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. Fig.4 shows the performance of four methods: THP-DC, linear decorrelating transmitter precoding (Dp.), DF-MUD, and linear decorrelating MUD Receiver (Dr.). The performance of THP-CC (not shown) is very close to that for THP-DC. The only difference is that the performance degradation due to the mod-2M operation affects users 1 through *K*-1 in the THP-DC system and users 2 through *K* in the THP-CC system. The SUB and the conventional system performance are also plotted 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-DC, the simulation results are very close to the theoretical performance since the end effect of mod-2M is slight for 8-PAM system. The best THP performance is for user 3. The THP achieves the SUB and has much lower error rate than DF-MUD and the 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-DC outperforms the linear methods and DF-MUD. Clearly, the THP schemes improve on the linear schemes for most users. Although THP approaches 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 error propagation when user powers are similar.

Example 3 We now investigate a 4-user, 16-PAM system in a flat fading channel. Assume the channel gains are i.i.d. Rayleigh fading, and the power of channel fading is normalized. Let the signal cross-correlation between any two users be 0.8, and A_1^2: A_2^2: A_3^2: A_4^2 = 8:4:2:1. Thus the first user is the strongest, and the last user is the weakest. The two linear decorrelating approaches have the same performance

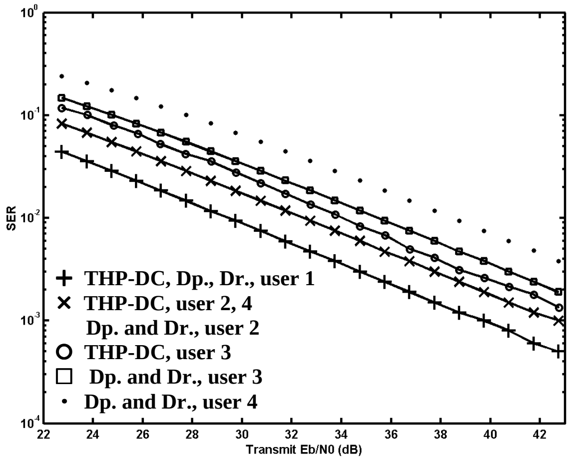


Fig. 5 THP, decorrelating precoding (Dp.) and decorrelating MUD (Dr.) in Rayleigh fading channel, 4 users, A_1^2: A_2^2: A_3^2: A_4^2 = 8:4:2:1, signal cross-correlation is 0.8. (Example 3)

under these given conditions. Since THP designs for CC and DC systems have similar performance, we only compare THP-DC with the linear decorrelating methods. As shown in Fig.5, user 1 achieves the best performance since its signal is the strongest. The three approaches result in similar SER for user 1 and user 2 as expected; while for the weakest users (user 3 and user 4) it is shown that THP-DC significantly outperforms the linear decorrelating schemes. The order of users presented here aids the weakest user. However, on the downlink, it might be desirable to order the users in the opposite order, so that a further user which needs larger transmit power is aided by the THP most. Thus, the overall average and peak transmit power can be reduced.

V. CONCLUSIONS AND FUTURE WORK

In this paper, we propose a transmitter-based nonlinear MAI rejection approach for synchronous channels with white Gaussian noise and flat Rayleigh fading. We describe two designs suitable for centralized and decentralized receiver systems, respectively. In particular, the THP-DC method significantly simplifies receiver structure by shifting signal processing to the common transmitter end. The ideal performance of the proposed THP techniques is the same as that of DF-MUD, while THP does not suffer from error propagation. It is also shown that THP schemes improve upon Tx-based and Rx-based linear decorrelating methods.

Future work includes applying THP to asynchronous and multipath channels, interference rejection for MIMO channels, and combining transmitter precoding and channel prediction techniques for fading channels.

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