Transmit Processing Techniques Based on Switched Interleaving and Limited Feedback for Interference Mitigation in Multiantenna MC-CDMA Systems

Yunlong Cai, *Member, IEEE*, Rodrigo C. de Lamare, *Senior Member, IEEE*, and Didier Le Ruyet

***Abstract*—In this paper, we propose transmit processing techniques based on novel switched interleaving, chipwise linear precoding, and limited feedback for both downlink and uplink multicarrier code-division multiple access (MC-CDMA) multiple-antenna systems. We develop transceiver structures with switched interleaving, linear precoding, and detectors for both uplink and downlink using limited-feedback techniques. In the proposed schemes, a set of possible chip interleavers is constructed and prestored at both the base station (BS) and mobile stations (MSs). For the downlink, a new hybrid transmit processing technique based on switched interleaving and chipwise precoding is proposed to suppress the multiuser interference. The BS and MSs are also equipped with another codebook of quantized downlink channel-state information (CSI). Each MS quantizes its own downlink CSI and feeds the index back to the BS through a low-rate feedback channel. Then, the selection function at the BS determines the optimum interleaver based on the CSI of all users. Moreover, a transmit processing technique for the uplink of multiple-antenna MC-CDMA systems, which requires a very low rate of feedback information, is also proposed. Codebook design methods for both interleavers and quantized CSI are also proposed. Simulation results show that the performance of the proposed techniques is significantly better than prior art.**

***Index Terms*—Chip interleaving, interference suppression, limited feedback, multicarrier code-division multiple access (MC-CDMA), multiple-antenna systems, precoding.**

I. Introduction

FUTURE generations of broadband wireless systems are expected to support a wide range of services and bit rates by employing a variety of techniques that can achieve the highest possible spectrum efficiency [1]. Multicarrier code-division multiple access (MC-CDMA), which is a combination of orthogonal frequency-division multiplexing (OFDM) and code-division multiple access (CDMA), has been attracting much attention [2]. The benefits of MC-CDMA include high spectral efficiency, easy adaptation to severe channel conditions without complex detection, and robustness to intersymbol interference (ISI) and fading caused by multipath propagation [3]. There are several variations of MC-CDMA, e.g., multicarrier direct-sequence CDMA (MC-DS-CDMA) proposed by DaSilva and Sousa [4] and multitone CDMA proposed by Vandendorpe [5]. These signals can easily be transmitted and received using the fast Fourier transform (FFT) without increasing the transmitter and receiver complexities and have the attractive feature of high spectral efficiency. Giannakis *et al.* proposed the generalized multicarrier (GMC) CDMA system and afforded an all-digital unifying framework, which encompasses single-carrier (SC) and several MC-CDMA systems [6].

Moreover, transmit processing techniques that are implemented at the base station (BS) with multiple antennas have received wide attention, because they require a simple receiver at the mobile station (MS). This condition leaves the BS with the task of precoding the signals in view of suppressing the multiuser interference (MUI) and adapting to the propagation channels. The essential premise of using transmit processing techniques is the knowledge of the channel-state information (CSI) at the transmitter. In time-division duplexing (TDD) systems, CSI can be obtained at the BS by exploiting reciprocity between the forward and reverse links. In frequency-division duplexing (FDD) systems, reciprocity is usually not available, but the BS can obtain knowledge of the downlink user channels by allowing the users to send a small number of feedback bits on the uplink. The limited-feedback approach has widely been investigated in multiuser multiple-input–multiple-output (MIMO) systems [7]–[12]. In particular, [8], [9], and [12] allow users to quantize some function of downlink CSI and send this information to the BS. When the channel is quantized, the user signals cannot perfectly be orthogonalized due to inherent quantization errors, the residual interference affects the system performance considerably, and it also requires a significant amount of bits for satisfactory performance. Thus, novel schemes that require a low rate of feedback bits, which can give better performance, are needed.

To the best of our knowledge, in the literature, no preprocessing works have considered the signal interleaving in the time domain, together with spatial processing for MC-CDMA systems. In this paper, we propose, for the first time, novel transmit processing techniques based on a

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switched-interleaving algorithm for both downlink and uplink MC-CDMA multiple-antenna systems. In the proposed schemes, a set of chip interleavers are constructed and prestored at both the BS and MSs. During the transmission, the optimum interleaver is chosen by the selection function at the BS. For the downlink systems, the summary of the contributions is listed as follows.

- A new hybrid transmit processing technique based on switched-interleaving and chipwise precoding techniques that require the feedback quantized CSI of all users is proposed to suppress the MUI.
- The relevant linear minimum-mean-square error (MMSE) receiver is designed for symbol recovery at the MS [13].
- Taking into account the interleaving and transmit precoding, we compute the effective spreading codes and obtain the received signal-to-interference-plus-noise ratio (SINR) as the selection criterion for choosing the optimum interleaver.
- We show the ability of the proposed downlink algorithm to deal with the channel estimation error at the receiver.

Note that preprocessing techniques for the uplink transmission are relatively unheeded due to the limitation of MS. To this end, we also propose a transmit processing technique and a transceiver for the uplink of multiple-antenna MC-CDMA systems, which requires a very low rate of feedback information. In particular, we will show that the proposed uplink scheme is very efficient and simple to design. The contributions for the uplink are summarized as follows.

- Instead of sending the quantized CSI, the BS feeds the index that corresponds to the optimum interleaver back to the MSs. The users will send data by using the interleaver that corresponds to the index that was sent from the BS in this particular channel situation.
- A preprocessing technique based on the proposed switched-interleaving scheme with low-rate feedback information is developed.
- We also present a design algorithm for the selection functions and the MMSE receivers for the uplink.

In addition, several chip-interleaving codebook design methods and channel quantization schemes are proposed for the downlink and uplink. Note that the codebooks are designed offline and each user applies the same optimum interleaver to process the symbols. The simulations show that the performance of the proposed techniques is significantly better.

This paper is structured as follows. Sections II and III describe the system models, the precoders and MMSE receivers, and the selection of parameters and optimization for the proposed downlink and uplink schemes, respectively. Limited-feedback timing structures and techniques for designing codebooks are described in Section IV. The simulation results are presented in Section V. Section VI draws the conclusions.

The superscripts $(.)^T$, $(.)^*$, $(.)^{-1}$, $(.)^\dagger$, and $(.)^H$ denote the transpose, elementwise conjugate, matrix inverse, Moore–Penrose pseudoinverse, and Hermitian transpose, respectively. Bold symbols denote matrices or vectors. The symbols $\otimes$, $E[.]$, $|.|$, $\|.\|$, $\Re(.)$, $\mathrm{sgn}\{.\}$, $(.)!$, and $\mathbf{I}_M$ represent the Kronecker product, expectation operator, norm of a scalar, norm of a vector, selecting the real part, signum function, factorial operator, and $M \times M$ identity matrix, respectively. The operations $(:, y)$, $(x, :)$, and $(.)_x$ denote taking the yth column of a matrix, the xth row of a matrix, and the xth element of a vector, respectively. The operation $(x : y)$ denotes generating a new vector by taking the elements from the xth to the yth entries of a vector.

II. Downlink Transmission and Reception

In this paper, we assume that the systems have control and feedback channels that can provide the required information. In practice, most standards have such channels [14]–[16]. The proposed downlink limited-feedback-based MC-CDMA multi-antenna model is presented in Fig. 1, left-hand side, where the solid line represents the transmission link, and the dashed line represents the limited-feedback link. We consider that the BS is equipped with multiple antennas and the MS is equipped with a single antenna, because the MSs are less likely to be equipped with multiple antennas than the BS. All the MSs and the BS are equipped with the codebook of quantized CSI and the same codebook of chip interleavers. The proposed downlink system works as follows.

- Each user selects an index from the codebook of the quantized CSI based on the downlink channel estimation and relays it to the BS by a limited-feedback channel.
- The feedback quantized CSIs are employed to compute the chipwise transmit precoder and the selection function to calculate and choose the index of the optimum interleaver from the codebook of interleavers at the BS.
- The downlink data is operated by the optimum interleaver, and the BS broadcasts the index of the optimum interleaver to all the MSs prior to data transmission.

A. Proposed Downlink System Model and Chipwise Precoder

The downlink transceiver structure of the proposed scheme is presented on the right-hand side of Fig. 1. We consider an uncoded synchronous binary phase-shift keying (BPSK) MC-CDMA system with K users, N chips per symbol, and N_t transmit antennas, where the $N \times 1$ vectors $\mathbf{s}_1, \ldots, \mathbf{s}_K$ denote the spreading codes. We assume that each block contains M symbols and that $\mathbf{b}_k(i) = [b_1^{(k)}(i), \ldots, b_M^{(k)}(i)]^T$ denotes the ith block data for user k, $b_m^{(k)}(i) \in \{\pm 1\}$, $m = 1, \ldots, M$, $k = 1, \ldots, K$. Here, we drop the index i for notation simplicity. For each user, the chip interleaver permutes one block of chips per time. The permutated chips of user k before the precoding procedure are given by

$$\mathbf{x}_k^{(l)} = A_k \mathbf{P}_l \mathbf{S}_k \mathbf{b}_k \tag{1}$$

where the quantity A_k is the amplitude that is associated with user k. The matrix $\mathbf{P}_l$ denotes the lth $MN \times MN$ interleaving matrix that is designed by the interleaving patterns of the codebook, where $l = 1, \ldots, 2^B$, B is the number of bits to represent the index of the interleaver, and 2^B is the length of the interleaver codebook. The quantity $\mathbf{S}_k = \mathbf{s}_k \otimes \mathbf{I}_M$ is the

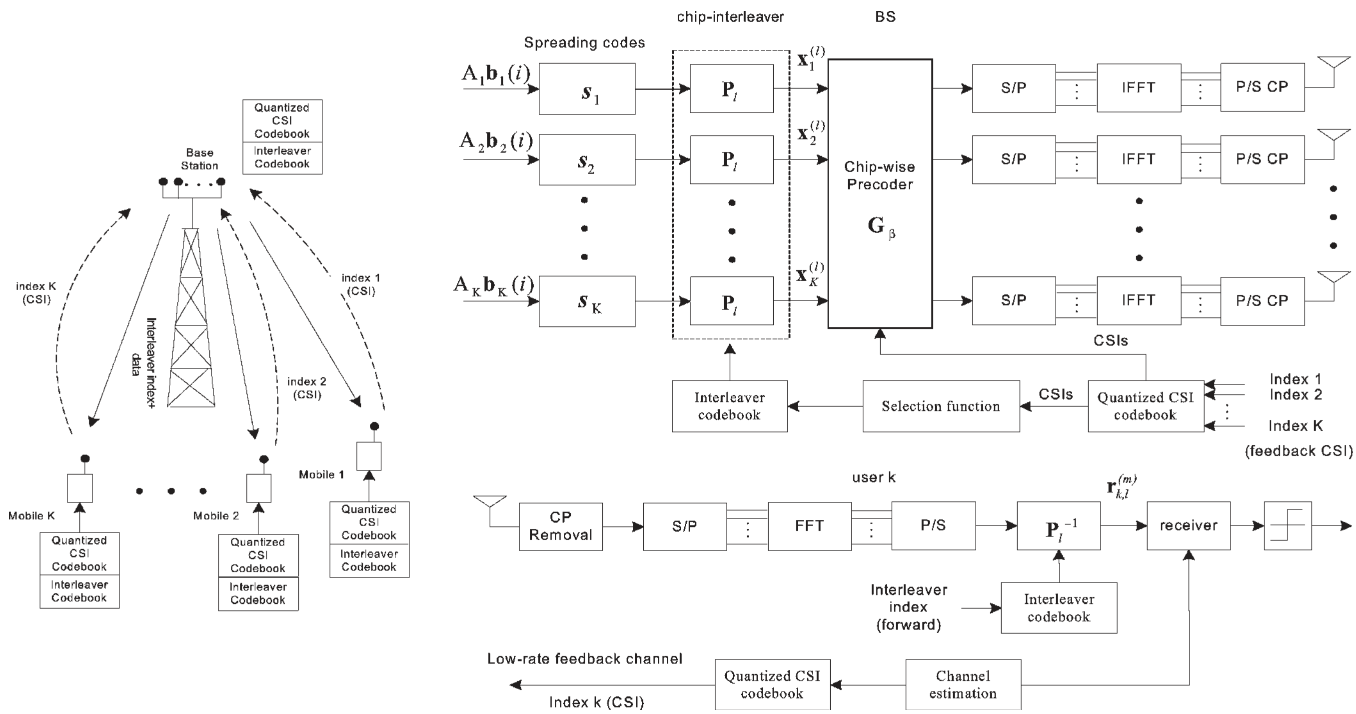


Fig. 1. Proposed downlink limited-feedback-based multiple-antenna MC-CDMA model and transceiver structure.

$MN \times M$ spreading code matrix. We define the $K \times MN$ matrix $\mathbf{X}^{(l)} = [\mathbf{x}_1^{(l)}, \ldots, \mathbf{x}_K^{(l)}]^T$, which consists of a block of permutated chips of all the users. Through the MN-point FFT and inverse fast Fourier transform (IFFT) and a cyclic prefix, the multipath fading channel can be divided into MN narrowband channels in the frequency domain [17], where we define a $K \times N_t$ matrix $\mathbf{D}_\beta$ as the equivalent channel matrix of the βth chip or subcarrier $\beta = 1, \ldots, MN$. The $K \times 1$ received vector $\mathbf{c}_{\beta,l}$ that consists of the βth received chips of all the users before deinterleaving denotes

$$\mathbf{c}_{\beta,l} = \mathbf{D}_\beta \mathbf{G}_\beta \mathbf{X}^{(l)}(:,\beta) + \tilde{\mathbf{n}}_\beta \tag{2}$$

where $\tilde{\mathbf{n}}_\beta$ denotes the $K \times 1$ noise vector, and $\mathbf{G}_\beta$ is the $N_t \times K$ precoding matrix of the βth chip. In this paper, the zero-forcing precoding is employed [7], which is given by

$$\mathbf{G}_\beta = \sqrt{\frac{U}{Tr\left(\mathbf{G}'_\beta \mathbf{G}_\beta^{H'}\right)}}\mathbf{G}'_\beta \tag{3}$$

where $\mathbf{G}'_\beta = \mathbf{D}_\beta^\dagger$, and U denotes the transmit power constraint $U = E[\|\mathbf{X}^{(l)}(:,\beta)\|^2]$.

B. Effective Spreading Code and Linear MMSE Receiver

We rewrite $\mathbf{G}_\beta$ as $\mathbf{G}_\beta = [\mathbf{g}_1^{(\beta)}, \ldots, \mathbf{g}_K^{(\beta)}]$, where $\mathbf{g}_k^{(\beta)}$ is the $N_t \times 1$ precoding vector for the kth user, and we define that the $MN \times 1$ vector $\bar{\mathbf{s}}_k$ is generated by stacking M copies of the kth user's spreading code vector on top of each other. The interleaved spreading code vector is given by $\mathbf{p}_{k,l} = \mathbf{P}_l \bar{\mathbf{s}}_k$. We obtain the $MN \times 1$ deinterleaved effective spreading code vector of the kth user for the lth interleaver that corresponds to the k_0th user's channel as

$$\tilde{\mathbf{s}}_{k,k_0,l} = \mathbf{P}_l^{-1} \begin{pmatrix} \mathbf{D}_1(k_0,:)\mathbf{g}_k^{(1)}(\mathbf{p}_{k,l})_1 \\ \vdots \\ \mathbf{D}_{MN}(k_0,:)\mathbf{g}_k^{(MN)}(\mathbf{p}_{k,l})_{MN} \end{pmatrix} \tag{4}$$

where $\mathbf{P}_l^{-1}$ denotes the deinterleaving matrix $k_0 = 1, \ldots, K$.

Then, we define two $K \times MN$ matrices $\mathbf{C}_l$ and $\mathbf{C}_l^r$, which are given by $\mathbf{C}_l = [\mathbf{c}_{1,l}, \ldots, \mathbf{c}_{MN,l}]$ and $\mathbf{C}_l^r = \mathbf{P}_l^{-1}\mathbf{C}_l^T$, and the received vector that corresponds to the mth symbol after deinterleaving for the desired user k_0 is given by

$$\begin{aligned} \mathbf{r}_{k_0,l}^{(m)} &= \mathbf{C}_l^r\left((m-1)N+1 : mN\right), k_0) \\ &= \sum_{k=1}^{K} A_k \tilde{\mathbf{s}}_{k,k_0,l}^{(m)} b_m^{(k)} + \bar{\mathbf{n}}_{m,k_0} \end{aligned} \tag{5}$$

where the $N \times 1$ vector $\tilde{\mathbf{s}}_{k,k_0,l}^{(m)} = \tilde{\mathbf{s}}_{k,k_0,l}((m-1)N+1 : mN)$ denotes the effective spreading code of the kth user that corresponds to the k_0th user's channel for the mth symbol, the vector $\bar{\mathbf{n}}_{m,k_0}$ is the $N \times 1$ deinterleaved noise vector, and $E[\bar{\mathbf{n}}_{m,k_0}\bar{\mathbf{n}}_{m,k_0}^H] = \sigma^2\mathbf{I}_N$. To obtain the MMSE receiver, we minimize

$$J_{MSE} = E\left[\left\|A_{k_0}b_m^{(k_0)} - \mathbf{w}_{m,k_0,l}^H \mathbf{r}_{k_0.l}^{(m)}\right\|^2\right]. \tag{6}$$

We take the gradient with respect to $\mathbf{w}_{m,k_0,l}^*$ and set it to zero. After further mathematical manipulations, we obtain the desired user's MMSE receiver of the lth interleaver for the mth symbol, which is given by

$$\mathbf{w}_{m,k_0,l} = \left(\sum_{k=1}^{K} A_k^2 \tilde{\mathbf{s}}_{k,k_0,l}^{(m)} \tilde{\mathbf{s}}_{k,k_0,l}^{(m)H} + \sigma^2 \mathbf{I}_N\right)^{-1} A_{k_0}^2 \tilde{\mathbf{s}}_{k_0,k_0,l}^{(m)}. \tag{7}$$

The complexity of the downlink chipwise precoder and the MMSE receiver is $O(N_t^3)$ and $O(N^3)$, respectively, due to the matrix inversion. To reduce the complexity, we can rely on advanced low-complexity parameter estimation algorithms [18]–[21].

C. Selection of Parameters and Optimization

By using the different interleaving patterns, we generate 2^B groups of effective spreading sequences with different cross correlations, which cause different levels of MUI. The BS transmitter is equipped with the selection function. Through the feedback quantized CSIs of all the users, the optimum interleaver index is chosen to maximize the sum received SINR over all the users in this codebook. The received SINR of the lth branch for the k_0th user is given in (8), which is computed as the ratio between the signals energy of user k_0 per block and the energy of interference plus noise in the same block, i.e.,

$$\begin{aligned}\mathrm{SINR}_l^{(k_0)} &= \frac{\sum_{m=1}^{M} E\left[\mathbf{w}_{m,k_0,l}^H A_{k_0}^2 \tilde{\mathbf{s}}_{k_0,k_0,l}^{(m)} \tilde{\mathbf{s}}_{k_0,k_0,l}^{(m)H} \mathbf{w}_{m,k_0,l}\right]}{\sum_{m=1}^{M} E\left[\mathbf{w}_{m,k_0,l}^H \mathbf{F}_{l,k_0} \mathbf{F}_{l,k_0}^H \mathbf{w}_{m,k_0,l}\right]} \\ &= \frac{\sum_{m=1}^{M}\left[\mathbf{w}_{m,k_0,l}^H \mathbf{R}_{s,k_0}^{(l)} \mathbf{w}_{m,k_0,l}\right]}{\sum_{m=1}^{M}\left[\mathbf{w}_{m,k_0,l}^H \mathbf{R}_{I,k_0}^{(l)} \mathbf{w}_{m,k_0,l}\right]}\end{aligned} \tag{8}$$

where the interference-plus-noise component $\mathbf{F}_{l,k_0} = \mathbf{r}_{k_0,l}^{(m)} - A_{k_0} \tilde{\mathbf{s}}_{k_0,k_0,l}^{(m)} b_m^{(k_0)}$, and the matrices $\mathbf{R}_{s,k_0}^{(l)} = A_{k_0}^2 \tilde{\mathbf{s}}_{k_0,k_0,l}^{(m)} \tilde{\mathbf{s}}_{k_0,k_0,l}^{(m)H}$, $\mathbf{R}_{I,k_0}^{(l)} = \mathbf{U}_{I,k_0}^{(l)} \mathbf{U}_{I,k_0}^{(l)H} + \sigma^2 \mathbf{I}$, and $\mathbf{U}_{I,k_0}^{(l)} = [A_1 \tilde{\mathbf{s}}_{1,k_0,l}^{(m)}, \ldots, A_{k_0-1} \tilde{\mathbf{s}}_{k_0-1,k_0,l}^{(m)}, A_{k_0+1} \tilde{\mathbf{s}}_{k_0+1,k_0,l}^{(m)}, \ldots, A_K \tilde{\mathbf{s}}_{K,k_0,l}^{(m)}]$, $l = 1, \ldots, 2^B$, $k_0 = 1, \ldots, K$. The optimum index l_{opt} for the downlink system maximizes the sum received SINR, as given by

$$l_{opt} = arg \max_{l=1\ldots 2^B} \left\{\sum_{k_0=1}^{K} \mathrm{SINR}_l^{(k_0)}\right\}. \tag{9}$$

The final output is given by

$$\hat{b}_{k_0,m}^{(f)} = \mathrm{sgn}\left\{\Re\left(\mathbf{w}_{m,k_0,l_{opt}}^H \mathbf{r}_{k_0,l_{opt}}^{(m)}\right)\right\} \tag{10}$$

where $\hat{b}_{k_0,m}^{(f)}$ is the mth estimated symbol within a block for the k_0th user.

The MMSE receiver $\mathbf{w}_{m,k_0,l}$ in (8) can be replaced by the effective spreading sequence $\tilde{\mathbf{s}}_{k_0,k_0,l}^{(m)}$ to reduce the complexity of the selection function; thus, (8) becomes

$$\begin{aligned}\mathrm{SINR}_l^{(k_0)} &= \frac{\sum_{m=1}^{M}\left[\tilde{\mathbf{s}}_{k_0,k_0,l}^{(m)H} \mathbf{R}_{s,k_0}^{(l)} \tilde{\mathbf{s}}_{k_0,k_0,l}^{(m)}\right]}{\sum_{m=1}^{M}\left[\tilde{\mathbf{s}}_{k_0,k_0,l}^{(m)H} \mathbf{R}_{I,k_0}^{(l)} \tilde{\mathbf{s}}_{k_0,k_0,l}^{(m)}\right]} \\ &= \frac{\sum_{m=1}^{M} A_{k_0}^2 \left\|\tilde{\mathbf{s}}_{k_0,k_0,l}^{(m)}\right\|^4}{\sum_{m=1}^{M}\sum_{k\neq k_0} A_k^2 \left|\rho_{k,k_0,l}^{(m)}\right|^2 + \sigma^2 \sum_{m=1}^{M}\left\|\tilde{\mathbf{s}}_{k_0,k_0,l}^{(m)}\right\|^2}\end{aligned} \tag{11}$$

where the cross correlation of the mth symbol is denoted by $\rho_{k,k_0,l}^{(m)} = \mathbf{s}_{k,k_0,l}^{(m)H} \mathbf{s}_{k_0,k_0,l}^{(m)}$, $k = 1, \ldots, K$, $k \neq k_0$. It is easy to show that (11) can be written as

$$\mathrm{SINR}_l^{(k_0)} = \frac{A_{k_0}^2 \left\|\tilde{\mathbf{s}}_{k_0,k_0,l}\right\|^4}{\sum_{k\neq k_0} A_k^2 \tilde{\mathbf{s}}_{k_0,k_0,l}^H \tilde{\mathbf{s}}_{k,k_0,l} + \sigma^2 \left\|\tilde{\mathbf{s}}_{k_0,k_0,l}\right\|^2}. \tag{12}$$

Note that the value of $\|\tilde{\mathbf{s}}_{k_0,k_0,l}\|^2$ is the same for all the interleavers. Thus, the received SINR of the k_0th user in this case changes as the cross-correlation summation $\sum_{k\neq k_0} \tilde{\mathbf{s}}_{k_0,k_0,l}^H \tilde{\mathbf{s}}_{k,k_0,l}$ building on the lth interleaver within one block. The optimum interleaver is chosen from the well-designed prestored interleaving codebook to maximize the received SINR for each fading block.

III. Uplink Transmission and Reception

The proposed uplink scheme and transceiver structure are presented in Fig. 2. Similar to the downlink, the BS with N_r receive antennas, and each MS with a single antenna is equipped with the same codebook of chip interleavers. Based on the estimated uplink CSI of each user, the BS feeds back an index that corresponds to the best available codebook entry to all the MSs, which select the same chip interleaver that corresponds to the feedback index to transmit signals.

A. Proposed Uplink System Model

Spatial processing techniques can be employed at the BS receiver to detect users' symbols [22]. The $MN \times 1$ received vector of the lth interleaver for the n_rth receive antenna after deinterleaving is given by

$$\bar{\mathbf{r}}_{n_r,l} = \sum_{k=1}^{K} \mathbf{P}_l^{-1} \mathbf{\Lambda}_{k,n_r} A_k \mathbf{P}_l \mathbf{S}_k \mathbf{b}_k + \mathbf{n}_{n_r} \tag{13}$$

where $n_r = 1, \ldots, N_r$, the quantity $\mathbf{n}_{n_r}$ is the $MN \times 1$ deinterleaved complex Gaussian noise vector of the n_rth receive antenna, and the matrix $\mathbf{\Lambda}_{k,n_r}$ denotes an $MN \times MN$ diagonal matrix, which is the equivalent frequency-domain uplink channel matrix between the kth user and the n_rth received antenna.

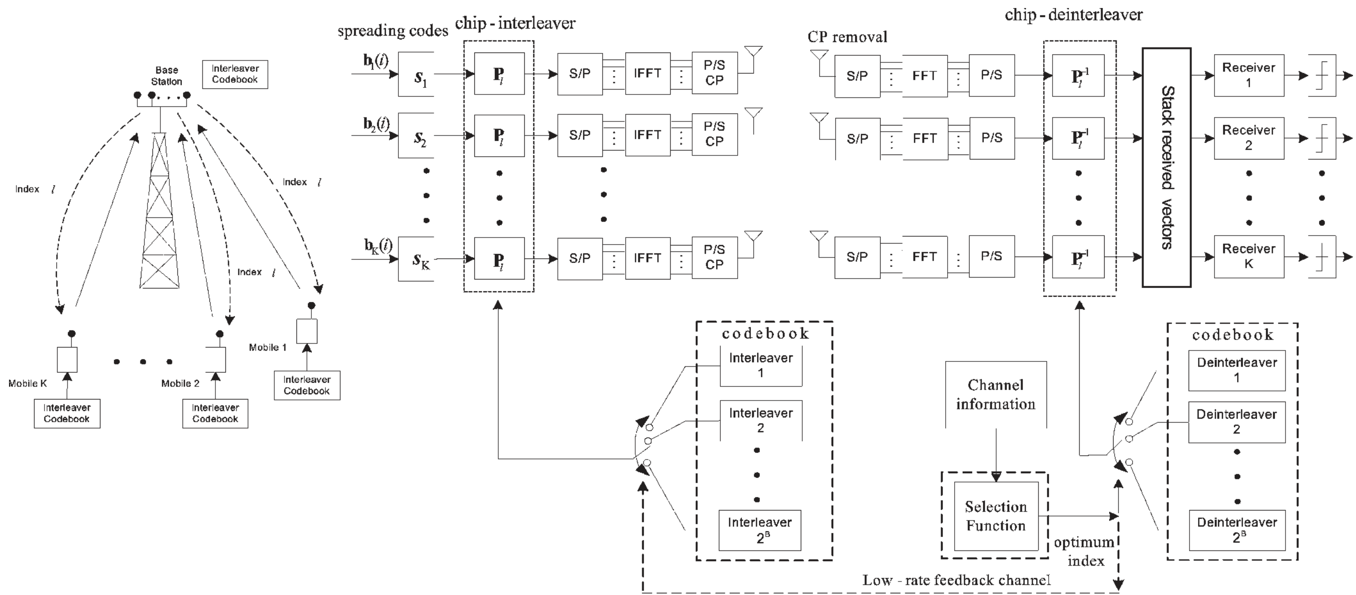


Fig. 2. Proposed uplink limited-feedback-based multiple-antenna MC-CDMA model and transceiver structure.

B. Linear MMSE Receiver

Due to the multiple antennas at the BS, we propose an MMSE receiver based on spatial–temporal processing techniques for the proposed uplink scheme. To reuse the space dimension, we stack the $N \times 1$ deinterleaved vectors $\bar{\mathbf{r}}_{n_r,l}((m-1)N+1:mN)$ from all the receive antenna elements on top of each other; thus, we obtain an $N_rN \times 1$ vector with regard to the mth symbol, i.e.,

$$\bar{\mathbf{r}}^s_{m,l} = \left[\bar{\mathbf{r}}^{(m)T}_{1,l}, \bar{\mathbf{r}}^{(m)T}_{2,l}, \ldots \bar{\mathbf{r}}^{(m)T}_{N_r,l}\right]^T \tag{14}$$

where the vector $\bar{\mathbf{r}}^{(m)}_{n_r,l} = \bar{\mathbf{r}}_{n_r,l}((m-1)N+1:mN)$. The final $MN \times 1$ effective spreading code vector of the lth interleaver for the kth user with regard to the n_rth receive antenna is given by

$$\tilde{\mathbf{p}}_{k,n_r,l} = \mathbf{P}_l^{-1}\boldsymbol{\Lambda}_{k,n_r}\mathbf{p}_{k,l}. \tag{15}$$

If we define the vector $\tilde{\mathbf{p}}^{(m)}_{k,n_r,l} = \tilde{\mathbf{p}}_{k,n_r,l}((m-1)N+1:mN)$, then the stacked received vector with regard to the mth symbol can be written as

$$\bar{\mathbf{r}}^s_{m,l} = \sum_{k=1}^{K} \tilde{\mathbf{p}}^s_{m,k,l} A_k b^{(k)}_m + \bar{\mathbf{n}}^s_m \tag{16}$$

where the $N_rN \times 1$ vector $\tilde{\mathbf{p}}^s_{m,k,l} = [\tilde{\mathbf{p}}^{(m)T}_{k,1,l}, \ldots, \tilde{\mathbf{p}}^{(m)T}_{k,N_r,l}]^T$, and $\bar{\mathbf{n}}^s_m$ is the $N_rN \times 1$ stacked noise vector $E[\bar{\mathbf{n}}^s_m \bar{\mathbf{n}}^{sH}_m] = \sigma^2\mathbf{I}_{N_rN}$. By following the same approach in the previous section, we obtain the uplink MMSE receiver expression of the lth interleaver for the k_0th user that corresponds to the mth symbol as

$$\bar{\mathbf{w}}_{m,k_0,l} = \left(\sum_{k=1}^{K} A_k^2 \tilde{\mathbf{p}}^s_{m,k,l}\tilde{\mathbf{p}}^{sH}_{m,k,l} + \sigma^2\mathbf{I}_{N_rN}\right)^{-1} A_{k_0}^2 \tilde{\mathbf{p}}^s_{m,k_0,l}. \tag{17}$$

The complexity of the uplink MMSE receiver is $O((N_rN)^3)$. Note that some other receivers also can be employed, e.g., adaptive linear receivers [23] and decision feedback receivers [24], [25].

C. Selection of Parameters and Optimization

For the uplink, the selection function is also equipped at the BS, which contains all the information of chip interleavers in this codebook, the estimated MSs' uplink channels, as well as the MMSE receivers. Similar to the downlink case, the received SINR expression for the lth interleaving pattern of the kth user has the same form as (8), where the vectors $\mathbf{w}_{m,k_0,l}$ and $\tilde{\mathbf{s}}^{(m)}_{k,k_0,l}$ are replaced by the uplink $N_tN \times 1$ receiver and the stacked effective spreading code $\tilde{\mathbf{p}}^s_{m,k,l}$, respectively. The expression of the uplink optimum index is the same as (9), and the final output for the k_0th user of the mth symbol is given by

$$\hat{b}^{(f)}_{k_0} = \mathrm{sgn}\left\{\Re\left(\bar{\mathbf{w}}^H_{m,k_0,l_{opt}}\bar{\mathbf{r}}^s_{m,l_{opt}}\right)\right\}. \tag{18}$$

IV. Design of Low-Rate Feedback Frame Structures and Codebooks

In this section, the frame structures of the proposed downlink and uplink low-rate feedback schemes are introduced. Then, the algorithms for designing codebooks for both quantized CSI and interleavers are described. To use a relatively small number of feedback bits, we focus on the time-domain CSI and separately quantize the channel direction and the channel norm. Then, we calculate the frequency-domain CSI based on the time-domain CSI. Note that the frequency-domain reduced-feedback quantization scheme in [26], [27] can also be employed.

A. Low-Rate Feedback Frame Structures

For both uplink and downlink schemes, preamble transmission and limited feedback are prior to payload transmission

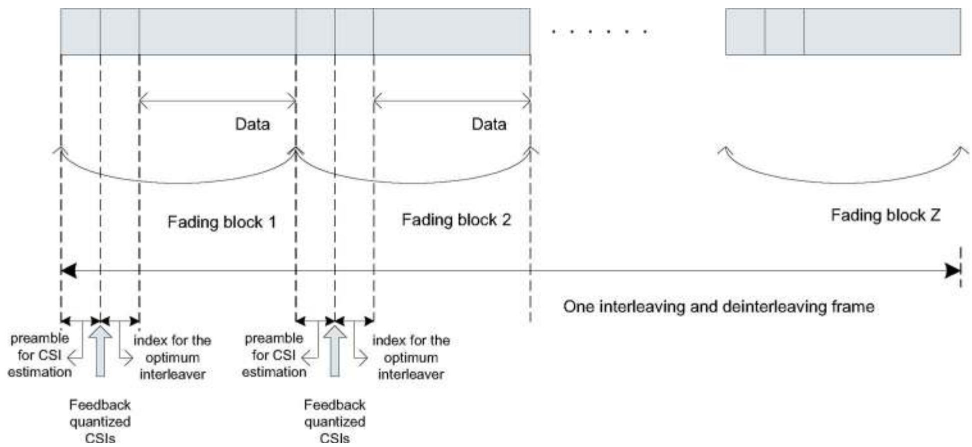


Fig. 3. Frame structure of the proposed downlink feedback scheme.

within each fading block. The CSI is estimated from the preamble at the receiver. Fig. 3 shows the frame structure of the proposed downlink feedback scheme. Each MS quantizes its own CSI and feeds back the index of the quantized CSI to the BS, which selects the optimum interleaver to preprocess the data. Payload transmission starts after the limited feedback, the first B bits that are broadcast are used to inform each MS the relevant optimum deinterleaver, and the rest of the payload is the preprocessed data. In this paper, we assume that the deinterleaver index is accurately known by the MSs. Because we consider that the channel is a block-fading channel, i.e., the channel coefficients can be treated as constants over one fading block, the feedback rate of the optimum index is once per fading block.

The proposed uplink scheme has a similar frame structure as the downlink scheme. However, the BS feeds back the index of the optimum interleaver to each MS. Thus, the uplink payload contains the preprocessed data only. We assume that the CSI is perfectly estimated at the receiver. Alternatively, we can employ a channel estimation algorithm that is trained with the preamble [34]. Furthermore, error-free transmission of feedback information is not possible if the feedback channel is noisy. In the next section, we will show the performance of the novel feedback schemes based on feedback channels with errors.

B. Design of Codebooks for Channel Direction and Norm Quantization

Because the channel is constant during a transmission block, to reduce the feedback bits for the downlink, we consider quantizing the multipath multiantenna fading channel in the time domain at the receiver. Let us write the channel matrix of the kth user as

$$\mathbf{H}_k = \begin{pmatrix} H^{(k)}(1,1) & H^{(k)}(1,2) & \dots & H^{(k)}(1,L_p) \\ H^{(k)}(2,1) & H^{(k)}(2,2) & \dots & H^{(k)}(2,L_p) \\ \vdots & \vdots & \vdots & \vdots \\ H^{(k)}(N_t,1) & H^{(k)}(N_t,2) & \dots & H^{(k)}(N_t,L_p) \end{pmatrix} \tag{19}$$

where L_p is the number of channel paths, and N_t denotes the number of transmit antennas at the BS. By using a vector quantization scheme [28], we can quantize the vectors of (19) per row or column, i.e., quantize the vector $\mathbf{H}_k(n_t,:)$ across the channel paths for each transmit antenna or the vector $\mathbf{H}_k(:,\alpha)$ across the transmit antennas for each channel path, where $n_t = 1,\dots,N_t$, and $\alpha = 1,\dots,L_p$. Define 2^{δ_1} and 2^{δ_2} as the direction codebook vector size and the norm codebook size, respectively. It is easy to see that the two quantization methods require $(\delta_1+\delta_2)N_t$ and $(\delta_1+\delta_2)L_p$ feedback bits, respectively. In the following discussion, we introduce the codebook design to quantize the vector across the transmit antennas, and the method is straightforward for the other method. Note that some similar methods in the time domain can be found in [33], [35], and [36], and some other quantization schemes can be applied in the frequency domain [26], [27].

The quantization of the channel direction information has been introduced by Narula *et al.* in [28], where the Lloyd algorithm was suggested for the design of the beamforming vector codebook. The authors in [29] and [30] showed that the codebook should be constructed by minimizing the maximum inner product between any two beamforming vectors in the codebook. We define the normalized channel vector, i.e., the channel direction as

$$\vec{\mathbf{h}} = \frac{\mathbf{h}}{\|\mathbf{h}\|} \tag{20}$$

where $\mathbf{h} = \mathbf{H}_k(:,\alpha)$, which denotes the $N_t \times 1$ channel vector across the transmit antennas per path, $\|\mathbf{h}\|$ is the channel norm, and $\vec{\mathbf{h}}$ is isotropically distributed in the N_t-dimension hypersphere of a unit radius. The receiver chooses the best quantized channel direction vector from a common codebook $\mathcal{T} = \{\vec{\mathbf{h}}_1^q,\dots,\vec{\mathbf{h}}_{2^{\delta_1}}^q\}$ in the maximum instantaneous correlation sense as

$$\vec{\mathbf{h}}_{opt}^q = \arg\max_{\vec{\mathbf{h}}_i^q \in \mathcal{T}} \left|\vec{\mathbf{h}}^H \vec{\mathbf{h}}_i^q\right|^2 \tag{21}$$

where $i = 1,\dots,2^{\delta_1}$.

An appropriate direction codebook is one that is designed to maximize the minimum chordal distance [29]. We have

$$\mathcal{T}_{opt} = \max_{\mathcal{T}\in\mathcal{C}^{N_t\times 2^{\delta_1}}} \min_{1\le i<j\le 2^{\delta_1}} d\left(\vec{\mathbf{h}}_i^q, \vec{\mathbf{h}}_j^q\right) \tag{22}$$

where

$$d\left(\vec{\mathbf{h}}_i^q, \vec{\mathbf{h}}_j^q\right) = \sqrt{1 - \left|\vec{\mathbf{h}}_i^{qH} \vec{\mathbf{h}}_j^q\right|^2} \tag{22}$$

and $\mathcal{C}^{N_t \times 2^{\delta_1}}$ denotes the $N_t \times 2^{\delta_1}$ complex matrix space. The Lloyd algorithm is described in the Appendix. To avoid the sum-rate degradation, we propose to use different direction codebooks for each user, and each user rotates the common codebook by a random unitary matrix to generate its own codebook [12]. With regard to the channel norm (scalar information), the algorithm is much simpler, and we use a nonuniform scalar quantizer.

C. Codebook of Interleavers

We assume that the lth interleaving order is $\phi_l(1), \phi_l(2), \ldots, \phi_l(MN)$, where $\phi_l(.)$ is defined as the order permutation function. For $\beta = 1, \ldots, MN$, the permutation matrix $\mathbf{P}_l$ is generated by filling the element of the βth row and the $\phi_l(\beta)$th column of an $MN \times MN$ zero matrix with 1. The βth row vector of the matrix $\mathbf{P}_l$ is given by

$$\mathbf{P}(\beta, :) = [\underbrace{0, \ldots, 0}_{\phi_l(\beta)-1}, 1, 0, \ldots, 0]. \tag{24}$$

The optimum interleaving codebook consists of $(MN)!$ interleavers. It is clearly impractical for any system when M and N are large numbers. Therefore, we need to select a subset of entries from the optimum codebook to build a practical suboptimum codebook that performs well. The first method is to randomly permutate the chips within one block, and we create the codebook by generating 2^B random-interleaving patterns. The second method is based on a block-interleaving method [31], where a block of chips comes into a matrix columnwise and goes out rowwise, and by varying the dimension of the matrix, we can obtain different interleaving patterns for the codebook entries. We assume that, for the lth interleaving pattern, the matrix dimension is $dx^{(l)} \times dy^{(l)}$, where $dx^{(l)} dy^{(l)} = MN$, and the permutation matrix $\mathbf{P}_l$ is given by

$$\mathbf{P}_l = \begin{pmatrix} \boldsymbol{\theta}_{1,l} \otimes \mathbf{I}_{dy^{(l)}} \\ \boldsymbol{\theta}_{2,l} \otimes \mathbf{I}_{dy^{(l)}} \\ \vdots \\ \boldsymbol{\theta}_{dx^{(l)},l} \otimes \mathbf{I}_{dy^{(l)}} \end{pmatrix} \tag{25}$$

where $\boldsymbol{\theta}_{\mu,l}$ is a $1 \times dx^{(l)}$ vector, $\mu = 1, \ldots, dx^{(l)}$, and

$$\boldsymbol{\theta}_{\mu,l} = [\underbrace{0, \ldots, 0}_{\mu-1}, 1, 0, \ldots, 0]. \tag{26}$$

We also propose a method for the interleaving codebook design. The basic principle is to build a codebook that contains the interleaving patterns with the maximum sum SINR (MASS). To implement the method, we need to conduct an extensive set of experiments and compute the sum SINRs for the indices of the random candidate patterns. The codebook is generated based on the statistics by choosing 2^B patterns with the maximum average sum SINR as the entries of the codebook. We define an $N_i \times N_e$ matrix $\mathbf{V}_{\mathrm{SINR}}$ as the storage of the sum SINRs for N_i possible interleavers over N_e testing channels, where N_i should be a large integer and practical for the experiment, N_e is the total number of experiments, and $\mathbf{V}_0$ is the list that contains all N_i interleaving patterns. The algorithm is summarized as follows.

Step 1. Initialization phase

- Initialize N_e and 2^B and choose an appropriate value for N_i. Set $\mathbf{v}_{idx}$, $\mathbf{V}_{\mathrm{SINR}}$, and $\mathbf{V}_{\mathrm{MASS}}$ with null.
- Generate N_i random-interleaving patterns, given the list of the interleavers to the matrix $\mathbf{V}_0$.

Step 2. Set $n_e = 1$.
Step 3. Set $l = 1$.
Step 4. Sum SINR calculation

- Generate the lth permutation matrix that corresponds to the lth entry in the interleaver list $\mathbf{V}_0$.
- The sum SINR of the lth interleaver entry is computed based on the permutation matrix $\mathbf{P}_l$, the n_eth testing channel matrix, and spreading sequences $\mathbf{s}_k$. Give it to the lth element of the column vector $\mathbf{V}_{\mathrm{SINR}}(:, n_e)$.

Step 5. $l \leftarrow l + 1$. *Loop back to Step 4 until* $l > N_i$.
Step 6. $n_e \leftarrow n_e + 1$. *Loop back to Step 3 until* $n_e > N_e$.
Step 7. Compute the average sum SINR.

- Based on the matrix $\mathbf{V}_{\mathrm{SINR}}$, by averaging the sum SINRs over the N_e testing channels, an $N_i \times 1$ vector $\mathbf{v}_{idx}$ is generated.

Step 8. Generate codebook.

- The final codebook $\mathbf{V}_{MASS}$ is generated by selecting 2^B patterns from $\mathbf{V}_0$ with the maximum average sum SINR according to $\mathbf{v}_{idx}$.

V. Simulations

In this section, we evaluate the performance of the proposed linear processing schemes with switched interleaving and compare them to other existing schemes, i.e., the conventional MC-CDMA system [2] and the MC-CDMA system that uses the chip-interleaving algorithm [31]. We adopt a simulation approach and conduct several experiments to verify the effectiveness of the proposed techniques. We carried out simulations to assess the average bit-error-rate (BER) performance of the interleaving algorithms for different loads, signal-to-noise ratios (SNR), the number of antennas, and the number of interleaving patterns. In this paper, our simulation results are based on an uncoded system with perfect CSI at the receiver. The length of the data block is set to $M = 8$ symbols, 128 subcarriers are used for each block, the random spreading code with a spreading gain $N = 16$ is generated for the simulations, and the length of the CP is enough to eliminate interblock interference. All channels have a profile with three taps whose powers are $p_0 = 0$ dB, $p_1 = -7$ dB, and $p_2 = -10$ dB, respectively, which are normalized, and the spacing between paths is $1/(MN)$ symbol duration. The sequence of channel coefficients is given by $h_l(i) = p_l \alpha_l(i)(l = 0, 1, 2)$, where $\alpha_l(i)$ are zero-mean circularly symmetric complex Gaussian random variables with unit variance. We have studied the proposed

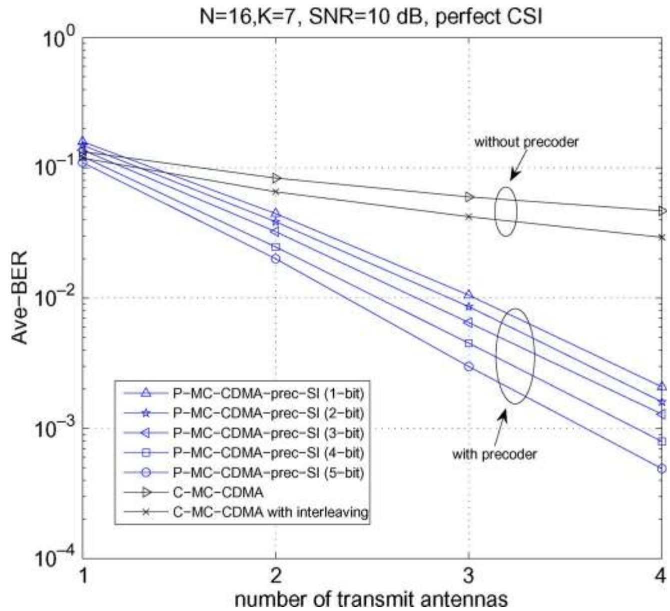


Fig. 4. Average BER performance versus the number of transmit antennas for the proposed downlink schemes with precoders and the conventional MC-CDMA schemes without precoders.

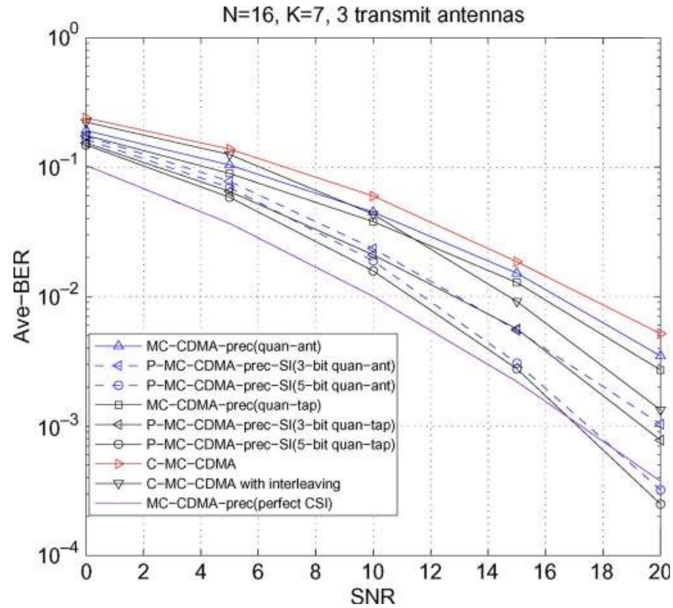


Fig. 5. Average BER performance versus SNR for the proposed downlink algorithms with different CSI quantization schemes and the conventional MC-CDMA systems.

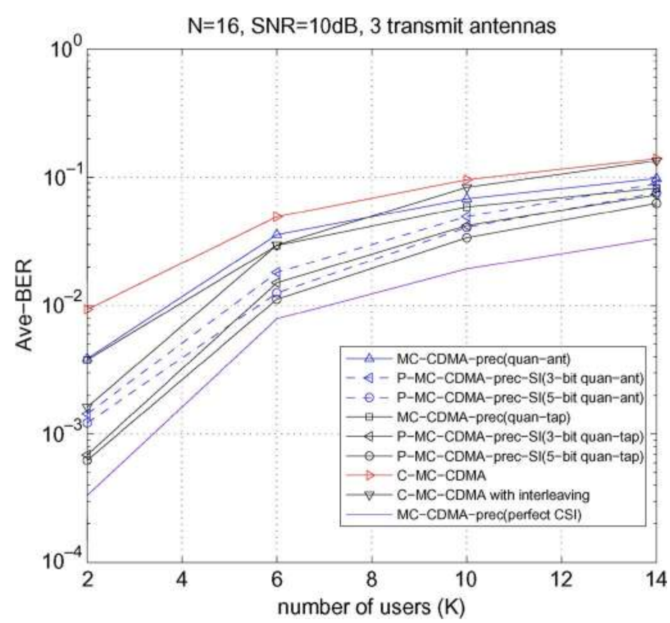


Fig. 6. Average BER performance versus the number of users for the proposed downlink algorithms with different CSI quantization schemes and the conventional MC-CDMA systems.

schemes with other channel profiles; however, we have opted for this approach to make the results easily reproducible. Note that the three-tap channel model has been used by previously reported works, e.g., [23]–[25]. Other channel models, e.g., the Universal Mobile Telecommunications System (UMTS)-Vehicular channels, can be also used; however, our studies indicate that the gains provided by the proposed schemes do not significantly change for the other channel models studies. Among the different schemes and quantization algorithms, we consider the following approaches:

- C-MC-CDMA: the conventional MC-CDMA system with the MMSE detector;
- P-MC-CDMA-prec-SI: the proposed preprocessing MC-CDMA system with switched interleaving and chipwise precoding schemes for the downlink;
- P-MC-CDMA-SI: the proposed preprocessing MC-CDMA system with switched-interleaving scheme for the uplink;
- MC-CDMA-prec: the conventional MC-CDMA system that employs the chipwise precoding scheme for the downlink;
- Perfect CSI: the perfect CSI at the transmitter;
- Quan-ant: vector quantization for the CSI across the channel paths per transmit antenna;
- Quan-tap: vector quantization for the CSI across the transmit antennas per channel path;
- B-bit: the proposed system employing B bits for the switched-interleaving scheme.

Fig. 4 shows the average BER performance of the proposed downlink switched-interleaving schemes combined with precoders and the conventional MC-CDMA schemes without precoders. In this experiment, we consider the scenario with a SNR of 10 dB and seven users. The knowledge of the CSI is given for the precoders at the transmitter, and the interleaving codebook is designed by using the random-interleaving method. The schemes with precoders are much better than the conventional MC-CDMA MMSE receivers. As the number of transmit antennas increases, the average BER decreases, and the gap between them becomes larger. Furthermore, the precoders with the proposed switched-interleaving scheme outperform the precoder without switched interleaving, and the performance improves as the number of the interleaving patterns increases.

The second experiment, which is shown in Figs. 5 and 6, considers the comparison in terms of the average BER of the time-domain CSI quantization scheme across the channel paths for each transmit antenna and across the transmit antennas for each channel path with the proposed downlink schemes. The interleaving codebook is designed by using the random-interleaving method. We use $\delta_1 = 10$ bits to feed back the channel direction and $\delta_2 = 6$ bits to feed back the channel norm. The Lloyd algorithm is used by these two CSI quantization methods. Note that, when we quantize the vector across the channel paths, a different codebook that is subject to the constraint of the normalized channel profile is designed. In particular, we show the average BER performance curves versus SNR and the number of users (K) for the analyzed schemes. The results in Fig. 5 indicate that, due to the large quantization error, the performance of the general precoding technique significantly decreases. The quantization scheme across the transmit antennas per tap is slightly better than the scheme across the taps per transmit antenna. In the case of three transmit antennas, the two quantization schemes both require 48 feedback bits for each user. The proposed downlink schemes outperform the general precoding algorithm without switched interleaving and the conventional MC-CDMA MMSE receiver with interleaving and without interleaving. In particular, the proposed downlink transmission scheme with 5 bits can save up to 5 dB compared with the general precoding algorithm without switched interleaving and can save up to more than 3 dB compared with the conventional MC-CDMA MMSE receiver with interleaving, at an average BER level of 10^{-2}. Fig. 6 indicates that the proposed scheme with 5 bits can support up to four more users at an average BER level of 10^{-2} compared with the conventional MC-CDMA MMSE receiver. As we increase the number of interleaving patterns, we achieve the performance of the general precoder with perfect CSI.

In the next experiment, we compare the codebooks of the interleavers that were created by the following three methods, as outlined in Section V: 1) the random-interleaving algorithm; 2) the block-interleaving algorithm; and 3) the MASS algorithm. In particular, we show the average BER performance curves versus the number of feedback bits for the uplink scenario. In this case, we consider $K = 5$, $\text{SNR} = 8$ dB, and 2 receive antennas at the BS. Note that the codebooks are designed offline. For the MASS algorithm, we set the number of simulations $N_e = 1000$ and the number of candidates $\beta = 100$ and 1000, and one block of symbols is transmitted per simulation. The results for an uplink system with $N = 16$ in the scenario of multipath fading channels are illustrated in Fig. 7. We can see that the best performance is achieved with the MASS algorithm, followed by the random- and block-interleaving methods. In particular, as we increase the number of candidates, the performance is improved for the MASS algorithm.

The results in Fig. 8 show the average BER performance versus SNR for the proposed downlink preprocessing scheme and the MC-CDMA system with transmit precoding using perfect and imperfect CSI at the receiver. Quantization across the transmit antennas for each channel path is employed. In the simulation, we assume that the imperfect channel coefficients are given by $\hat{h}_v^{(\alpha,k)}(i) \approx h_v^{(\alpha,k)}(i) + \varepsilon_v^{(\alpha,k)}(i)$, where $v = 1, \ldots, N_t$, $\alpha = 1, \ldots, L_p$, $k = 1, \ldots, K$, and $\varepsilon_v^{(\alpha,k)}(i)$ denotes a complex Gaussian random variable with zero mean and

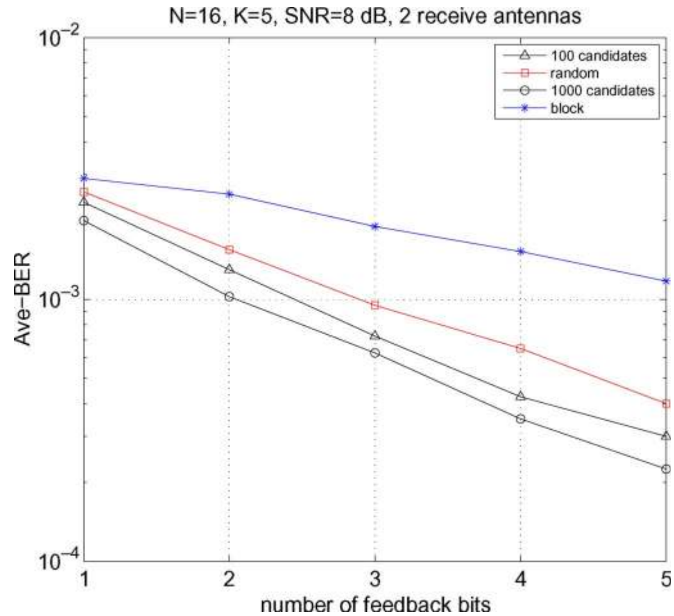


Fig. 7. Average BER performance versus the number of feedback bits for different interleaving codebooks. $N_e = 1000$. $\beta = 100$.

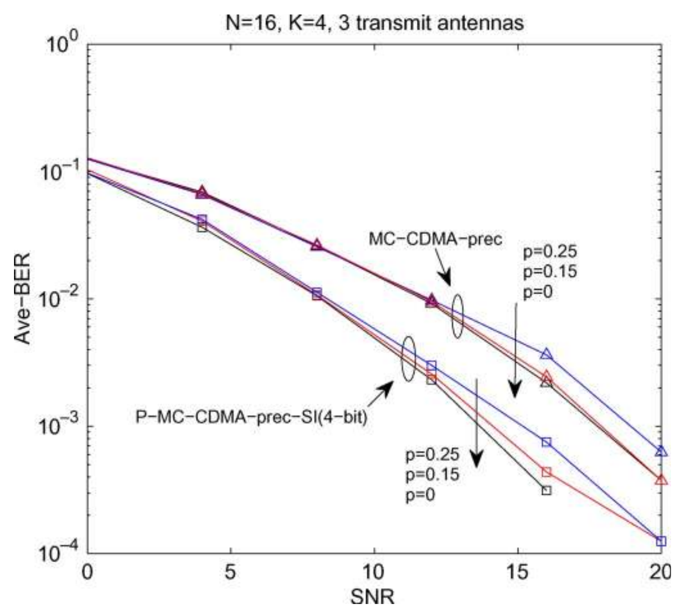


Fig. 8. Average BER performance versus SNR for the proposed downlink preprocessing scheme with perfect and imperfect CSI at the receiver. Quantization across the transmit antennas for each channel path is employed, where $\delta_1 = 10$, and $\delta_2 = 6$.

variance p^2. The imperfect channel vector of the αth path of the kth user is given by $\hat{\mathbf{h}}^{(\alpha,k)} = [\hat{h}_1^{(\alpha,k)}(i), \ldots, \hat{h}_{N_t}^{(\alpha,k)}(i)]^T$. Thus, the variance approaches the MSE of the channel estimation error at the receiver [32]. We compare the proposed downlink preprocessing scheme to the conventional MC-CDMA system with precoding based on the quantization errors with different p. Here, we select $p = 0, 0.15, 0.25$ and employ 10 feedback bits for the channel direction, six feedback bits for the channel norm, and four users and 16 interleavers for the proposed scheme. We can see that the performance decreases as the value p increases. In particular, the curves with $p = 0.15$ are close to the ones with perfect CSI at the receiver. The case with

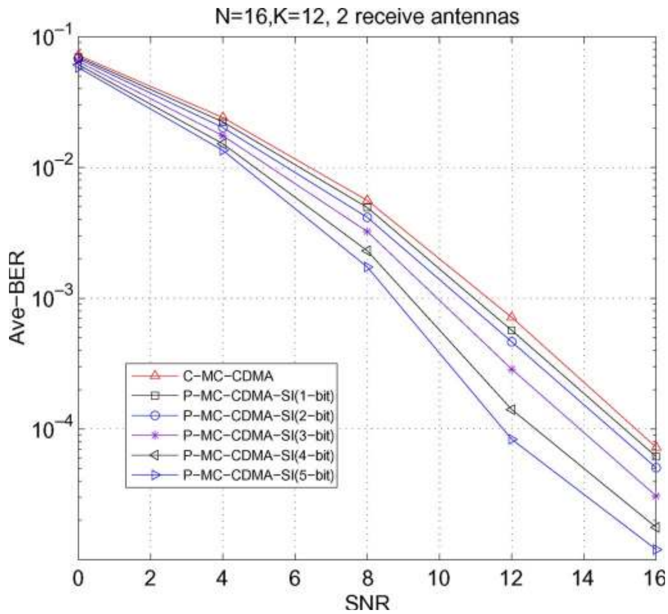


Fig. 9. Average BER performance versus SNR for the proposed uplink switched-interleaving preprocessing schemes and the conventional MC-CDMA MMSE receiver.

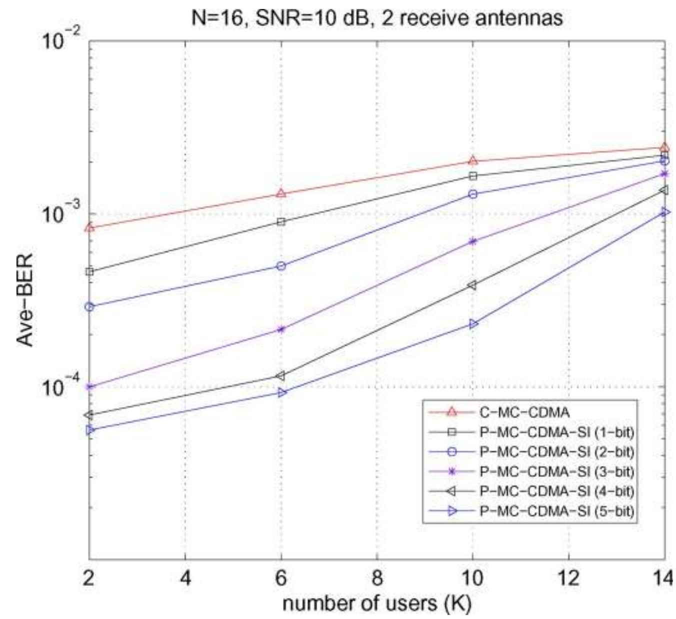


Fig. 10. Average BER performance versus the number of users for the proposed uplink switched-interleaving preprocessing schemes and the conventional MC-CDMA MMSE receiver.

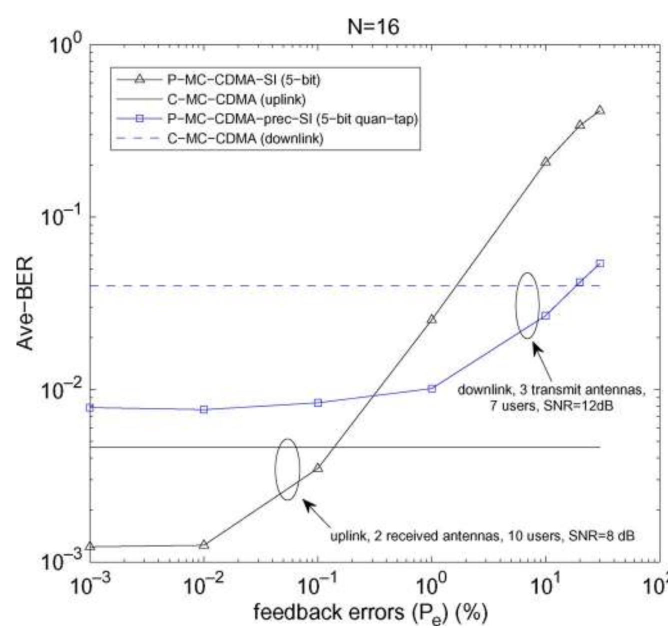


Fig. 11. Average BER performance versus the percentage of each user's feedback errors for the proposed downlink and uplink schemes.

$p = 0.25$ has almost 1.5-dB degradation, compared with the perfect CSI case at a BER level of 10^{-3}. This case shows the ability of our proposed downlink preprocessing algorithms to deal with channel estimation errors. We can relax the channel estimation algorithm at the receiver and reduce the computational complexity to meet the associated performance.

Now, let us consider the experiments of our proposed uplink preprocessing structure equipped with a different number of feedback bit configurations. We compare the performance in terms of the average BER of the proposed limited-feedback structures with MMSE receivers, i.e., 1-, 2-, 3-, 4-, and 5-bits feedback, respectively, and the conventional MC-CDMA system with MMSE receiver. In particular, we show the average BER performance curves versus the SNR and the number of users K for the analyzed schemes. In this simulation, the interleaving codebook is designed by the MASS algorithm. At the receiver, the BS is equipped with two antennas. The results in Figs. 9 and 10, which are based on $K = 12$ and $SNR = 10$ dB, respectively, indicate that the best performance is achieved with the novel switched-interleaving preprocessing scheme with five feedback bits, and we can see that the average BER decreases as the number of feedback bits increases. In particular, the proposed scheme with 5 bits can save up to 3 dB and support up to eight more users, which is near an average BER level of 10^{-3}, compared with the conventional MC-CDMA MMSE receiver.

Finally, Fig. 11 illustrates the average BER performance versus the percentage of each user's feedback errors for both uplink and downlink scenarios. Here, we use 5 bits for the proposed switched-interleaving schemes. The interleaving codebooks are based on the random-interleaving method. In particular, the downlink scheme quantizes the CSI across the transmit antennas per channel path. We use a structure based on a frame format where the indices are converted to 0 and 1 s. This frame of 1 and 0 s with the feedback information is transmitted over a binary symmetric channel with an associated probability of error Pe. The burst errors scenario in the limited-feedback channel can easily be transferred to the case of the binary symmetric channel by employing a conventional bit interleaver. As we increase the feedback errors for each user, the performance of the proposed limited-feedback schemes decreases. In the case of the downlink scheme, the performance decreases fast after 1%, and compared with the conventional MC-CDMA, it starts to lose at 20%. The performance of the uplink decays faster than the downlink scheme, because the interleaver is not in accordance with the deinterleaver due to the feedback errors for the uplink scenario, which creates a significant detection error for a fading block. To ensure that the errors are controlled, channel-coding techniques should be employed for the feedback channels with errors.

VI. Conclusion

In this paper, we have proposed linear preprocessing schemes based on switched-interleaving techniques with limited feedback for both downlink and uplink MC-CDMA systems. The chipwise linear precoder and relevant MMSE receivers were introduced. The selection functions were also proposed to choose the optimum interleaver from the codebook. The CSI quantization scheme based on the Lloyd algorithm and three methods for interleaving codebook design were described. The results showed that the proposed interleaving and detection schemes significantly outperform existing algorithms and support systems with higher loads. We remark that our proposed algorithms can also be extended to take into account coded, distributed, and other types of communication systems.

Appendix
Lloyd Algorithm for the Channel Direction Codebook Design

Step 1. Initialization phase.

- Generate a training sequence that consists of source vectors $\mathbf{h}$ with coefficients that are independent and identically distributed (i.i.d.) with a complex Gaussian distribution with zero mean and unit variance.

.

Step 2. Set $t = 1$.

Step 3. Nearest neighbor rule.

- All input vectors $\mathbf{h}$ that are closer to the codeword $\vec{\mathbf{h}}^q_{i,t-1}$ than any other codeword should be assigned to the neighborhood of $\vec{\mathbf{h}}^q_{i,t-1}$ or region Ω_i.
- $\mathbf{h} \in \Omega_i$ if and only if $d(\mathbf{h}, \vec{\mathbf{h}}^q_{i,t-1}) \leq d(\mathbf{h}, \vec{\mathbf{h}}^q_{j,t-1}), \forall i, j = 1, \ldots, 2^{\delta_1}$.

Step 4. Centroid condition.

- Take the ith region Ω_i as an example, whose local correlation matrix $\mathbf{\Sigma}_i := E[\mathbf{h}\mathbf{h}^H | \mathbf{h} \in \Omega_i]$.

  According to the centroid condition, the optimal vector $\vec{\mathbf{h}}^q_{i,t}$ should maximize $\boldsymbol{\omega}_i^H \mathbf{\Sigma}_i \boldsymbol{\omega}_i$ subject to the unit norm constraint, i.e.,

$$\vec{\mathbf{h}}^q_{i,t} = \underset{\boldsymbol{\omega}_i^H \boldsymbol{\omega}_i = 1}{\arg\max}\, \boldsymbol{\omega}_i^H \mathbf{\Sigma}_i \boldsymbol{\omega}_i = \mathbf{u}_i \tag{27}$$

  where $\mathbf{u}_i$ is the eigenvector that corresponds to the largest eigenvalue of $\mathbf{\Sigma}_i$.

Loop back to Step 3 until convergence.

References

[1] L. Hanzo, L.-L. Yang, E.-L. Kuan, and K. Yen, *Single- and Multi-carrier DS-CDMA: Multiuser Detection, Space–Time Spreading, Synchronization, Networking and Standards*. New York: IEEE Press-Wiley, Jun. 2003.

[2] S. Hara and R. Prasad, "Overview of multicarrier CDMA," *IEEE Commun. Mag.*, vol. 35, no. 12, pp. 126–133, Dec. 1997.

[3] K. Fazel and S. Kaiser, *Multicarrier and Spread Spectrum Systems*. Chichester, U.K.: Wiley, 2003.

[4] V. M. Dasilva and E. S. Sousa, "Performance of orthogonal CDMA codes for quasisynchronous communication systems," in *Proc. IEEE ICUPC*, Ottawa, ON, Canada, Oct. 1993, pp. 995–999.

[5] L. Vandendorpe, "Multitone direct sequence CDMA system in an indoor wireless environment," in *Proc. IEEE 1st Symp. Commun. Veh. Technol. Benelux*, Delft, The Netherlands, Oct. 1993, pp. 4.1.1–4.1.8.

[6] G. B. Giannakis, P. A. Anghel, and Z. Wang, "Generalized multicarrier CDMA: Unification and linear equalization," *EURASIP J. Appl. Signal Process.*, vol. 2005, no. 5, pp. 743–756, 2005.

[7] A. D. Dabbagh and D. J. Love, "Multiple-antenna MMSE-based downlink precoding with quantized feedback or channel mismatch," *IEEE Trans. Commun.*, vol. 56, no. 11, pp. 1859–1868, Nov. 2008.

[8] N. Jindal, "MIMO broadcast channels with finite-rate feedback," *IEEE Trans. Inf. Theory*, vol. 52, no. 11, pp. 5045–5060, Nov. 2006.

[9] C. Swannack, G. W. Wornell, and E. Uysal-Biyikoglu, "Efficient quantization for feedback in MIMO broadcasting systems," in *Proc. IEEE Asilomar Conf. Signals, Syst., Comput.*, Oct./Nov. 2006, pp. 784–788.

[10] M. Trivellato, F. Boccardi, and F. Tosato, "User selection schemes for MIMO broadcast channels with limited feedback," in *Proc. IEEE Veh. Technol. Conf.*, Apr. 2007, pp. 2089–2093.

[11] T. Yoo, N. Jindal, and A. Goldsmith, "Multiantenna downlink channels with limited feedback and user selection," *IEEE J. Sel. Areas Commun.*, vol. 25, no. 7, pp. 1478–1491, Sep. 2007.

[12] P. Ding, D. J. Love, and M. D. Zoltowski, "Multiple-antenna broadcast channels with shape feedback and limited feedback," *IEEE Trans. Signal Process.*, vol. 55, no. 7, pp. 3417–3428, Jul. 2007.

[13] Y. Cai, R. C. de Lamare, and D. Le Ruyet, "Linear precoding based on switched interleaving and limited feedback for interference suppression in downlink multiantenna MC-CDMA systems," in *Proc. IEEE Int. Conf. Acoust., Speech, Signal Process.*, Dallas, TX, Mar. 2010, pp. 3322–3325.

[14] R. T. Derryberry, S. D. Gray, D. M. Ionescu, G. Mandyam, and B. Raghothaman, "Transmit diversity in 3G CDMA systems," *IEEE Commun. Mag.*, vol. 40, no. 4, pp. 68–75, Apr. 2002.

[15] H. Holma and A. Toskala, *WCDMA for UMTS: Radio Access for Third-Generation Mobile Communications*. Chichester, U.K.: Wiley, 2002.

[16] C. R. Murthy, J. Zheng, and B. D. Rao, "Performance of quantized methods for equal gain transmission with noisy feedback channels," *IEEE Trans. Signal Process.*, vol. 56, no. 6, pp. 2451–2460, Jun. 2008.

[17] Z. Wang and G. B. Giannakis, "Wireless multicarrier communications," *IEEE Signal Process. Mag.*, vol. 17, no. 3, pp. 29–48, May 2000.

[18] R. C. de Lamare and R. Sampaio-Neto, "Reduced-rank adaptive filtering based on joint iterative optimization of adaptive filters," *IEEE Signal Process. Lett.*, vol. 14, no. 12, pp. 980–983, Dec. 2007.

[19] R. C. de Lamare and R. Sampaio-Neto, "Adaptive reduced-rank processing based on joint and iterative interpolation, decimation and filtering," *IEEE Trans. Signal Process.*, vol. 57, no. 7, pp. 2503–2514, Jul. 2009.

[20] R. C. de Lamare and R. Sampaio-Neto, "Reduced-rank space–time-adaptive interference suppression with joint iterative least squares algorithms for spread-spectrum systems," *IEEE Trans. Veh. Technol.*, vol. 59, no. 3, pp. 1217–1228, Mar. 2010.

[21] Y. Cai and R. C. de Lamare, "Low-complexity variable-step-size mechanism for code-constrained constant modulus stochastic gradient algorithms applied to CDMA interference suppression," *IEEE Trans. Signal Process.*, vol. 57, no. 1, pp. 313–323, Jan. 2009.

[22] M. Jankiraman, *Space–Time Codes and MIMO Systems*. Norwood, MA: Artech House, 2004.

[23] R. C. de Lamare and R. Sampaio-Neto, "Adaptive interference suppression for DS-CDMA systems based on interpolated FIR filters with adaptive interpolators in multipath channels," *IEEE Trans. Veh. Technol.*, vol. 56, no. 5, pp. 2457–2474, Sep. 2007.

[24] R. C. de Lamare and R. Sampaio-Neto, "Minimum mean-square-error iterative successive parallel arbitrated decision feedback detectors for DS-CDMA systems," *IEEE Trans. Commun.*, vol. 56, no. 5, pp. 778–789, May 2008.

[25] Y. Cai and R. C. de Lamare, "Space–Time adaptive MMSE multiuser decision feedback detectors with multiple-feedback interference cancellation for CDMA systems," *IEEE Trans. Veh. Technol.*, vol. 58, no. 8, pp. 4129–4140, Oct. 2009.

[26] T. Pande, D. J. Love, and J. V. Krogmeier, "Reduced feedback MIMO–OFDM precoding and antenna selection," *IEEE Trans. Signal Process.*, vol. 55, no. 5, pp. 2284–2293, May 2007.

[27] J. Choi and R. W. Heath, Jr., "Interpolation-based transmit beamforming for MIMO–OFDM with limited feedback," *IEEE Trans. Signal Process.*, vol. 53, no. 11, pp. 4125–4135, Nov. 2005.

[28] A. Narula, M. J. Lopez, M. D. Trott, and G. W. Wornell, "Efficient use of side information in multiple-antenna data transmission over fading channels," *IEEE J. Sel. Areas Commun.*, vol. 16, no. 8, pp. 1423–1436, Oct. 1998.

[29] D. J. Love, R. W. Heath, Jr., and T. Strohmer, "Grassmanian beamforming for multiple-input–multiple-output wireless systems," *IEEE Trans. Inf. Theory*, vol. 49, no. 10, pp. 2735–2747, Oct. 2003.

[30] K. K. Mukkavilli, A. Sabarwal, E. Erkip, and B. Aazhang, "On beamforming with finite-rate feedback in multiple-antenna systems," *IEEE Trans. Inf. Theory*, vol. 49, no. 10, pp. 2562–2579, Oct. 2003.

[31] Y. Na, M. Saquib, and M. Z. Win, "Pilot-aided chip-interleaved DS-CDMA transmission over time-varying channels," *IEEE J. Sel. Areas Commun.*, vol. 24, no. 1, pp. 151–160, Jan. 2006.

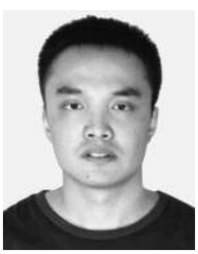
[32] N. S. Jayant and P. Noll, *Digital Coding of Waveforms: Principles and Applications to Speech and Video*. Englewood Cliffs, NJ: Prentice-Hall, 1984.

[33] D. J. Love, R. W. Heath, Jr., W. Santipach, and M. L. Honig, "What is the value of limited feedback for MIMO channels?" *IEEE Commun. Mag.*, vol. 42, no. 10, pp. 54–59, Oct. 2004.

[34] Y. Cai, R. C. de Lamare, and R. Fa, "Linear interference suppression for spread spectrum systems with switched interleaving and limited feedback," in *Proc. Wireless Commun. Netw. Conf.*, Budapest, Hungary, Sep. 2009, pp. 1–6.

[35] H. Shirani-Mehr and G. Caire, "Channel-state feedback schemes for multiuser MIMO–OFDM downlink," *IEEE Trans. Commun.*, vol. 57, no. 9, pp. 2713–2723, Sep. 2009.

[36] Y. Liang, R. Schober, and W. Gerstacker, "Time-domain transmit beamforming for MIMO–OFDM systems," in *Proc. GLOBECOM*, 2007, pp. 3579–3584.



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