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Space-Time Chip Equalization for Maximum Diversity Space-Time Block Coded DS-CDMA Downlink Transmission

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In the downlink of DS-CDMA, frequency-selectivity destroys the orthogonality of the user signals and introduces multiuser interference (MUI). Space-time chip equalization is an efficient tool to restore the orthogonality of the user signals and suppress the MUI. Furthermore, multiple-input multiple-output (MIMO) communication techniques can result in a significant increase in capacity. This paper focuses on space-time block coding (STBC) techniques, and aims at combining STBC techniques with the original single-antenna DS-CDMA downlink scheme. This results into the so-called space-time block coded DS-CDMA downlink schemes, many of which have been presented in the past. We focus on a new scheme that enables both the maximum multiantenna diversity and the maximum multipath diversity. Although this maximum diversity can only be collected by maximum likelihood (ML) detection, we pursue suboptimal detection by means of space-time chip equalization, which lowers the computational complexity significantly. To design the space-time chip equalizers, we also propose efficient pilot-based methods. Simulation results show improved performance over the space-time RAKE receiver for the space-time block coded DS-CDMA downlink schemes that have been proposed for the UMTS and IS-2000 W-CDMA standards.

Keywords and phrases: downlink CDMA, space-time block coding, space-time chip equalization.

1. INTRODUCTION

Direct sequence code division multiple access (DS-CDMA) has emerged as the predominant multiple access technique for 3G cellular systems. In the downlink of DS-CDMA, orthogonal user signals are transmitted from the base station. All these signals are distorted by the same channel when propagating to the desired mobile station. Hence, when this channel is frequency-selective, the orthogonality of the user signals is destroyed and severe multiuser interference (MUI) is introduced. Space-time chip equalization can then restore the orthogonality of the user signals and suppress the MUI [1, 2, 3, 4].

Multiple-input multiple-output (MIMO) systems, on the other hand, have recently been shown to realize a significant increase in capacity for rich scattering environments [5, 6, 7]. Both space division multiplexing (SDM) [8, 9] and space-time coding (STC) [10, 11, 12] are popular MIMO communication techniques. SDM techniques mainly aim at an increase in throughput by transmitting different data streams from the different transmit antennas. However, SDM typically requires as many receive as transmit antennas, which seriously impairs a cost-efficient implementation at the mobile station. STC techniques, on the other hand, mainly aim at an increase in performance by introducing spatial and temporal correlation in the transmitted data streams. As opposed to SDM, STC supports any number of receive antennas, and thus enables a cost-efficient implementation at the mobile station. In this perspective, space-time block coding (STBC) techniques, introduced in [11] for two transmit antennas and later generalized in [12] for any number of transmit antennas, are particularly appealing because they facilitate maximum likelihood (ML) detection with simple linear processing. However, these STBC techniques have originally been developed for signaling over frequency-flat channels, and do not enable the maximum multiantenna and multipath diversity present in frequency-selective channels. Therefore, improved STBC techniques have recently been developed for signaling over frequency-selective channels [13, 14, 15]. The STBC technique proposed in [13] enables the maximum multiantenna diversity, and although it is presented as a technique that provides the maximum multipath diversity, it is not possible to prove it without any proper discussion on how to treat the edge effects at the beginning and the end of a burst. If the edge effects are handled by a cyclic prefix as in [14], maximum multipath diversity is not guaranteed. On the other hand, if the edge effects are handled by a zero postfix as in [15], maximum multipath diversity is guaranteed.

Up till now, research on STBC techniques has mainly focused on single-user communication links. In this paper, we aim at combining STBC techniques with the original single-antenna DS-CDMA downlink scheme, resulting into so-called space-time block coded DS-CDMA downlink schemes. As an example, we mention the space-time block coded DS-CDMA downlink schemes that have been proposed for the UMTS and IS-2000 W-CDMA standards, both special cases of the so-called space-time spreading scheme presented in [16], which consists of a mixture of the original single-antenna DS-CDMA downlink scheme and the STBC technique of [12]. However, this scheme does not enable the maximum multiantenna and multipath diversity present in frequency-selective channels. A second example is the space-time block coded DS-CDMA downlink scheme presented in [17], which consists of the original single-antenna DS-CDMA downlink scheme followed by the STBC technique of [14]. However, this scheme only enables the maximum multiantenna diversity but not the maximum multipath diversity (due to the fact that maximum multipath diversity is not provided by the STBC technique of [14]). Therefore, in this paper, we consider the space-time block coded DS-CDMA downlink scheme that consists of the original single-antenna DS-CDMA downlink scheme followed by the STBC technique of [15]. This scheme enables both the maximum multiantenna diversity and the maximum multipath diversity (due to the fact that maximum multipath diversity is provided by the STBC technique of [15]). Although this maximum diversity can only be collected by ML detection, we pursue suboptimal detection by means of space-time chip equalization, which lowers the computational complexity significantly. Note that this suboptimal detection technique can also be applied to the STBC technique of [15] on its own, without combining it with the original single-antenna DS-CDMA downlink scheme.

Assuming there are J transmit antennas, the straightforward way to implement space-time chip equalization is to apply J space-time chip equalizers to recover the J transmitted space-time block coded multiuser chip sequences, then to apply space-time decoding to recover J subsequences of the original multiuser chip sequence, and finally, to perform simple despreading. Since this comes down to an equalization problem with J sources, we need $J + 1$ chip rate sampled outputs at each mobile station for a finite-length zero-forcing (ZF) solution to exist (i.e., $J + 1$ receive antennas if the antennas are sampled at chip rate). However, we will show that the space-time chip equalization and space-time decoding operations can be swapped, which allows us to first apply space-time decoding, then to apply J space-time chip equalizers to recover J subsequences of the original multiuser chip sequence, and finally, to perform simple despreading. Since this comes down to J equalization problems with only one source, we need only two chip rate sampled outputs at each mobile station for a finite-length ZF solution to exist (i.e., two receive antennas if the antennas are sampled at chip rate). To design the space-time chip equalizers, we finally propose efficient pilot-based methods.

In Section 2, we discuss the transceiver design of the proposed space-time block coded DS-CDMA system. We distinguish between the transmitter design, the channel model, and the receiver design, where the latter is based on space-time chip equalization. In Section 3, we then propose two pilot-based methods for practical space-time chip equalizer design. We show some simulation results in Section 4. In Section 5, we finally draw our conclusions.

Notation

We use upper (lower) bold face letters to denote matrices (vectors). Superscripts $*$, T, and H represent conjugate, transpose, and Hermitian, respectively. Further, $\lfloor\cdot\rfloor$ represents the flooring operation, and $\mathcal{E}\{\cdot\}$ represents the expectation operation. We denote the $N \times N$ identity matrix as $\mathbf{I}_N$ and the $M \times N$ all-zero matrix as $\mathbf{0}_{M\times N}$. Next, $[\mathbf{A}]_{m,n}$ denotes the entry at position (m, n) of the matrix $\mathbf{A}$. Finally, $\operatorname{diag}\{\mathbf{a}\}$ represents the diagonal matrix with the vector $\mathbf{a}$ on the diagonal.

2. TRANSCEIVER DESIGN

We consider the downlink of a space-time block coded DS-CDMA system. We assume the base station is equipped with J transmit antennas, and the mobile station is equipped with M receive antennas. In the following, we discuss the transmitter design, the channel model, and the receiver design.

2.1. Transmitter design

At the base station, a space-time block coded DS-CDMA downlink scheme transforms $\{s_u[k]\}_{u=1}^{U}$ and $s_p[k]$, where $s_u[k]$ is the uth user's data symbol sequence and $s_p[k]$ is the pilot symbol sequence, into J space-time block coded multiuser chip sequences $\{u_j[n]\}_{j=1}^{J}$.

We consider the space-time block coded DS-CDMA downlink scheme that consists of the original single-antenna CDMA downlink transmission scheme followed by the STBC technique of [15]. This scheme enables both the maximum multiantenna diversity and the maximum multipath diversity. For simplicity, we will focus on the case of $J = 2$ transmit antennas. Extensions to more than two transmit antennas

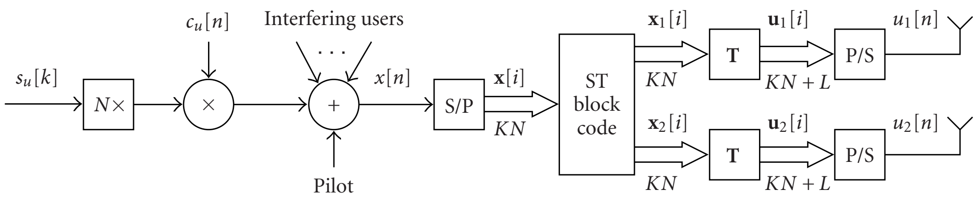

FIGURE 1: Proposed space-time block coded DS-CDMA downlink scheme.

($J > 2$) are straightforward and can be developed following the design rules presented in [18].

Figure 1 depicts the proposed space-time block coded DS-CDMA downlink scheme ($N\times$ repeats each sample N times, whereas "S/P" and "P/S" represent a serial-to-parallel and parallel-to-serial conversion, respectively). First, the original multiuser chip sequence $x[n]$ is constructed:

$$x[n] := \sum_{u=1}^{U} s_u[\lfloor n/N \rfloor]c_u[n] + s_p[\lfloor n/N \rfloor]c_p[n], \quad (1)$$

where $c_u[n]$ is the uth user's code sequence and $c_p[n]$ is the pilot code sequence. We assume that both $c_u[n]$ and $c_p[n]$ are normalized and consist of a multiplication of a user/pilot specific orthogonal Walsh-Hadamard spreading code of length N and a base-station specific long scrambling code. Note that the above pilot insertion technique is similar to the so-called common pilot channel (CPICH) [19] in forthcoming 3G systems. Second, the original multiuser chip sequence $x[n]$ is serial-to-parallel converted into the $1 \times KN$ multiuser chip block sequence $\mathbf{x}[i]$:

$$\mathbf{x}[i] := [x[iKN], \dots, x[(i+1)KN - 1]]. \quad (2)$$

Third, the multiuser chip block sequence $\mathbf{x}[i]$ is transformed into the two $1 \times KN$ block sequences $\mathbf{x}_1[i]$ and $\mathbf{x}_2[i]$:

$$\begin{aligned} &\begin{bmatrix} \mathbf{x}_1[2i] & \mathbf{x}_1[2i+1] \\ \mathbf{x}_2[2i] & \mathbf{x}_2[2i+1] \end{bmatrix} \\ &\quad := \begin{bmatrix} \mathbf{x}[2i] & -\mathbf{x}^*[2i+1]\mathbf{P}_{KN} \\ \mathbf{x}[2i+1] & \mathbf{x}^*[2i]\mathbf{P}_{KN} \end{bmatrix}, \end{aligned} \quad (3)$$

where $\mathbf{P}_N$ is an $N \times N$ permutation matrix that performs a reversal of the entries, that is, $[\mathbf{P}_N]_{n,n'} = \delta[n + n' - N - 1]$. Fourth, we add a zero postfix of length L to each block of the block sequence $\mathbf{x}_j[i]$, resulting into the $1 \times (KN + L)$ block sequence $\mathbf{u}_j[i]$: $\mathbf{u}_j[i] := \mathbf{x}_j[i]\mathbf{T}$, where $\mathbf{T}$ is the $KN \times (KN+L)$ zero postfix insertion matrix: $\mathbf{T} := [\mathbf{I}_{KN}, \mathbf{0}_{KN\times L}]$. Finally, the block sequence $\mathbf{u}_j[i]$ is parallel-to-serial converted into the space-time block coded multiuser chip sequence $u_j[n]$:

$$[u_j[i(KN+L)], \dots, u_j[(i+1)(KN+L) - 1]] := \mathbf{u}_j[i], \quad (4)$$

which is transmitted at the jth transmit antenna with rate $1/T_c$ (the chip rate).

2.2. Channel model

Assuming the mth receive antenna is sampled at the chip rate, the received sequence at the mth receive antenna can be written as

$$y_m[n] = \sum_{j=1}^{2} \sum_{l=0}^{L} h_{m,j}[l]u_j[n-l] + e_m[n], \quad (5)$$

where $e_m[n]$ is the additive noise at the mth receive antenna and $h_{m,j}[l]$ is the channel from the jth transmit antenna to the mth receive antenna, including transmit and receive filters. We assume that $h_{m,j}[l]$ is FIR with order $L_{j,m}$ and that L is a known upper bound on $\max_{j,m}\{L_{j,m}\}$. Note that L was also chosen as the zero postfix length in Section 2.1.

2.3. Receiver design

A first option is to serial-to-parallel convert the received sequence $y_m[n]$ into the $1 \times (KN + L)$ received block sequence $\mathbf{y}_m[i]$:

$$\mathbf{y}_m[i] := [y_m[i(KN+L)], \dots, y_m[(i+1)(KN+L) - 1]], \quad (6)$$

then to apply space-time decoding and Viterbi equalization as in [18], and finally, to perform simple despreading. This detection technique is overall ML, but leads to a very large computational complexity. That is why we pursue suboptimal detection by means of space-time chip equalization, which lowers the computational complexity significantly. Note that this suboptimal detection technique can also be applied to the STBC technique of [15] on its own, without combining it with the original single-antenna DS-CDMA downlink scheme.

We first introduce some new notation. Defining the $M\times 1$ vector

$$\mathbf{y}[n] := [y_1[n], \dots, y_M[n]]^T, \quad (7)$$

we can write

$$\mathbf{y}[n] = \sum_{j=1}^{2} \sum_{l=0}^{L} \mathbf{h}_j[l]u_j[n-l] + \mathbf{e}[n], \quad (8)$$

where $\mathbf{e}[n]$ is similarly defined as $\mathbf{y}[n]$, and

$$\mathbf{h}_j[l] := [h_{1,j}[l], \dots, h_{M,j}[l]]^T. \quad (9)$$

Further, defining the $(Q+1)M \times KN$ matrix

$$\mathbf{Y}[i] := \begin{bmatrix} \mathbf{y}[i(KN+L)] & \cdots & \mathbf{y}[i(KN+L)+KN-1] \\ \vdots & \vdots & \vdots \\ \mathbf{y}[i(KN+L)+Q] & \cdots & \mathbf{y}[i(KN+L)+KN-1+Q] \end{bmatrix}, \tag{10}$$

we can write

$$\mathbf{Y}[i] = \sum_{j=1}^{2} \mathcal{H}_j \mathbf{U}_j[i] + \mathbf{E}[i], \tag{11}$$

where $\mathbf{E}[i]$ is similarly defined as $\mathbf{Y}[i]$,

$$\mathcal{H}_j := \begin{bmatrix} \mathbf{h}_j[L] & \cdots & \mathbf{h}_j[0] & \mathbf{0}_{M\times 1} & \cdots & \mathbf{0}_{M\times 1} \\ \mathbf{0}_{M\times 1} & \mathbf{h}_j[L] & \cdots & \mathbf{h}_j[0] & \cdots & \mathbf{0}_{M\times 1} \\ \vdots & \vdots & \ddots & & \ddots & \vdots \\ \mathbf{0}_{M\times 1} & \mathbf{0}_{M\times 1} & \cdots & \mathbf{h}_j[L] & \cdots & \mathbf{h}_j[0] \end{bmatrix},$$

$$\mathbf{U}_j[i] := \begin{bmatrix} u_j[i(KN+L)-L] & \cdots & u_j[i(KN+L)-L+KN-1] \\ \vdots & \vdots & \vdots \\ u_j[i(KN+L)+Q] & \cdots & u_j[i(N+L)+Q+KN-1] \end{bmatrix}. \tag{12}$$

The parameter Q basically represents the order of the adopted space-time chip equalizer. This equalizer order Q is usually chosen to be close to the channel order L. For the sake of conciseness, we assume $Q = L$. However, the proposed results can easily be extended to other values of the equalizer order Q.

Choosing $Q = L$, it is clear from the zero postfix insertion that $\mathbf{U}_j[i]$ can be expressed as

$$\mathbf{U}_j[i] = \mathcal{T}(\mathbf{x}_j[i]) := \begin{bmatrix} \mathbf{x}_j[i]\mathbf{J}_{KN}^{(-L)} \\ \vdots \\ \mathbf{x}_j[i]\mathbf{J}_{KN}^{(L)} \end{bmatrix}, \tag{13}$$

with $\mathbf{J}_N^{(l)}$ the $N \times N$ shift matrix with $[\mathbf{J}_N^{(l)}]_{n,n'} = \delta[n-n'-l]$ (note that $\mathbf{J}_N^{(0)} = \mathbf{I}_N$).

To proceed, the straightforward way is to apply two space-time chip equalizers on $\mathbf{Y}[i]$ to recover $\mathbf{x}_1[i]$ and $\mathbf{x}_2[i]$, then to apply space-time decoding to recover $\mathbf{x}[2i]$ and $\mathbf{x}[2i+1]$, and finally, to perform simple despreading. Since this comes down to an equalization problem with two sources, we need three chip rate sampled receive antennas at each mobile station for a finite-length ZF solution to exist (for $J > 2$ transmit antennas, we need $J+1$ chip rate sampled receive antennas at each mobile station). However, we will show that the space-time chip equalization and space-time decoding operations can be swapped, which allows us to first apply space-time decoding on $\mathbf{Y}[2i]$ and $\mathbf{Y}[2i+1]$, then to apply two space-time chip equalizers to recover $\mathbf{x}[2i]$ and $\mathbf{x}[2i+1]$, and finally, to perform simple despreading. Since this comes down to two equalization problems with only one source, we need only two chip rate sampled receive antennas at each mobile station for a finite-length ZF solution to exist (even for $J > 2$ transmit antennas, we need only two chip rate sampled receive antennas at each mobile station). The latter option clearly has more degrees of freedom to tackle the equalization problem, and therefore leads to a better performance. This option is explained in more detail next.

2.3.1. Space-time decoding

Using (11) and (13), we can write $\mathbf{Y}[2i]$ and $\mathbf{Y}[2i+1]$ as

$$\begin{aligned} \mathbf{Y}[2i] &= \mathcal{H}_1\mathcal{T}(\mathbf{x}_1[2i]) + \mathcal{H}_2\mathcal{T}(\mathbf{x}_2[2i]) + \mathbf{E}[2i], \\ \mathbf{Y}[2i+1] &= \mathcal{H}_1\mathcal{T}(\mathbf{x}_1[2i+1]) + \mathcal{H}_2\mathcal{T}(\mathbf{x}_2[2i+1]) \\ &\quad + \mathbf{E}[2i+1]. \end{aligned} \tag{14}$$

Since $\mathbf{x}_1[2i+1] = -\mathbf{x}_2^*[2i]\mathbf{P}_{KN}$ (see (3)), we can derive from (13) that

$$\begin{aligned} \mathcal{T}(\mathbf{x}_1[2i+1]) &= \begin{bmatrix} \mathbf{x}_1[2i+1]\mathbf{J}_{KN}^{(-L)} \\ \vdots \\ \mathbf{x}_1[2i+1]\mathbf{J}_{KN}^{(L)} \end{bmatrix} \\ &= -\begin{bmatrix} \mathbf{x}_2^*[2i]\mathbf{P}_{KN}\mathbf{J}_{KN}^{(-L)} \\ \vdots \\ \mathbf{x}_2^*[2i]\mathbf{P}_{KN}\mathbf{J}_{KN}^{(L)} \end{bmatrix} \\ &= -\begin{bmatrix} \mathbf{x}_2^*[2i]\mathbf{J}_{KN}^{(L)} \\ \vdots \\ \mathbf{x}_2^*[2i]\mathbf{J}_{KN}^{(-L)} \end{bmatrix}\mathbf{P}_{KN} \\ &= -\mathbf{P}_{2L+1}\begin{bmatrix} \mathbf{x}_2^*[2i]\mathbf{J}_{KN}^{(-L)} \\ \mathbf{x}_2^*[2i]\mathbf{J}_{KN}^{(L)} \end{bmatrix}\mathbf{P}_{KN} \\ &= -\mathbf{P}_{2L+1}\mathcal{T}^*(\mathbf{x}_2[2i])\mathbf{P}_{KN}. \end{aligned} \tag{15}$$

Similarly, since $\mathbf{x}_2[2i+1] = \mathbf{x}_1^*[2i]\mathbf{P}_{KN}$ (see (3)), we can derive from (13) that

$$\mathcal{T}(\mathbf{x}_2[2i+1]) = \mathbf{P}_{2L+1}\mathcal{T}^*(\mathbf{x}_1[2i])\mathbf{P}_{KN}. \tag{16}$$

Conjugating $\mathbf{Y}[2i+1]$ and multiplying it to the right-hand side with $\mathbf{P}_{KN}$, we then arrive at

$$\begin{aligned} \mathbf{Y}^*[2i+1]\mathbf{P}_{KN} &= \mathcal{H}_1^*\mathcal{T}^*(\mathbf{x}_1[2i+1])\mathbf{P}_{KN} + \mathcal{H}_2^*\mathcal{T}^*(\mathbf{x}_2[2i+1])\mathbf{P}_{KN} \\ &\quad + \mathbf{E}^*[2i+1]\mathbf{P}_{KN} \\ &= -\mathcal{H}_1^*\mathbf{P}_{2L+1}\mathcal{T}(\mathbf{x}_2[2i]) + \mathcal{H}_2^*\mathbf{P}_{2L+1}\mathcal{T}(\mathbf{x}_1[2i]) \\ &\quad + \mathbf{E}^*[2i+1]\mathbf{P}_{KN}, \end{aligned} \tag{17}$$

where the second equality is due to (15) and (16). Stacking $\mathbf{Y}[2i]$ and $\mathbf{Y}^*[2i+1]\mathbf{P}_{KN}$:

$$\bar{\mathbf{Y}}[i] := \begin{bmatrix} \mathbf{Y}[2i] \\ \mathbf{Y}^*[2i+1]\mathbf{P}_{KN} \end{bmatrix}, \tag{18}$$

and using the fact that $\mathbf{x}_1[2i] = \mathbf{x}[2i]$ and $\mathbf{x}_2[2i] = \mathbf{x}[2i+1]$ (see (3)), we finally obtain

$$\bar{\mathbf{Y}}[i] = \mathcal{H}\bar{\mathbf{X}}[i] + \bar{\mathbf{E}}[i], \tag{19}$$

where $\bar{\mathbf{E}}[i]$ is similarly defined as $\bar{\mathbf{Y}}[i]$,

$$\begin{aligned} \mathcal{H} &:= \begin{bmatrix} \mathcal{H}_1 & \mathcal{H}_2 \\ \mathcal{H}_2^* \mathbf{P}_{2L+1} & -\mathcal{H}_1^* \mathbf{P}_{2L+1} \end{bmatrix}, \\ \bar{\mathbf{X}}[i] &:= \begin{bmatrix} \mathcal{T}(\mathbf{x}[2i]) \\ \mathcal{T}(\mathbf{x}[2i+1]) \end{bmatrix}. \end{aligned} \tag{20}$$

2.3.2. Space-time chip equalization

We now apply two space-time chip equalizers on $\bar{\mathbf{Y}}[i]$: $\mathbf{f}_e$ and $\mathbf{f}_o$. The $1 \times 2(L+1)M$ space-time chip equalizer $\mathbf{f}_e$ is designed to extract the even multiuser chip block $\mathbf{x}[2i]$, whereas the $1 \times 2(L+1)M$ space-time chip equalizer $\mathbf{f}_o$ is designed to extract the odd multiuser chip block $\mathbf{x}[2i+1]$:

$$\hat{\mathbf{x}}[2i] = \mathbf{f}_e \bar{\mathbf{Y}}[i], \qquad \hat{\mathbf{x}}[2i+1] = \mathbf{f}_o \bar{\mathbf{Y}}[i]. \tag{21}$$

Note that $\mathbf{x}[2i]$ and $\mathbf{x}[2i+1]$ are two distinct rows of $\bar{\mathbf{X}}[i]$.

A first possibility is to apply two ZF space-time chip equalizers, completely eliminating the interchip interference (ICI) at the expense of potentially excessive noise enhancement:

$$\begin{aligned} \mathbf{f}_e &= \mathbf{i}_e \left(\mathcal{H}^H \mathbf{R}_e^{-1} \mathcal{H}\right)^{-1} \mathcal{H}^H \mathbf{R}_e^{-1}, \\ \mathbf{f}_o &= \mathbf{i}_o \left(\mathcal{H}^H \mathbf{R}_e^{-1} \mathcal{H}\right)^{-1} \mathcal{H}^H \mathbf{R}_e^{-1}, \end{aligned} \tag{22}$$

where $\mathbf{i}_e$ is a $1 \times (4L+2)$ unit vector with a one in the $(L+1)$th position, $\mathbf{i}_o$ is a $1 \times (4L+2)$ unit vector with a one in the $(3L+2)$th position, and $\mathbf{R}_e := 1/(KN)\mathcal{E}\{\bar{\mathbf{E}}[i]\bar{\mathbf{E}}^H[i]\}$. A second possibility is to apply two minimum mean-squared error (MMSE) space-time chip equalizers, balancing ICI elimination with noise enhancement:

$$\begin{aligned} \mathbf{f}_e &= \mathbf{i}_e \left(\mathcal{H}^H \mathbf{R}_e^{-1} \mathcal{H} + \mathbf{R}_x^{-1}\right)^{-1} \mathcal{H}^H \mathbf{R}_e^{-1}, \\ \mathbf{f}_o &= \mathbf{i}_o \left(\mathcal{H}^H \mathbf{R}_e^{-1} \mathcal{H} + \mathbf{R}_x^{-1}\right)^{-1} \mathcal{H}^H \mathbf{R}_e^{-1}, \end{aligned} \tag{23}$$

where $\mathbf{R}_x := 1/(KN)\mathcal{E}\{\bar{\mathbf{X}}[i]\bar{\mathbf{X}}^H[i]\}$.

Assuming the additive noise sequences $\{e_m[n]\}_{m=1}^M$ are mutually uncorrelated and white with variance σ_e^2, we can write $\mathbf{R}_e = \sigma_e^2 \mathbf{I}_{2(L+1)M}$. Furthermore, assuming the data symbol sequences $\{s_u[n]\}_{u=1}^U$ are mutually uncorrelated and white with variance σ_s^2, the original multiuser chip sequence $x[n]$ is white with variance $\sigma_x^2 = \sigma_s^2 J/N$ (justified by the long scrambling code), and we can write $\mathbf{R}_x = \sigma_x^2 \operatorname{diag}\{[\mathbf{r}_x, \mathbf{r}_x]\} = \sigma_s^2 J/N \operatorname{diag}\{[\mathbf{r}_x, \mathbf{r}_x]\}$, where $\mathbf{r}_x = [(KN-L)/(KN), \ldots, (KN-1)/(KN), 1, (KN-1)/(KN), \ldots, (KN-L)/(KN)]$.

2.3.3. Despreading

We define the $1 \times KU$ multiuser data symbol block $\mathbf{s}[i]$ as

$$\mathbf{s}[i] := [\mathbf{s}_1[i], \ldots, \mathbf{s}_U[i]], \tag{24}$$

where $\mathbf{s}_u[i]$ is the uth user's $1 \times K$ data symbol block given by

$$\mathbf{s}_u[i] := [s_u[iK], \ldots, s_u[(i+1)K-1]]. \tag{25}$$

Note that the $1 \times K$ pilot symbol block $\mathbf{s}_p[i]$ is similarly defined as $\mathbf{s}_u[i]$. We further define the multiuser code matrix $\mathbf{C}[i]$ as

$$\mathbf{C}[i] := [\mathbf{C}_1[i]^T, \ldots, \mathbf{C}_U[i]^T]^T, \tag{26}$$

where $\mathbf{C}_u[i]$ is the uth user's code matrix given by

$$\mathbf{C}_u[i] := \begin{bmatrix} \mathbf{c}_u[iK] & & \\ & \ddots & \\ & & \mathbf{c}_u[(i+1)K-1] \end{bmatrix}, \tag{27}$$

with $\mathbf{c}_u[k] := [c_u[kN], \ldots, c_u[(k+1)N-1]]$. Note that the pilot code matrix $\mathbf{C}_p[i]$ is similarly defined as $\mathbf{C}_u[i]$. It is then clear from (1) that the multiuser chip block $\mathbf{x}[i]$ can be expressed as

$$\begin{aligned} \mathbf{x}[i] &= \sum_{u=1}^{U} \mathbf{s}_u[i]\mathbf{C}_u[i] + \mathbf{s}_p[i]\mathbf{C}_p[i] \\ &= \mathbf{s}[i]\mathbf{C}[i] + \mathbf{s}_p[i]\mathbf{C}_p[i]. \end{aligned} \tag{28}$$

Hence, by despreading the multiuser chip block $\mathbf{x}[i]$ with the uth user's code matrix $\mathbf{C}_u[i]$, we obtain

$$\mathbf{s}_u[i] = \mathbf{x}[i]\mathbf{C}_u^H[i] \tag{29}$$

because $\mathbf{C}_p[i]\mathbf{C}_u^H[i] = \mathbf{0}_{K\times K}$, $\mathbf{C}_{u'}[i]\mathbf{C}_u^H[i] = \mathbf{0}_{K\times K}$ for $u \neq u'$, and $\mathbf{C}_u[i]\mathbf{C}_u^H[i] = \mathbf{I}_K$. Therefore, once $\mathbf{x}[i]$ has been estimated, we can find an estimate for $\mathbf{s}_u[i]$ by simple despreading:

$$\hat{\mathbf{s}}_u[i] = \hat{\mathbf{x}}[i]\mathbf{C}_u^H[i]. \tag{30}$$

Plugging (30) into (21), we thus obtain

$$\begin{aligned} \hat{\mathbf{s}}_u[2i] &= \mathbf{f}_e \bar{\mathbf{Y}}[i]\mathbf{C}_u^H[2i], \\ \hat{\mathbf{s}}_u[2i+1] &= \mathbf{f}_o \bar{\mathbf{Y}}[i]\mathbf{C}_u^H[2i+1]. \end{aligned} \tag{31}$$

From these equations, it is also clear that the order of equalization and despreading can be reversed. In other words, we can first despread $\bar{\mathbf{Y}}[i]$ with $\mathbf{C}_u[2i]$ and $\mathbf{C}_u[2i+1]$, and then perform space-time chip equalization on both results.

3. PRACTICAL SPACE-TIME CHIP EQUALIZER DESIGN

In this section, we focus on practical space-time chip equalizer design. In [20, 21], we have developed two pilot-based space-time chip equalizer design methods for the original single-antenna DS-CDMA downlink scheme: a *training-based* method and a *semiblind* method. In this section, these two methods are appropriately modified and applied to the proposed space-time coded DS-CDMA downlink scheme. We consider a burst of $2I$ data symbol blocks.

The goal of the *training-based* method is to compute the uth user's even and odd data symbol blocks $\{\mathbf{s}_u[2i]\}_{i=1}^{I}$ and $\{\mathbf{s}_u[2i+1]\}_{i=1}^{I}$ from $\{\bar{\mathbf{Y}}[i]\}_{i=1}^{I}$, based on the even and odd pilot symbol blocks $\{\mathbf{s}_p[2i]\}_{i=1}^{I}$ and $\{\mathbf{s}_p[2i+1]\}_{i=1}^{I}$, the even and odd pilot code matrices $\{\mathbf{C}_p[2i]\}_{i=1}^{I}$ and $\{\mathbf{C}_p[2i+1]\}_{i=1}^{I}$, and the uth user's even and odd code matrices $\{\mathbf{C}_u[2i]\}_{i=1}^{I}$ and $\{\mathbf{C}_u[2i+1]\}_{i=1}^{I}$.

The goal of the *semiblind* method is to compute the uth user's even and odd data symbol blocks $\{\mathbf{s}_u[2i]\}_{i=1}^{I}$ and $\{\mathbf{s}_u[2i+1]\}_{i=1}^{I}$ from $\{\bar{\mathbf{Y}}[i]\}_{i=1}^{I}$, based on the even and odd pilot symbol blocks $\{\mathbf{s}_p[2i]\}_{i=1}^{I}$ and $\{\mathbf{s}_p[2i+1]\}_{i=1}^{I}$, the even and odd pilot code matrices $\{\mathbf{C}_p[2i]\}_{i=1}^{I}$ and $\{\mathbf{C}_p[2i+1]\}_{i=1}^{I}$, and the even and odd multiuser code matrices $\{\mathbf{C}[2i]\}_{i=1}^{I}$ and $\{\mathbf{C}[2i+1]\}_{i=1}^{I}$. Note that the semiblind method requires the knowledge of the active codes. This knowledge can be obtained by means of a limited feedback from the base station to the mobile station (only the indices of the active codes have to be fed back). However, this knowledge can also be obtained by first adopting the training-based method to design a space-time chip equalizer, and then comparing for each code the energy obtained after equalization and despreading with some threshold in order to decide whether this code is active or not.

For the sake of conciseness, we will only focus on block implementations. These block implementations might look rather complex, but they form the basis for practical low-complexity adaptive implementations, which can be derived in a similar fashion as done in [20, 21].

For the sake of simplicity, we make the following assumptions:

(A1) the matrix $\boldsymbol{\mathcal{H}}$ has full column rank $4L+2$;

(A2) the matrices $\bar{\mathbf{X}}[2i]$ and $\bar{\mathbf{X}}[2i+1]$ have full row rank $4L+2$ for all $i \in \{1,\dots,I\}$.

The first assumption requires that $2(L+1)(M-1) \geq 2L$, which means we need only $M \geq 2$ receive antennas at each mobile station (even for $J > 2$ transmit antennas, we need only $M \geq 2$ receive antennas at each mobile station). The second assumption requires that $4L+2 \leq KN$. Note that these assumptions are not really necessary for the proposed methods to work. The only true requirement is that $\mathbf{x}[2i]$ and $\mathbf{x}[2i+1]$ belong to the row space of $\bar{\mathbf{Y}}[i]$ for all $i \in \{1,\dots,I\}$. Assumptions (A1) and (A2) are sufficient but not necessary conditions for this. However, they considerably simplify the analysis.

Assume no noise is present. Because of assumption (A1), the row space of $\bar{\mathbf{Y}}[i]$ equals the row space of $\bar{\mathbf{X}}[i]$. Hence, there exist two $1 \times 2(L+1)M$ space-time chip equalizers $\mathbf{f}_e$ and $\mathbf{f}_o$, for which

$$\begin{aligned}\mathbf{f}_e\bar{\mathbf{Y}}[i] - \mathbf{x}[2i] &= \mathbf{0}_{1\times KN},\\ \mathbf{f}_o\bar{\mathbf{Y}}[i] - \mathbf{x}[2i+1] &= \mathbf{0}_{1\times KN}.\end{aligned} \tag{32}$$

Because of assumption (A2), these two space-time chip equalizers $\mathbf{f}_e$ and $\mathbf{f}_o$ are ZF. By using (28), we then obtain

$$\begin{aligned}\mathbf{f}_e\bar{\mathbf{Y}}[i] - \mathbf{s}[2i]\mathbf{C}[2i] - \mathbf{s}_p[2i]\mathbf{C}_p[2i] &= \mathbf{0}_{1\times KN},\\ \mathbf{f}_o\bar{\mathbf{Y}}[i] - \mathbf{s}[2i+1]\mathbf{C}[2i+1] - \mathbf{s}_p[2i+1]\mathbf{C}_p[2i+1] &= \mathbf{0}_{1\times KN}.\end{aligned} \tag{33}$$

3.1. Training-based method

By despreading (33) with the even and odd pilot code matrices $\mathbf{C}_p[2i]$ and $\mathbf{C}_p[2i+1]$, we obtain

$$\begin{aligned}\mathbf{f}_e\bar{\mathbf{Y}}[i]\mathbf{C}_p^H[2i] - \mathbf{s}_p[2i] &= \mathbf{0}_{1\times K},\\ \mathbf{f}_o\bar{\mathbf{Y}}[i]\mathbf{C}_p^H[2i+1] - \mathbf{s}_p[2i+1] &= \mathbf{0}_{1\times K}\end{aligned} \tag{34}$$

because $\mathbf{C}[i]\mathbf{C}_p^H[i] = \mathbf{0}_{K\times K}$ and $\mathbf{C}_p[i]\mathbf{C}_p^H[i] = \mathbf{I}_K$. The training-based method solves (34) for $\mathbf{f}_e$ and $\mathbf{f}_o$ for all $i \in \{1,\dots,I\}$. In the noisy case, this leads to the following least squares (LS) problems:

$$\begin{aligned}&\min_{\mathbf{f}_e}\left\{\sum_{i=1}^{I}\left\|\mathbf{f}_e\bar{\mathbf{Y}}[i]\mathbf{C}_p^H[2i] - \mathbf{s}_p[2i]\right\|^2\right\},\\ &\min_{\mathbf{f}_o}\left\{\sum_{i=1}^{I}\left\|\mathbf{f}_o\bar{\mathbf{Y}}[i]\mathbf{C}_p^H[2i+1] - \mathbf{s}_p[2i+1]\right\|^2\right\},\end{aligned} \tag{35}$$

which can be interpreted as follows. The space-time decoded output matrix $\bar{\mathbf{Y}}[i]$ is first equalized with the even and odd space-time chip equalizers $\mathbf{f}_e$ and $\mathbf{f}_o$, and then despread with the even and odd pilot code matrices $\mathbf{C}_p[2i]$ and $\mathbf{C}_p[2i+1]$. The resulting even and odd vectors $\mathbf{f}_e\bar{\mathbf{Y}}[i]\mathbf{C}_p^H[2i]$ and $\mathbf{f}_o\bar{\mathbf{Y}}[i]\mathbf{C}_p^H[2i+1]$ should then be as close as possible in an LS sense to the even and odd pilot symbol blocks $\mathbf{s}_p[2i]$ and $\mathbf{s}_p[2i+1]$ for all $i \in \{1,\dots,I\}$. The solutions of (35) can be written as

$$\begin{aligned}\hat{\mathbf{f}}_e &= \left(\sum_{i=1}^{I}\mathbf{s}_p[2i]\mathbf{C}_p[2i]\bar{\mathbf{Y}}^H[i]\right)\\ &\quad\times\left(\sum_{i=1}^{I}\bar{\mathbf{Y}}[i]\mathbf{C}_p^H[2i]\mathbf{C}_p[2i]\bar{\mathbf{Y}}^H[i]\right)^{-1},\\ \hat{\mathbf{f}}_o &= \left(\sum_{i=1}^{I}\mathbf{s}_p[2i+1]\mathbf{C}_p[2i+1]\bar{\mathbf{Y}}^H[i]\right)\\ &\quad\times\left(\sum_{i=1}^{I}\bar{\mathbf{Y}}[i]\mathbf{C}_p^H[2i+1]\mathbf{C}_p[2i+1]\bar{\mathbf{Y}}^H[i]\right)^{-1}.\end{aligned} \tag{36}$$

The obtained space-time chip equalizers $\hat{\mathbf{f}}_e$ and $\hat{\mathbf{f}}_o$ are subsequently used to estimate the uth user's even and odd data symbol blocks $\mathbf{s}_u[2i]$ and $\mathbf{s}_u[2i+1]$ for all $i \in \{1,\dots,I\}$:

$$\begin{aligned}\hat{\mathbf{s}}_u[2i] &= \hat{\mathbf{f}}_e\bar{\mathbf{Y}}[i]\mathbf{C}_u^H[2i],\\ \hat{\mathbf{s}}_u[2i+1] &= \hat{\mathbf{f}}_o\bar{\mathbf{Y}}[i]\mathbf{C}_u^H[2i+1].\end{aligned} \tag{37}$$

These soft estimates are fed into a decision device that determines the nearest constellation point.

3.2. Semiblind method

The semiblind method directly solves (33) for $(\mathbf{f}_e, \mathbf{s}[2i])$ and $(\mathbf{f}_o, \mathbf{s}[2i+1])$ for all $i \in \{1,\ldots,I\}$. In the noisy case, this leads to the following LS problems:

$$\begin{gathered}\min_{(\mathbf{f}_e,\{\mathbf{s}[2i]\}_{i=1}^I)} \left\{ \sum_{i=1}^{I} \left\| \mathbf{f}_e \bar{\mathbf{Y}}[i] - \mathbf{s}[2i]\mathbf{C}[2i] - \mathbf{s}_p[2i]\mathbf{C}_p[2i] \right\|^2 \right\}, \\ \min_{(\mathbf{f}_o,\{\mathbf{s}[2i+1]\}_{i=1}^I)} \left\{ \sum_{i=1}^{I} \left\| \mathbf{f}_o \bar{\mathbf{Y}}[i] - \mathbf{s}[2i+1]\mathbf{C}[2i+1] - \mathbf{s}_p[2i+1]\mathbf{C}_p[2i+1] \right\|^2 \right\}.\end{gathered} \tag{38}$$

Since we are interested in $\mathbf{f}_e$ and $\mathbf{f}_o$, we can first solve (38) for $\mathbf{s}[2i]$ and $\mathbf{s}[2i+1]$ for all $i \in \{1,\ldots,I\}$, which results into

$$\begin{aligned}\hat{\mathbf{s}}[2i] &= \mathbf{f}_e \bar{\mathbf{Y}}[i]\mathbf{C}^H[2i], \\ \hat{\mathbf{s}}[2i+1] &= \mathbf{f}_o \bar{\mathbf{Y}}[i]\mathbf{C}^H[2i+1]\end{aligned} \tag{39}$$

because $\mathbf{C}[i]\mathbf{C}_p^H[i] = \mathbf{0}_{K\times K}$ and $\mathbf{C}_p[i]\mathbf{C}_p^H[i] = \mathbf{I}_K$. Substituting $\hat{\mathbf{s}}[2i]$ and $\hat{\mathbf{s}}[2i+1]$ in (38) leads to the following LS problems:

$$\begin{gathered}\min_{\mathbf{f}_e} \left\{ \sum_{i=1}^{I} \left\| \mathbf{f}_e \bar{\mathbf{Y}}[i]\left(\mathbf{I}_{KN} - \mathbf{C}^H[2i]\mathbf{C}[2i]\right) - \mathbf{s}_p[2i]\mathbf{C}_p[2i] \right\|^2 \right\}, \\ \min_{\mathbf{f}_o} \left\{ \sum_{i=1}^{I} \left\| \mathbf{f}_o \bar{\mathbf{Y}}[i]\left(\mathbf{I}_{KN} - \mathbf{C}^H[2i+1]\mathbf{C}[2i+1]\right) - \mathbf{s}_p[2i+1]\mathbf{C}_p[2i+1] \right\|^2 \right\},\end{gathered} \tag{40}$$

which can be interpreted as follows. The space-time decoded output matrix $\bar{\mathbf{Y}}[i]$ is first equalized with the even and odd space-time chip equalizers $\mathbf{f}_e$ and $\mathbf{f}_o$ and then projected on the orthogonal complement of the subspace spanned by the even and odd multiuser code matrices $\mathbf{C}[2i]$ and $\mathbf{C}[2i+1]$. The resulting even and odd vectors $\mathbf{f}_e\bar{\mathbf{Y}}[i](\mathbf{I}_{KN} - \mathbf{C}^H[2i]\mathbf{C}[2i])$ and $\mathbf{f}_o\bar{\mathbf{Y}}[i](\mathbf{I}_{KN} - \mathbf{C}^H[2i+1]\mathbf{C}[2i+1])$ should then be as close as possible in an LS sense to the even and odd pilot chip blocks $\mathbf{s}_p[2i]\mathbf{C}_p[2i]$ and $\mathbf{s}_p[2i+1]\mathbf{C}_p[2i+1]$ for all $i \in \{1,\ldots,I\}$. The solutions of (40) can be written as

$$\begin{aligned}\hat{\mathbf{f}}_e &= \left(\sum_{i=1}^{I} \mathbf{s}_p[2i]\mathbf{C}_p[2i]\bar{\mathbf{Y}}^H[i] \right) \times \left(\sum_{i=1}^{I} \bar{\mathbf{Y}}[i]\left(\mathbf{I}_{KN} - \mathbf{C}^H[2i]\mathbf{C}[2i]\right)\bar{\mathbf{Y}}^H[i] \right)^{-1}, \\ \hat{\mathbf{f}}_o &= \left(\sum_{i=1}^{I} \mathbf{s}_p[2i+1]\mathbf{C}_p[2i+1]\bar{\mathbf{Y}}^H[i] \right) \times \left(\sum_{i=1}^{I} \bar{\mathbf{Y}}[i]\left(\mathbf{I}_{KN} - \mathbf{C}^H[2i+1]\mathbf{C}[2i+1]\right)\bar{\mathbf{Y}}^H[i] \right)^{-1}.\end{aligned} \tag{41}$$

The obtained space-time chip equalizers $\hat{\mathbf{f}}_e$ and $\hat{\mathbf{f}}_o$ are subsequently used to estimate the uth user's even and odd data symbol blocks $\mathbf{s}_u[2i]$ and $\mathbf{s}_u[2i+1]$ for all $i \in \{1,\ldots,I\}$:

$$\begin{aligned}\hat{\mathbf{s}}_u[2i] &= \hat{\mathbf{f}}_e \bar{\mathbf{Y}}[i]\mathbf{C}_u^H[2i], \\ \hat{\mathbf{s}}_u[2i+1] &= \hat{\mathbf{f}}_o \bar{\mathbf{Y}}[i]\mathbf{C}_u^H[2i+1].\end{aligned} \tag{42}$$

These soft estimates are fed into a decision device that determines the nearest constellation point.

With some algebraic manipulations, it is easy to prove that (40) is equivalent to

$$\begin{gathered}\min_{\mathbf{f}_e} \left\{ \sum_{i=1}^{I} \left\| \mathbf{f}_e \bar{\mathbf{Y}}[i]\mathbf{C}_p^H[2i] - \mathbf{s}_p[2i] \right\|^2 + \left\| \mathbf{f}_e \bar{\mathbf{Y}}[i]\left(\mathbf{I}_{KN} - \mathbf{C}^H[2i]\mathbf{C}[2i] - \mathbf{C}_p^H[2i]\mathbf{C}_p[2i]\right) \right\|^2 \right\}, \\ \min_{\mathbf{f}_o} \left\{ \sum_{i=1}^{I} \left\| \mathbf{f}_o \bar{\mathbf{Y}}[i]\mathbf{C}_p^H[2i+1] - \mathbf{s}_p[2i+1] \right\|^2 + \left\| \mathbf{f}_o \bar{\mathbf{Y}}[i]\left(\mathbf{I}_{KN} - \mathbf{C}^H[2i+1]\mathbf{C}[2i+1] - \mathbf{C}_p^H[2i+1]\mathbf{C}_p[2i+1]\right) \right\|^2 \right\}.\end{gathered} \tag{43}$$

This shows that (40) naturally decouples into a training-based part and a blind part (hence the name *semiblind*). The training-based part corresponds to (35). The blind part can be interpreted as follows. The space-time decoded output matrix $\bar{\mathbf{Y}}[i]$ is first equalized with the even and odd space-time chip equalizers $\mathbf{f}_e$ and $\mathbf{f}_o$ and then projected on the orthogonal complement of the subspace spanned by the even and odd multiuser code matrices $\mathbf{C}[2i]$ and $\mathbf{C}[2i+1]$ and the even and odd pilot code matrices $\mathbf{C}_p[2i]$ and $\mathbf{C}_p[2i+1]$. The resulting even and odd vectors $\mathbf{f}_e\bar{\mathbf{Y}}[i](\mathbf{I}_{KN} - \mathbf{C}^H[2i]\mathbf{C}[2i] - \mathbf{C}_p^H[2i]\mathbf{C}_p[2i])$ and $\mathbf{f}_o\bar{\mathbf{Y}}[i](\mathbf{I}_{KN} - \mathbf{C}^H[2i+1]\mathbf{C}[2i+1] - \mathbf{C}_p^H[2i+1]\mathbf{C}_p[2i+1])$ should then be as small as possible in an LS sense for all $i \in \{1,\ldots,I\}$. Note that when the user load increases, the orthogonal complement of the subspace spanned by the even and odd multiuser code matrices $\mathbf{C}[2i]$ and $\mathbf{C}[2i+1]$ and the even and odd pilot code matrices $\mathbf{C}_p[2i]$ and $\mathbf{C}_p[2i+1]$ decreases in dimension. As a result, the information that the blind part contributes to the training-based part diminishes, and the semiblind method converges to the training-based method. In the extreme case when the system is fully loaded, that is, $N = U - 1$, the orthogonal complement of the subspace spanned by the even and odd multiuser code matrices $\mathbf{C}[2i]$ and $\mathbf{C}[2i+1]$ and the even and odd pilot code matrices $\mathbf{C}_p[2i]$ and $\mathbf{C}_p[2i+1]$ is empty, that is, $\mathbf{I}_{KN} - \mathbf{C}^H[2i]\mathbf{C}[2i] - \mathbf{C}_p^H[2i]\mathbf{C}_p[2i] = \mathbf{0}_{KN\times KN}$ and $\mathbf{I}_{KN} - \mathbf{C}^H[2i+1]\mathbf{C}[2i+1] - \mathbf{C}_p^H[2i+1]\mathbf{C}_p[2i+1] = \mathbf{0}_{KN\times KN}$. Hence, the blind part does not contribute any additional information to the training-based part, and the semiblind method reduces to the training based method, that is, (43) reduces to (35).

4. SIMULATION RESULTS

In this section, we compare the proposed space-time chip equalizer for the proposed space-time coded downlink CDMA transmission scheme with the space-time RAKE receiver for the space-time spreading scheme, which encompasses the space-time coded downlink CDMA transmission schemes that have been proposed for the UMTS and IS-2000 W-CDMA standards [16]. We do not consider channel codes when comparing the above transceivers. Otherwise, it will not be very clear whether a performance gain is due to the transceiver or the channel code. Moreover, the influence of channel codes on performance has been studied extensively in literature. In W-CDMA, the target coded BER typically is 10^{-6}, which boils down to an uncoded BER of 10^{-2} with a convolutional code of rate 1/2, constraint length 7, and soft decision Viterbi [22]. Therefore, we compare the different transceivers at an uncoded BER of 10^{-2} in the sequel.

We consider a downlink CDMA system with a spreading factor of $N = 32$, $J = 2$ transmit antennas at the base station, and $M = 2$ receive antennas at each mobile station. We assume that all channels are independent. We further assume that each channel $h_{j,m}[n]$ is FIR with order $L_{j,m} = 3$ and has independent Rayleigh fading channel taps of equal variance σ_h^2. Note that the bandwidth efficiency of the proposed space-time coded downlink CDMA transmission scheme is $\epsilon_1 = KU/(KN + L)$, whereas the bandwidth efficiency of the space-time spreading scheme is $\epsilon_2 = U/N$. Hence, in order to make a fair comparison between the two systems, their spectral efficiencies should be comparable. We therefore take $K = 5$ and $L = 3$ for the proposed space-time coded downlink CDMA transmission scheme, which results into $\epsilon_1/\epsilon_2 \approx 0.98$. We assume QPSK modulated data symbols, and define the signal-to-noise ratio (SNR) as the received bit energy over the noise power:

$$\begin{aligned} \text{SNR} &= \frac{\sigma_s^2/2 \sum_{j=1}^{2} \sum_{l=0}^{L} \mathcal{E}\left\{\left\|\mathbf{h}_j[l]\right\|^2\right\}}{\sigma_e^2} \\ &= \frac{2(L+1)\sigma_s^2\sigma_h^2}{\sigma_e^2}. \end{aligned} \tag{44}$$

Two test cases are investigated.

Test case 1

We first assume that the pilot enables us to obtain perfect channel knowledge at the receiver. We then compare the proposed MMSE space-time chip equalizer for the proposed space-time coded downlink CDMA transmission scheme with the MMSE space-time RAKE receiver for the space-time spreading scheme (see [23, 24]), which is different from the matched space-time RAKE receiver for the space-time spreading scheme (see [16]) because it uses an MMSE filter instead of a matched filter to combine the finger outputs. It has been shown in [23, 24] that for the space-time spreading scheme, the MMSE space-time RAKE receiver significantly outperforms the matched space-time RAKE receiver. Figures 2, 3, and 4 compare the performance of the two transceivers for $U = 1$, $U = 15$, and $U = 31$ users, respectively. The performance results are averaged over 1000 random channel realizations, where for each channel realization, we consider 10 random data and noise realizations corresponding to $I = 10$ (100 data symbols per user). Also shown is the theoretical performance of $\sum_{j,m}(L_{j,m} + 1) = 16$-fold diversity over Rayleigh fading channels [22].

First of all, we see that the proposed transceiver comes close to extracting the maximum diversity at low-to-medium

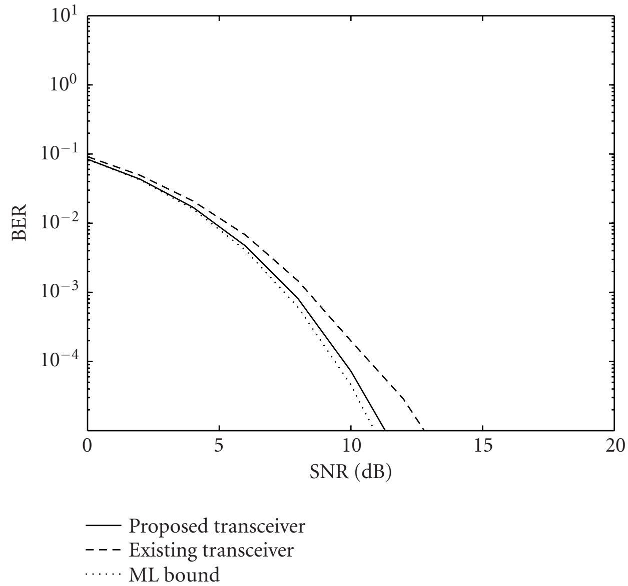

FIGURE 2: Performance comparison for $U = 1$.

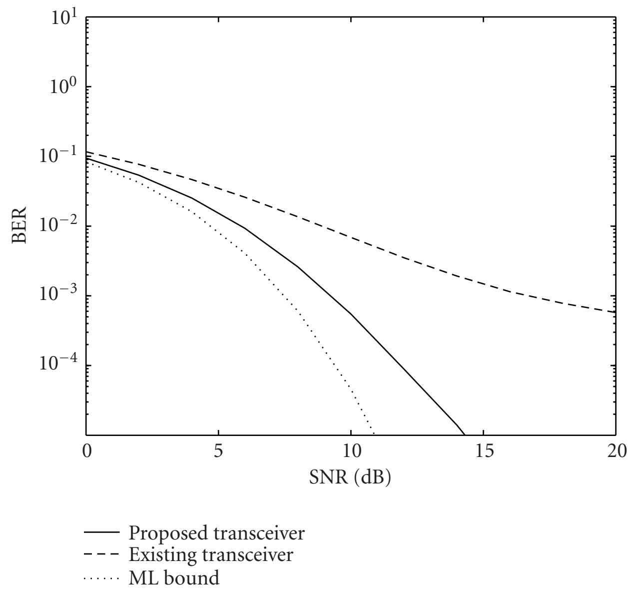

FIGURE 3: Performance comparison for $U = 15$.

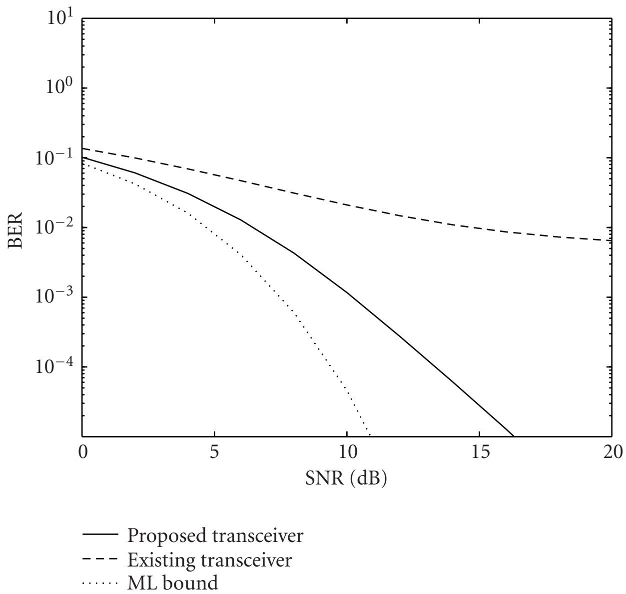

Figure 4: Performance comparison for $U = 31$.

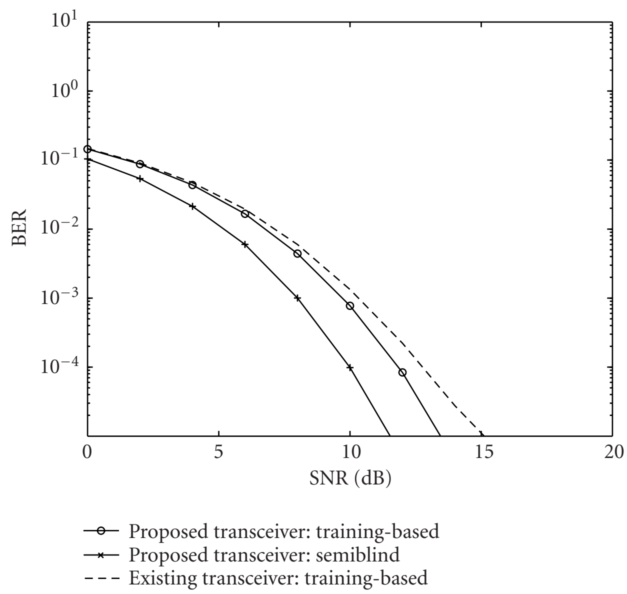

Figure 5: Performance of pilot-based methods for $U = 1$.

user loads. More specifically, at a BER of 10^{-2}, the proposed transceiver incurs a 0.1, 1, and 1.8 dB loss compared to the theoretical ML bound for $U = 1$, $U = 15$, and $U = 31$ users, respectively. The existing transceiver, on the other hand, performs poorly at medium-to-high user loads. At a BER of 10^{-2}, it incurs a 0.5, 3, and 8.2 dB performance loss compared to the proposed transceiver for $U = 1$, $U = 15$, and $U = 31$ users, respectively. The existing transceiver is not capable of completely suppressing the MUI at high SNR. This results into a flooring of the BER at high SNR. Note that the flooring level increases with the number of users U.

Test case 2

We now investigate the performance of the pilot-based methods. Note that for the space-time spreading scheme, it is easy to derive a training-based method to estimate the combining filter of the space-time RAKE receiver based on the knowledge of the pilot. The performance results are again averaged over 1000 random channel realizations, where for each channel realization, we consider 10 random data and noise realizations corresponding to $I = 10$ (100 data symbols per user). Figures 5, 6, and 7 compare the performance of the different methods for $U = 1$, $U = 15$, and $U = 31$ users, respectively.

First of all, we observe that the difference between the training-based method and the semiblind method for the proposed transceiver decreases with an increasing user load, as indicated in Section 3.2. Next, we observe that the training-based method for the existing transceiver performs much worse than the training-based and semiblind methods for the proposed transceiver at medium-to-high user loads. Finally, note that for the proposed transceiver, the MMSE performance discussed in test case 1 can be viewed as the convergence point of the training-based and semiblind methods as I goes to infinity. Comparing the figures of test case 2 with the figures of test case 1, we observe that for $I = 10$, the training-based method is still far from the MMSE performance, whereas the semiblind method is already very close to the MMSE performance. Hence, as I increases, the semiblind method converges faster to the MMSE performance than the training-based method.

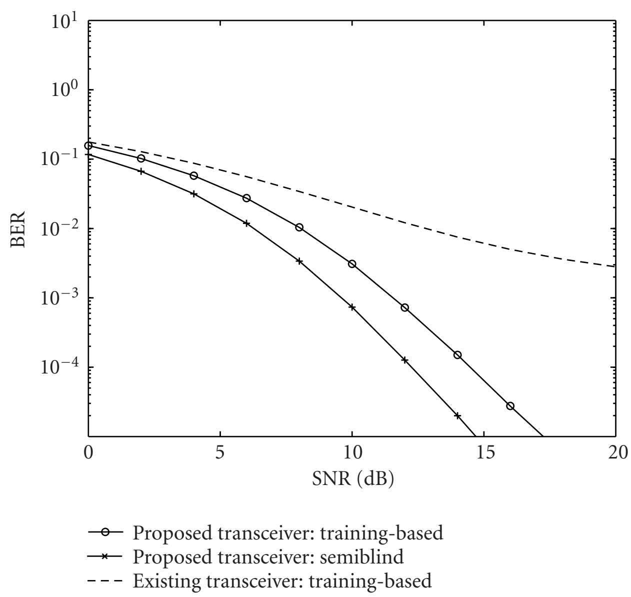

Figure 6: Performance of pilot-based methods for $U = 15$.

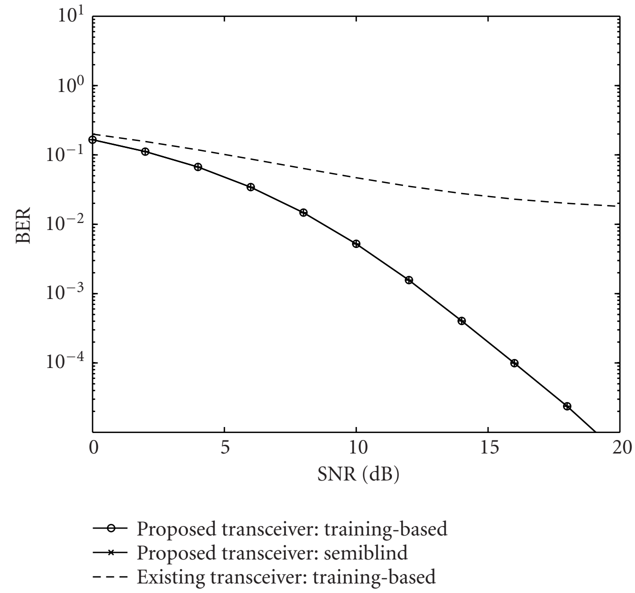

Figure 7: Performance of pilot-based methods for $U = 31$.

5. CONCLUSIONS

We have aimed at combining STBC techniques with the original single-antenna DS-CDMA downlink scheme, resulting into the so-called space-time block coded DS-CDMA downlink schemes. Many space-time block coded DS-CDMA downlink transmission schemes can be considered. We have focussed on a new scheme that enables both the maximum multiantenna diversity and the maximum multipath diversity. Although this maximum diversity can only be collected by ML detection, we have pursued suboptimal detection by means of space-time chip equalization, which lowers the computational complexity significantly. To design the space-time chip equalizers, we have also proposed efficient pilot-based methods. Simulation results have shown improved performance over the space-time RAKE receiver for the space-time block coded DS-CDMA downlink schemes that have been proposed for the UMTS and IS-2000 W-CDMA standards.

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