LINEAR PRECODING VERSUS LINEAR MULTIUSER DETECTION IN THE DOWNLINK OF TDD-CDMA SYSTEMS

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ABSTRACT

In this paper, we compare two classes of linear interference suppression techniques for downlink TDD-CDMA systems, namely, linear multiuser detection methods (receiver processing) and linear precoding methods (transmitter processing). For the linear precoding schemes, we assume that the channel state information (CSI) is available only at the transmitter but not at the receiver. We propose several precoding techniques and the corresponding power control algorithms. The performance metric used in the comparisons is the total power required at the transmitter to achieve certain QoS at the receiver. Our results reveal that in general multiuser detection and precoding offer similar performance; but in certain scenarios, precoding can bring a substantial performance improvement. These results motivate the use of precoding techniques to reduce the complexity of the mobile terminals (only a matched-filter to the *own* spreading sequence is required without CSI).

1. INTRODUCTION

In the uplink CDMA wireless systems, it is assumed that the base station has access to all users' channel state information (CSI) and spreading signatures; and multiuser detection (MUD) has been shown to be an effective way to combat interference and increase data throughput [1]. For the downlink, on the other hand, one can transfer the signal processing for interference suppression from the mobile receiver to the transmit base station by using precoding techniques. This is feasible if the base station has access to the CSI of all active mobile units, e.g., in time-division duplex (TDD) systems where the base station can exploit the channel reciprocity if the time difference between uplink and downlink transmission is shorter than the channel coherence time, or by using channel prediction techniques. The simplest precoding method is pre-RAKE [2], which mitigates the multipath interference without considering the multiuser interference (MUI). Linear precoding techniques to remove the MUI and multipath interference were proposed in [3]. Nonlinear precoding techniques have been shown to offer superior performance although they complicate the receiver and the transmitter, since a modulo operation has to be implemented at both sides of the communication link [4, 5]. Note that most work on linear precoding assumes that each user implements a RAKE receiver and hence assumes the knowledge of CSI at the receiver [3, 5].

In this paper, we consider linear precoders with ultra-simple receivers, i.e., only a fixed matched-filter to the spreading sequence without CSI. We propose several linear precoders and the corresponding power control algorithms to meet certain performance at the receiver. We also consider the performance comparisons between linear precoding and linear MUD. The comparison metric is the total required power at the transmitter to achieve a minimum QoS requirement at each of the receivers. Our results show that linear precoding offers similar performance as linear MUD in most cases; but in some specific cases, linear precoding is more effective. These results motivate the use of linear precoding techniques in the downlink of TDD-CDMA systems. Among the advantages of using linear precoding we have:

- Receiver terminals are limited to a fixed matched-filter to the *own* spreading sequence. This translates into a power consumption reduction and decrease in price of the terminals since they do not have to perform sophisticated signal processing for channel estimation and interference mitigation. Note that variations in channel conditions and number of active users in the network do not affect the receiver operations.

- Less amount of control data is required in the precoding solution. The reason is that in MUD, every user requires to know the own channel response plus the spreading sequences of all other active users in the network. Moreover, mobile units do not need to be informed when users are added to (or removed from) the network.

- Power control is easy to implement with linear precoding since the receiver has information about the quality of each link and it does not require extra feedback information. Note that MUD requires a feedback link to find the power loading value assigned to each user.

- User scheduling based on the knowledge of CSI can be implemented jointly with linear precoding to increase the system throughput.

The remainder of this paper is organized as follows. In Section 2 we briefly summarize two well-known linear MUD methods and the corresponding power control algorithms. In Section 3 we propose several forms of linear precoding schemes and discuss their properties. In Section 4 we present simulation comparisons between linear MUD and linear precoding. Finally, Section 5 concludes the paper.

2. LINEAR MUD METHODS

We consider a K-user discrete-time synchronous multipath CDMA system. Define $b_k[i]$ from a constellation $\mathcal{A}$ as the symbol of the k-th user transmitted during the i-th symbol interval with $\mathbb{E}\{|b[i]|^2\} = 1$ and $\boldsymbol{b}[i] = [b_1[i], ..., b_K[i]]^T$. Denote N as the spreading factor and $\boldsymbol{s}_k = [s_{k,1}, ..., s_{k,N}]^T$ as the normalized spreading waveform of the k-th user. Then, the signal transmitted from the base station during the i-th symbol interval can be written as $\boldsymbol{p}[i] = \boldsymbol{S}\boldsymbol{A}\boldsymbol{b}[i]$, where $\boldsymbol{S} = [\boldsymbol{s}_1, \boldsymbol{s}_2, ..., \boldsymbol{s}_K]$ is the matrix of spreading waveforms; and $\boldsymbol{A} = \mathrm{diag}(A_1, ..., A_K)$ contains the user signal amplitudes. The vector $\boldsymbol{p}[i]$ is passed through a parallel-to-serial converter and transmitted over the multipath channel. The path delays are assumed to be an integral number of chip periods. Denote the multipath channel seen by the k-th user as $\boldsymbol{f}_k = [f_{k,1}, f_{k,2}, ..., f_{k,L}]^T$, where L is the number of resolvable paths and $f_{k,l}$ is the complex fading gain corresponding to the l-th path of the k-th user. We assume that $L < N$. At the k-th user's receiver, the $N \times 1$ received signal during N consecutive chip intervals corresponding to $\boldsymbol{b}[i]$ is given by

$$\boldsymbol{r}_k[i] = \underbrace{\boldsymbol{F}_k\boldsymbol{S}}_{\boldsymbol{H}_k}\boldsymbol{A}\boldsymbol{b}[i] + \boldsymbol{n}_k[i] \tag{1}$$

with

$$\boldsymbol{F}_k = \begin{bmatrix} f_{k,1} & 0 & \cdots & \cdots & 0 \\ \vdots & \ddots & \ddots & & \vdots \\ f_{k,L} & \ddots & f_{k,1} & \ddots & \\ 0 & \ddots & & \ddots & 0 \\ 0 & \cdots & f_{k,L} & \cdots & f_{k,1} \end{bmatrix}_{N\times N}, \tag{2}$$

where $\boldsymbol{r}_k[i] = [r_{k,1}[i], ..., r_{k,N}[i]]^T$ is the received signal, $\boldsymbol{n}_k[i] \sim \mathcal{N}_c\left(\boldsymbol{0}, \sigma^2\boldsymbol{I}_N\right)$ is the complex white Gaussian noise vector at the k-th receiver, and $\boldsymbol{H}_k = \boldsymbol{F}_k\boldsymbol{S}$. Notice that we have assumed that ISI can be ignored either by being truncated or by inserting a guard interval. At the k-th receiver, a linear detector to recuperate the signal $b_k[i]$ can be represented by an N-dimensional vector $\boldsymbol{w}_k \in \mathbb{C}^N$, which is correlated with the received signal $\boldsymbol{r}_k[i]$ in (1) to obtain $z_k[i] = \boldsymbol{w}_k^H\boldsymbol{r}_k[i]$, and the k-th mobile unit makes a decision $\hat{b}_k[i] = \mathcal{Q}(z_k[i])$, where $\mathcal{Q}$ rounds to the closest point in the constellation.

Linear Decorrelating Detector: The decorrelating detector completely eliminates the multiuser interference (MUI) and interchip interference (ICI), at the expense of enhancing the noise. The linear decorrelating detector for user k is given by [1]

$$\boldsymbol{w}_k = \boldsymbol{H}_k^{\dagger H}\boldsymbol{e}_k = \boldsymbol{H}_k(\boldsymbol{H}_k^H\boldsymbol{H}_k)^{-1}\boldsymbol{e}_k, \tag{3}$$

where $\boldsymbol{e}_k$ denotes a K-dimensional vector with all entries zeros, except for the k-th entry, which is 1. The output of this detector is given by

$$z_k[i] = \boldsymbol{w}_k^H\boldsymbol{r}_k[i] = A_kb_k[i] + \boldsymbol{w}_k^H\boldsymbol{n}_k[i] \tag{4}$$

and therefore

$$\mathrm{SINR}_k = \frac{A_k^2}{\sigma^2\|\boldsymbol{w}_k\|^2}, \tag{5}$$

where SINR_k is the signal-to-interference-plus-noise ratio for the k-th user. Suppose that the QoS requirement for user k is such that $\mathrm{SINR}_k \geq \gamma_k$, where γ_k is the minimum acceptable SINR value for user k. Hence we have $A_k^2 = \sigma^2\gamma_k\|\boldsymbol{w}_k\|^2$. And the total required transmit power is given by

$$P_T = \sum_{k=1}^{K} A_k^2 = \sum_{k=1}^{K} \sigma^2\gamma_k\,\boldsymbol{e}_k^H(\boldsymbol{S}^H\boldsymbol{F}_k^H\boldsymbol{F}_k\boldsymbol{S})^{-H}\boldsymbol{e}_k. \tag{6}$$

Linear MMSE Detector: The linear MMSE detector for user k is given by [1]

$$\begin{aligned} \boldsymbol{w}_k &= \arg\min_{\boldsymbol{w}_k\in\mathbb{C}^N} \mathbb{E}\left\{|b_k[i] - \boldsymbol{w}_k^H\boldsymbol{r}_k[i]|^2\right\} \\ &= A_k(\boldsymbol{H}_k\boldsymbol{A}^2\boldsymbol{H}_k^H + \sigma^2\boldsymbol{I}_N)^{-1}\boldsymbol{H}_k\boldsymbol{e}_k. \end{aligned} \tag{7}$$

The SINR for this detector is given by

$$\mathrm{SINR}_k = \frac{A_k^2\|\boldsymbol{w}_k^H\boldsymbol{H}_k\boldsymbol{e}_k\|^2}{\sum_{j\neq k} A_j^2\|\boldsymbol{w}_k^H\boldsymbol{H}_k\boldsymbol{e}_j\|^2 + \sigma^2\|\boldsymbol{w}_k\|^2}. \tag{8}$$

We seek to minimize the total power P_T such that $\mathrm{SINR}_k \geq \gamma_k$. The iterative power control algorithm for linear MMSE MUD proposed in [6] can be extended to the downlink scenario. At the $(n+1)$-th iteration, the MMSE filter $\boldsymbol{w}_k(n+1)$ is constructed using the current power matrix $\boldsymbol{A}(n)$. Then, the power matrix $\boldsymbol{A}(n+1)$ is updated using the new filter coefficients $\boldsymbol{w}_k(n+1)$.

3. LINEAR PRECODING SCHEMES

In this section we consider different approaches to implement linear precoding assuming that the transmitter has perfect CSI.

Algorithm 1 Power control algorithm for linear MMSE MUD in the downlink

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INPUT: H_k, γ_k, σ^2.
FOR n = 0,1,2,... DO
  FOR k = 1,2,...,K DO
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$$\begin{aligned} \boldsymbol{w}_k(n+1) &= (\boldsymbol{H}_k\boldsymbol{A}^2(n)\boldsymbol{H}_k^H+\sigma^2\boldsymbol{I})^{-1}A_k(n)\boldsymbol{H}_k\boldsymbol{e}_k \\ A_k^2(n+1) &= \gamma_k\frac{\sum_{j=1,j\neq k}^K A_j^2(n)\|\boldsymbol{w}_k^H(n+1)\boldsymbol{H}_k\boldsymbol{e}_j\|^2}{\|\boldsymbol{w}_k^H(n+1)\boldsymbol{H}_k\boldsymbol{e}_k\|^2} \\ &\quad +\gamma_k\frac{\sigma^2(\boldsymbol{w}_k^H(n+1)\boldsymbol{w}_k(n+1))}{\|\boldsymbol{w}_k^H(n+1)\boldsymbol{H}_k\boldsymbol{e}_k\|^2} \end{aligned} \quad (9)$$

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  END FOR;
END FOR;
OUTPUT: assigned powers A_k and linear
MMSE filters w_k, k = 1,...,K.
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3.1. Chip-wise Linear Precoding

We assume that each mobile unit employs only a filter matched to its *own* spreading sequence, and it does not need to know other users' spreading sequences or to estimate the channel.

Denote the symbol by symbol chip-wise precoding operation as $\boldsymbol{p}[i] = \boldsymbol{M}_c\boldsymbol{A}\boldsymbol{b}[i]$, where $\boldsymbol{p}[i]$ is the precoded symbol vector and $\boldsymbol{M}_c \in \mathbb{C}^{N\times K}$ is the chip-wise linear precoding matrix. Note that in chip-wise precoding, we do not explicitly use any spreading matrix at the transmitter. This is, the precoder takes K symbols and outputs the spread vector of length N. Hence the spreading and precoding operations are effectively combined. The vector $\boldsymbol{p}[i]$ is passed through a parallel-to-serial converter and transmitted through the channel. The received signal at the kth receiver is given by

$$\boldsymbol{r}_k[i] = \boldsymbol{F}_k\boldsymbol{A}\boldsymbol{M}_c\boldsymbol{b}[i] + \boldsymbol{n}_k[i], \quad (10)$$

where $\boldsymbol{M}_c \in \mathbb{C}^{N\times K}$ is the chip-wise precoding matrix. At each receiver k, the matched-filter $\boldsymbol{s}_k$ is applied to $\boldsymbol{r}_k[i]$. By stacking the outputs of all K matched-filters we obtain

$$\underbrace{\begin{bmatrix} \boldsymbol{s}_1^H\boldsymbol{r}_1[i] \\ \boldsymbol{s}_2^H\boldsymbol{r}_2[i] \\ \vdots \\ \boldsymbol{s}_K^H\boldsymbol{r}_K[i] \end{bmatrix}}_{\boldsymbol{y}[i]} = \underbrace{\begin{bmatrix} \boldsymbol{s}_1^H\boldsymbol{F}_1 \\ \boldsymbol{s}_2^H\boldsymbol{F}_2 \\ \vdots \\ \boldsymbol{s}_K^H\boldsymbol{F}_K \end{bmatrix}}_{\boldsymbol{H}_c}\boldsymbol{M}_c\boldsymbol{A}\boldsymbol{b}[i] + \underbrace{\begin{bmatrix} \boldsymbol{s}_1^H\boldsymbol{n}_1[i] \\ \boldsymbol{s}_2^H\boldsymbol{n}_2[i] \\ \vdots \\ \boldsymbol{s}_K^H\boldsymbol{n}_K[i] \end{bmatrix}}_{\boldsymbol{v}[i]}. \quad (11)$$

Here the channel matrix $\boldsymbol{H}_c$ has dimension $K \times N$ with $N \geq K$. The k-th receiver makes a decision $\hat{b}_k[i] = \mathcal{Q}(y_k[i])$. Therefore the precoder design problem involves designing the precoding matrix $\boldsymbol{M}_c$ such that $\boldsymbol{r}[i]$ is as close to $\boldsymbol{b}[i]$ as possible.

Chip-wise MMSE Precoding: The linear MMSE chip-wise precoder chooses the precoding matrix $\boldsymbol{M}_c$ to minimize $\mathbb{E}\{\|\boldsymbol{b}-\boldsymbol{y}\|^2\}$. Using an argument similar to [7], $\boldsymbol{M}_c$ is given by

$$\boldsymbol{M}_c = \boldsymbol{H}_c^\dagger = \boldsymbol{H}_c^H(\boldsymbol{H}_c\boldsymbol{H}_c^H)^{-1}. \quad (12)$$

It is easily seen that the SINR for each user is given by

$$\text{SINR}_k = \frac{A_k^2}{\sigma^2}, \;\; k = 1, ..., K. \quad (13)$$

If we assume that the required SINR for user k is γ_k, the required power assigned to the k-th user becomes $A_k^2 = \sigma^2\gamma_k$. Due to the precoding matrix, the required total transmit power becomes

$$P_T = \text{tr}(\boldsymbol{H}_c^\dagger\boldsymbol{A}^2\boldsymbol{H}_c^{\dagger H}) = \text{tr}(\boldsymbol{A}^2(\boldsymbol{H}_c\boldsymbol{H}_c^H)^{-1}). \quad (14)$$

Remark: Note that under a fixed transmit power budget P_T, the linear MMSE precoder is given by $\boldsymbol{M}_c = \beta\boldsymbol{H}_c^\dagger$ with $\beta = \sqrt{P_T/\text{tr}(\boldsymbol{A}^2(\boldsymbol{H}_c\boldsymbol{H}_c^H)^{-1})}$ and $\text{SINR}_k = \frac{(\beta A_k)^2}{\sigma^2}$.

Chip-wise Wiener Precoding: The Wiener precoder for multiple-antenna systems is proposed in [8] and can be used in our chip-wise system model. The Wiener precoder matrix $\boldsymbol{M}_c$ and constant β minimize $\mathbb{E}\left\{\|\boldsymbol{b}[i]-\beta^{-1}\boldsymbol{y}[i]\|^2\right\}$, subject to $\mathbb{E}\{\|\boldsymbol{M}_c\boldsymbol{A}\boldsymbol{b}[i]\|^2\} = P_T$. Given the total transmit power P_T, the Wiener precoder is given by

$$\boldsymbol{M}_c = \beta\boldsymbol{F}^{-1}\boldsymbol{H}_c^H, \;\text{ with }\; \beta = \sqrt{\frac{P_T}{\text{tr}\left(\boldsymbol{F}^{-2}\boldsymbol{H}_c^H\boldsymbol{A}^2\boldsymbol{H}_b\right)}} \quad (15)$$

and

$$\boldsymbol{F} = \boldsymbol{H}_c^H\boldsymbol{H}_c + \frac{K\sigma^2}{P_T}\boldsymbol{I}_N. \quad (16)$$

Next we propose a power loading algorithm for the chip-wise Wiener precoder. Consider the signal model (11). Define $\boldsymbol{G} = \boldsymbol{H}_c\boldsymbol{M}_c$. Then we can write $y_k[i] = A_k\boldsymbol{G}_{kk}b_k[i] + \sum_{i=1,i\neq k}^K A_i\boldsymbol{G}_{ki}b_i[i] + v_k[i], k = 1, ..., K$. In the Wiener precoder $\boldsymbol{M}_c$ is not the pseudo-inverse of $\boldsymbol{H}_c$ and therefore $\boldsymbol{G}$ is not a diagonal matrix. Hence, for a fixed loading matrix $\boldsymbol{A}$, the received SINR is given by

$$\text{SINR}_k = \frac{A_k^2\|\boldsymbol{G}_{kk}\|^2}{\sigma^2 + \sum_{i=1,i\neq k}^K A_i^2\|\boldsymbol{G}_{ki}\|^2}. \quad (17)$$

To achieve the target SINR γ_k for each user k, we need to find the optimal powers $A_k^2, k = 1, ..., K$. Now, different from the linear MMSE precoding, the power allocation problem is coupled with the problem of finding the optimal precoding matrix. Following the ideas of [6] we propose the following iterative algorithm to solve the joint problem. In the algorithm we first fix the power loading values $\boldsymbol{A}(n)$ to find the precoding matrix and then, based on the precoding matrix, the power loading values are updated. Simulations show that the algorithm converges in about two or three iterations.

Algorithm 2 Power control algorithm for Wiener precoder

INPUT: $\boldsymbol{H}_c, \sigma^2$ and $\gamma_k, k = 1, ..., K$;
FOR $n = 1, 2, ...$ DO
$\boldsymbol{F}(n+1) = \boldsymbol{H}_c^H \boldsymbol{H}_c + \frac{K\sigma^2}{P_T(n)} \boldsymbol{I}_N$
$\beta(n+1) = \sqrt{\frac{P_T(n)}{\mathrm{tr}(\boldsymbol{F}^{-2}(n+1)\boldsymbol{H}_c^H \boldsymbol{A}^2(n)\boldsymbol{H}_c)}}$
$\boldsymbol{M}_c(n+1) = \beta(n+1)\boldsymbol{F}^{-1}(n+1)\boldsymbol{H}_c^H$;
$\boldsymbol{G}(n+1) = \boldsymbol{H}_c \boldsymbol{M}_c(n+1)$;
FOR $k = 1 : K$ DO
$A_k^2(n+1) = \gamma_k \frac{\sum_{i=1, i\neq k}^{K} A_i^2(n) \|\boldsymbol{G}_{ki}(n+1)\|^2 + \sigma^2}{\|\boldsymbol{G}_{kk}(n+1)\|^2}$;
END;
$P_T(n+1) = E\{\|\boldsymbol{M}_c(n+1)\boldsymbol{A}(n+1)\boldsymbol{b}\|^2\} = \mathrm{tr}(\boldsymbol{M}_c(n+1)\boldsymbol{A}^2(n+1)\boldsymbol{M}_c^H(n+1))$;
END FOR;
OUTPUT: precoding matrix $\boldsymbol{M}_c(n+1)$, and assigned powers $\boldsymbol{A}(n+1)$

4. SIMULATION RESULTS

Linear precoding vs. linear MUD – total transmit power: We compare linear MUD with linear precoding. We assume that each mobile user experiences an independent multipath channel $\boldsymbol{f}_k = [f_{k,1}, ..., f_{k,L}]$ with $L = 3$ resolvable paths, and the transmitter has perfect CSI of all users. The path gains are generated according to $f_{k,i} \sim \mathcal{N}_c(0, \frac{1}{L})$. We compare the CDF of the required total power P_T at the transmitter to achieve a target SINR $\gamma_k = 13$dB, $\forall k$, in each of the four following schemes: (a) linear decorrelating MUD [cf. Eq.(6)]; (b) linear MMSE MUD [cf. Alg. 1]; (c) chip-wise linear MMSE precoder, [cf. Eq.(14)]; and (d) chip-wise Wiener precoder [cf. Alg. 2]. Simulations are performed for spreading gain $N = 31$, with Gold and random spreading sequences. Fig. 1 shows the results with $K = 15$ users and Fig. 2 shows the results with $K = 27$ users. It is seen that with Gold codes, MUD is slightly better (although only 0.5dB of difference with linear precoding when 15 users are considered), whereas with random codes linear precoding largely outperforms MUD. Notice that the Wiener precoder is slightly better than the MMSE precoder. It is also seen that the total power required in the precoding solutions is almost independent of the chosen spreading sequences and therefore, an outage event is less likely to occur. Although the linear MMSE MUD solution seems to be quite effective with Gold codes, we recall that it is unlikely to be implemented in the downlinks of most wireless systems due to the amount of required feedback information to implement perfect power control and other issues discussed in Section 1. Also notice that the linear decorrelator offers very poor performance in heavily loaded systems, which does not occur to the linear MMSE linear precoder.

Linear precoding vs. linear MUD – BER performance: Fig. 3 and Fig. 4 show the BER performance of the various linear MUD and linear precoding methods. We also consider the bit-wise linear MMSE precoding with a RAKE

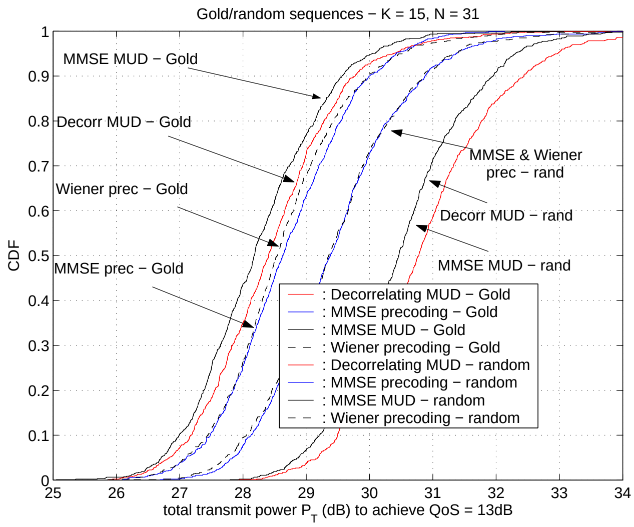

Fig. 1. Linear precoding vs. linear MUD: CDF of the required power P_T at the transmitter to achieve $\gamma_k = 13$dB, $\forall k$. Spreading gain $N = 31$, $K = 15$ users.

receiver proposed in [3]. The difference with the linear MMSE precoder considered in the Section 3.1 is that the receiver must also estimate the channel and apply a RAKE receiver, consequently, increasing the number of pilot symbols and the complexity of the receiver. We discuss this method only for comparison since we seek precoding solutions with simple receivers with no receiver CSI. The results are averaged over 100 channel realization and QPSK modulation is employed. Recall that the linear MMSE precoder is equivalent to the transmitter counterpart of the decorrelator. For the decorrelating MUD we consider perfect power loading to achieve the same SNR across the users. It is seen that the linear MMSE precoder with RAKE only performs slightly better with Gold sequences in the very low SNR region. In all the other cases, the chip-wise linear MMSE precoder obtains much better results. On the other hand, the chip-wise MMSE precoder obtains much better results than the decorrelating MUD, especially in heavily loaded systems. These results are due to the outage events of the decorrelating MUD observed in Fig. 1 and Fig. 2. Again, it is seen that the BER performance of the chip-wise precoding solution is almost independent of the chosen spreading sequence.

5. CONCLUSIONS

In this work, we have compared the performance of linear precoding and linear MUD in the downlink of TDD-CDMA systems. We have proposed different linear precoding schemes and power loading algorithms. Our results interestingly reveal that in general precoding can outperform the more complex MUD. These results strongly motivate the

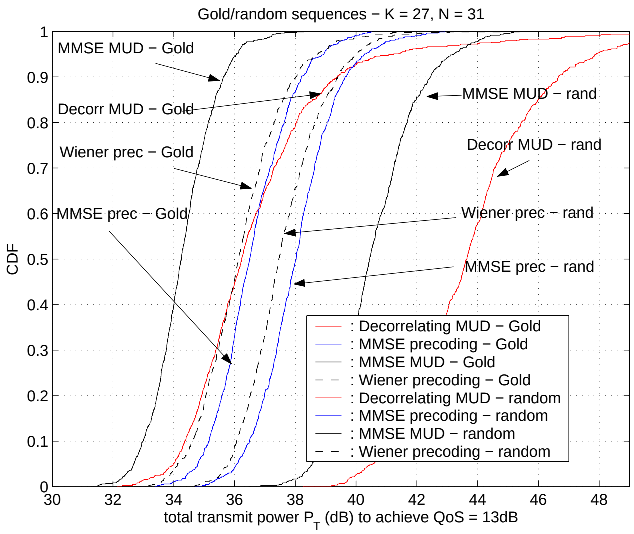

Fig. 2. Linear precoding vs. linear MUD: CDF of the required power P_T at the transmitter to achieve $\gamma_k = 10$dB, $\forall k$. Spreading gain $N = 31$, $K = 27$ users.

use of transmit precoding in the downlink of TDD-CDMA systems due to the multiple advantages over MUD, including the simple implementation of power control and user scheduling, reduction of pilot symbols, and the reduction of the power consumption and complexity at the mobile unit.

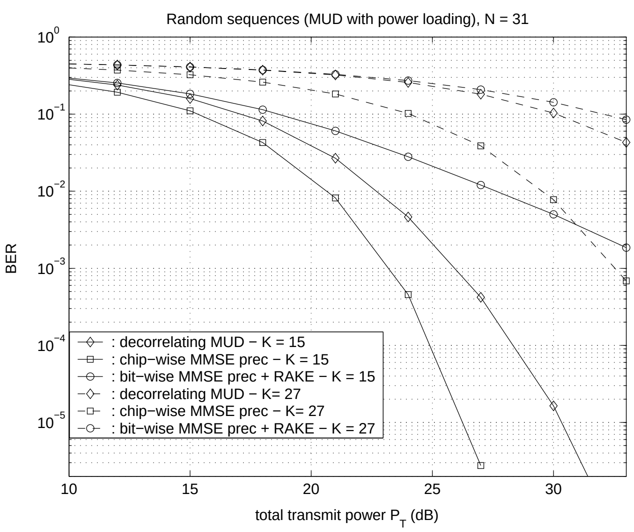

Fig. 3. Linear precoding vs. linear MUD: BER with random spreading sequences ($N = 31$, $K = 15$ and $K = 27$).

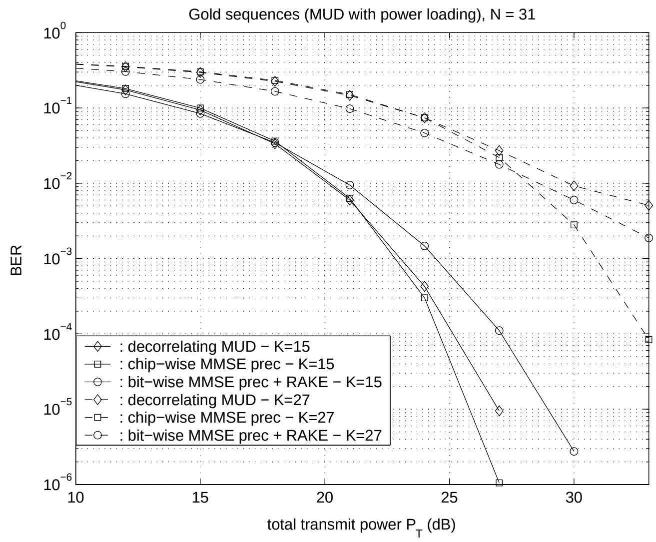

Fig. 4. Linear precoding vs. linear MUD: BER with Gold spreading sequences ($N = 31$, $K = 15$ and $K = 27$).

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