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Space-Time Coded MC-CDMA: Blind Channel Estimation, Identifiability, and Receiver Design

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Integrating the strengths of multicarrier (MC) modulation and code division multiple access (CDMA), MC-CDMA systems are of great interest for future broadband transmissions. This paper considers the problem of channel identification and signal combining/detection schemes for MC-CDMA systems equipped with multiple transmit antennas and space-time (ST) coding. In particular, a subspace-based blind channel identification algorithm is presented. Identifiability conditions are examined and specified which guarantee unique and perfect (up to a scalar) channel estimation when knowledge of the noise subspace is available. Several popular single-user based signal combining schemes, namely the maximum ratio combining (MRC) and the equal gain combining (EGC), which are often utilized in conventional single-transmit-antenna-based MC-CDMA systems, are extended to the current ST-coded MC-CDMA (STC-MC-CDMA) system to perform joint combining and decoding. In addition, a linear multiuser minimum mean square error (MMSE) detection scheme is also presented, which is shown to outperform the MRC and EGC at some increased computational complexity. Numerical examples are presented to evaluate and compare the proposed channel identification and signal detection/combining techniques.

Keywords and phrases: MC-CDMA, OFDM, space-time coding, blind channel identification, identifiability, multiuser detection.

1. INTRODUCTION

Multicarrier (MC) technologies (e.g., [1] and the references therein), in particular OFDM (orthogonal frequency division multiplexing), are considered very promising for future broadband data services in fading environments. OFDM has been proposed in various standards, for example, digital audio/video broadcasting [2, 3] and wireless local area networks including the IEEE802.11a [4, 5] and HIPERLAN/2 [6, 7, 8]. Recently, the combination of MC modulation with CDMA (code division multiple access) has been of significant interest as a means to take such advantages as bandwidth efficiency, fading resilience, and interference suppression capability which are crucial in future broadband data transmissions [9, 10].

There are several multiple access schemes based on the combination of MC modulation and CDMA, including multicarrier CDMA (MC-CDMA), multicarrier DS-CDMA (MC-DS-CDMA), and multitone CDMA (MT-CDMA) (see [11] and the references therein). In OFDM based MC-CDMA systems, the frequency selective fading channel is divided into a number of narrowband subchannels that are (approximately) frequency nonselective. At the transmitter, the information symbol is first spread by a spreading code, followed by OFDM modulation (via the inverse fast Fourier transform, IFFT) such that each chip of the spreading code is modulated by one subcarrier [11, 12, 13]. A cyclic prefix (CP) of length longer than the delay spread of the impulse response of the multipath channel is then inserted into each OFDM symbol before transmission. At the receiver, the CP, which contains the intersymbol interference (ISI), is first removed. The ISI-free signal is then FFT (fast Fourier transform) converted, which effectively performs OFDM demodulation. Finally, the OFDM demodulated signals corresponding to different subcarriers are combined to produce decision variables for detection.

While orthogonal spreading codes are often used to maintain orthogonality among different user transmissions in the forward link of single-carrier CDMA systems, for example, IS-95 [14], user orthogonality is usually lost in MC-CDMA systems when operating in frequency selective environments which gives rise to multiuser interference (MUI) [11]. This is due to that different subcarriers in an MC system are subject to different attenuation as well as different phase shift. To reduce MUI, several signal combining schemes have been proposed (e.g., [11] and the references therein). These combining schemes, however, require the knowledge of the channel status information (CSI). The CSI has to be estimated at the receiver either with the help of pilot carriers [15] or through blind (self-recovering) identification schemes [16, 17].

It has been shown that significant performance improvement can be obtained when adaptive modulation is used with OFDM [18]. Adaptive modulation allocates subcarriers to users based on the instantaneous channel gain so that the subcarriers can be used more effectively [19]. Optimized bit-loading and power allocation are important techniques to increase the capacity of OFDM systems [19]. The implementation of adaptive modulation and bit/power loading, however, requires the CSI at the transmitter, which is usually estimated at the receiver and relayed back to the transmitter through some feedback links.

In this paper, we consider MC-CDMA systems equipped with multiple transmit antennas and some space-time (ST) block coded transmission schemes [20] in order to provide transmit diversity to the receiver. In particular, we consider the simple and yet effective Alamouti's ST coding scheme that involves two transmit antennas [21]. While it has been found that transmit diversity and ST coding in MC-CDMA systems can significantly increase the transmission rate and improve the overall system performance, channel estimation for such systems is a challenging task, due to the fact that the received signal is mixed both temporally and spatially [22]. We present herein a subspace-based blind channel identification algorithm for ST-coded MC-CDMA (STC-MC-CDMA) systems. We examine the associated identifiability issue and specify conditions under which unique and perfect (up to a scalar) channel estimation is guaranteed when exact knowledge of the noise subspace is available.

Subspace-based estimation has been extensively studied recently and various applications have been reported (see [23, 24, 25, 26, 27, 28] and the references therein). For example, blind channel identification via subspace estimation was first introduced in [23], where a single-user TDMA-like system in frequency selective channels was considered. Subspace-based blind channel identification for single-carrier CDMA systems were studied in [25, 26], where the related identifiability issue was also investigated. Multiuser detection, using subspace decomposition for single-carrier CDMA systems, was considered in [27]. Subspace-based channel estimation for asynchronous single-carrier CDMA was studied in [28] where the problem of code acquisition was also stressed. The proliferation and success of subspace-based estimation in the above systems make it a well-motivated effort to extend this technique and investigate its performance in STC-MC-CDMA systems, where channel estimation is known to be very challenging [22]. Identifiability of subspace-based estimation for STC-MC-CDMA is also an important issue by itself since existing identifiability results for single-carrier CDMA (e.g., [26]) cannot be applied here, due to the use of MC modulation and ST coding.

Once the estimates of the channel response are available at the receiver, several signal combining schemes can be implemented to demodulate and decode the ST-coded transmission. Specifically, we extend herein two signal combining schemes, namely the maximum ratio combining (MRC) and the equal gain combining (EGC) which are often utilized in conventional single-transmit-antenna-based MC-CDMA systems [11], to STC-MC-CDMA to perform joint combining and decoding. While both MRC and EGC are single-user detection schemes based on *per-subcarrier* combining, we also present a linear MMSE (minimum mean square error) multiuser detector which performs *joint-subcarrier* combining and decoding. The performance of the MRC, EGC, and MMSE combining/detection schemes is compared with one another via numerical examples. We also illustrate the performance gain of STC-MC-CDMA systems over conventional MC-CDMA systems without transmit diversity.

The rest of this paper is organized as follows. In Section 2, we introduce the baseband data model for STC-MC-CDMA systems and formulate the problem of interest. The subspace-based blind channel identification algorithm is presented in Section 3. We also discuss therein the identifiability conditions and implementation issues of the proposed identification algorithm. In Section 4, we discuss the MRC, EGC, and linear MMSE combing/detection schemes for STC-MC-CDMA systems. Section 5 contains numerical examples which evaluate the performance of the proposed blind channel identification and signal detection/combining schemes. Finally, we conclude the paper in Section 6.

Notation

Vectors (matrices) are denoted by boldface lower (upper) case letters; all vectors are column vectors; superscripts $*$, T, and H denote the complex conjugate, transpose, and conjugate transpose, respectively; $\mathbf{I}_N$ denotes the $N \times N$ identity matrix; $\mathbf{0}$ denotes an all-zero vector/matrix; $\mathrm{ran}(\cdot)$ denotes the range space of a matrix argument; $E\{\cdot\}$ denotes the statistical expectation; $\otimes$ denotes the matrix Kronecker product [29]; and $\mathrm{diag}\{\cdot\}$ denotes a diagonal matrix.

2. PROBLEM FORMULATION

2.1. System model

Consider a synchronous (downlink) K-user STC-MC-CDMA system equipped with two transmit antennas, Tx1 and Tx2, and one receive antenna, Rx. We focus herein on the downlink (i.e., base station to mobile) since multiple antennas are more often installed at the base station than at the mobile; furthermore, the downlink is also considered the bottleneck direction in future mobile networks since future data transmissions will be highly asymmetric, requiring much faster downlink transmission rates than uplink transmission rates [30]. Figure 1 depicts the diagram of a baseband STC-MC-CDMA system with the Alamouti's ST coding scheme [21] shown only for user k. The

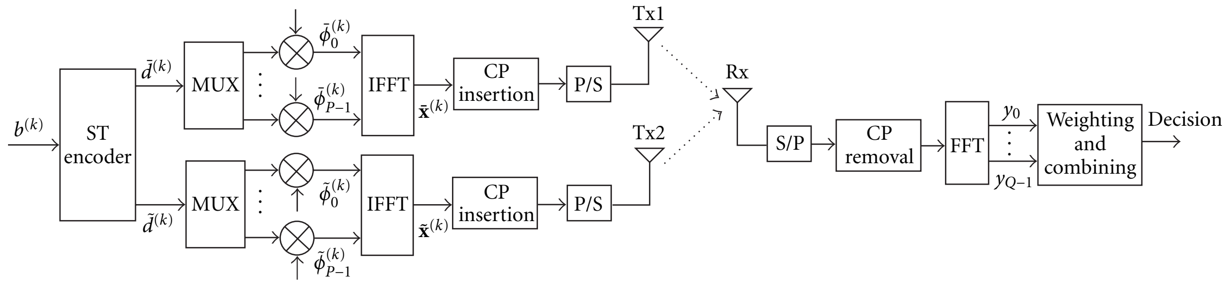


Figure 1: Diagram of a baseband STC-MC-CDMA system.

frequency-selective channel between each Tx and Rx over the entire system bandwidth is divided into Q narrowband subchannels. With a proper choice of Q, each subchannel can be treated as (approximately) frequency nonselective. At the transmitter, the ST encoder (specified in Section 2.2) first maps each user's incoming symbol stream $\{b^{(k)}(n)\}$ drawn from some unit-energy constellation $\mathscr{B}$ into two ST-coded streams: $\{\bar{d}^{(k)}(n)\}$ and $\{\tilde{d}^{(k)}(n)\}$.[1] Next, each of the two ST-coded streams is multiplexed into Q parallel substreams and subsequently spread by a distinctive spreading code (see Figure 1). In this system, each user is assigned to two spreading codes of processing gain P, $\bar{\boldsymbol{\phi}}^{(k)} = [\bar{\phi}_0^{(k)}, \dots, \bar{\phi}_{P-1}^{(k)}]^T$ and $\tilde{\boldsymbol{\phi}}^{(k)} = [\tilde{\phi}_0^{(k)}, \dots, \tilde{\phi}_{P-1}^{(k)}]^T$ to spread symbols transmitted from Tx1 and Tx2, respectively. A trade-off is often made between the bandwidth allocated to an MC-CDMA system and the processing gain P such that Q is an integer multiple of P [11]. Here, we assume that $P = Q$ to simplify the presentation; in the event that $Q > P$, multiple user symbols can be spread and transmitted across the entire system bandwidth simultaneously [11]. After spreading, the IFFT of the spread signal is computed (to perform OFDM modulation) along with CP insertion. Finally, the STC-MC-CDMA signal is parallel-to-serial (P/S) converted and sent out by the two Tx's. At the receiver, the received signal is first serial-to-parallel (S/P) converted and followed by CP removal and FFT (to perform OFDM demodulation). The FFT processor outputs are then weighted and combined to generate decision variables by utilizing channel estimates obtained by either some training or blind channel estimation scheme.

Let $\bar{\mathbf{u}}^{(k)}(n) = [\bar{u}_0^{(k)}(n), \dots, \bar{u}_{P-1}^{(k)}(n)]^T$ which consists of samples resulting from the spreading of $\{\bar{d}^{(k)}(n)\}$. That is,

$$\begin{aligned} \bar{\mathbf{u}}^{(k)}(n) &= \bar{d}^{(k)}(n)\bar{\boldsymbol{\phi}}^{(k)} \\ &= \left[\bar{d}^{(k)}(n)\bar{\phi}_0^{(k)}, \dots, \bar{d}^{(k)}(n)\bar{\phi}_{P-1}^{(k)}\right]^T. \end{aligned} \tag{1}$$

The frequency selective channel between Tx1 and Rx is divided into $Q = P$ subchannels with normalized center frequencies (subcarriers) $\varphi_p = p/P$, $p = 0, 1, \dots, P-1$. (The channel between Tx2 and Rx is similarly divided.) Note that spreading is performed in the frequency domain as in standard MC-CDMA systems [11]. The subcarriers at frequencies $\{\varphi_p\}_{p=0}^{P-1}$ are orthogonal to each other. Hence, when (frequency) coherent demodulation is utilized, interference between subcarriers can be avoided [31]. Unlike plain OFDM systems which suffer from channel fading (i.e., some subcarriers are severely attenuated), MC-CDMA systems are insensitive to channel fading since each information symbol is transmitted across the entire system bandwidth.

The baseband discrete-time signal of user k corresponding to the pth sample of the nth OFDM symbol is (see [1])

$$\bar{x}_p^{(k)}(n) = \sum_{p=0}^{P-1} \bar{d}^{(k)}(n)\bar{\phi}_p^{(k)} e^{j2\pi p\varphi_p}, \quad p = 0, 1, \dots, P-1. \tag{2}$$

Equivalently, the OFDM symbol can be obtained by computing the P-point IFFT of $\bar{\mathbf{u}}^{(k)}(n)$,

$$\bar{\mathbf{x}}^{(k)}(n) = \boldsymbol{\mathcal{F}}^H \bar{\mathbf{u}}^{(k)}(n), \tag{3}$$

where $\bar{\mathbf{x}}^{(k)}(n) = [\bar{x}_0^{(k)}(n), \dots, \bar{x}_{P-1}^{(k)}(n)]^T$ and $\boldsymbol{\mathcal{F}}$ denotes the $P \times P$ FFT matrix with the (p,q)th element given by $P^{-1/2}\exp(-j2\pi pq/P)$ (and so $\boldsymbol{\mathcal{F}}^H$ denotes the $P \times P$ IFFT matrix). The signal from all K users to be transmitted from Tx1 is thus given by

$$\bar{\mathbf{x}}(n) = \sum_{k=1}^{K} \bar{\mathbf{x}}^{(k)}(n) = \sum_{k=1}^{K} \boldsymbol{\mathcal{F}}^H \bar{\mathbf{u}}^{(k)}(n). \tag{4}$$

Before transmission, a CP of length L is inserted in $\bar{\mathbf{x}}(n)$ to combat ISI. The length L is chosen such that $L \geq M$, where M denotes the maximum length of the channel between Tx1 and Rx (and the channel between Tx2 and Rx as well). The CP consists of the last L samples of $\bar{\mathbf{x}}(n)$ and is inserted at the beginning of $\bar{\mathbf{x}}(n)$ [32], resulting in an (composite) OFDM symbol of $(P + L)$ samples.

The frequency selective channel between Tx1 (resp., Tx2) and Rx is modeled as an FIR (finite-duration impulse response) filter with at most M taps [33]. Let

$$\begin{aligned} \bar{\mathbf{h}} &\triangleq \left[\bar{h}(0), \dots, \bar{h}(M-1)\right]^T, \\ \tilde{\mathbf{h}} &\triangleq \left[\tilde{h}(0), \dots, \tilde{h}(M-1)\right]^T, \end{aligned} \tag{5}$$

[1]Throughout the paper, $(\bar{\cdot})$ (resp., $(\tilde{\cdot})$) is designated to quantities associated with the first (resp., the second) transmit antenna.

which collect the channel coefficients. Let $\bar{\mathbf{g}}$ and $\tilde{\mathbf{g}}$ be the frequency response of $\bar{\mathbf{h}}$ and $\tilde{\mathbf{h}}$, respectively. That is,

$$\bar{\mathbf{g}} = \mathcal{F}_M \bar{\mathbf{h}} = \left[\bar{g}(0), \ldots, \bar{g}(P-1)\right]^T, \qquad \tilde{\mathbf{g}} = \mathcal{F}_M \tilde{\mathbf{h}} = \left[\tilde{g}(0), \ldots, \tilde{g}(P-1)\right]^T, \tag{6}$$

where $\mathcal{F}_M \in \mathbb{C}^{P\times M}$ is composed of the first M columns of the $P \times P$ FFT matrix $\mathcal{F}$.

At the receiver, the CP of each OFDM symbol is first removed, followed by OFDM demodulation (via a P-point FFT). Let $\mathbf{y}(n) = [y_0(n), \ldots, y_{P-1}(n)]^T$ consisting of the output samples of the P-point FFT processor applied to the nth OFDM symbol. Then,

$$\mathbf{y}(n) = \sum_{k=1}^{K} \left[\bar{d}^{(k)}(n)\bar{\mathbf{\Phi}}^{(k)}\bar{\mathbf{g}} + \tilde{d}^{(k)}(n)\tilde{\mathbf{\Phi}}^{(k)}\tilde{\mathbf{g}}\right] + \mathbf{v}(n), \tag{7}$$

where

$$\bar{\mathbf{\Phi}}^{(k)} = \operatorname{diag}\left\{\bar{\phi}_0^{(k)}, \ldots, \bar{\phi}_{P-1}^{(k)}\right\}, \qquad \tilde{\mathbf{\Phi}}^{(k)} = \operatorname{diag}\left\{\tilde{\phi}_0^{(k)}, \ldots, \tilde{\phi}_{P-1}^{(k)}\right\}, \tag{8}$$

and $\mathbf{v}(n) = [v_0(n), \ldots, v_{P-1}(n)]^T$ consisting of the channel noise samples (after FFT). To have a more compact expression for $\mathbf{y}(n)$, define

$$\mathbf{d}^{(k)}(n) \triangleq \left[\bar{d}^{(k)}(n), \tilde{d}^{(k)}(n)\right]^T, \qquad \mathbf{\Phi}^{(k)} \triangleq \left[\bar{\mathbf{\Phi}}^{(k)}, \tilde{\mathbf{\Phi}}^{(k)}\right], \qquad \mathcal{G} \triangleq \begin{bmatrix} \bar{\mathbf{g}} & \mathbf{0} \\ \mathbf{0} & \tilde{\mathbf{g}} \end{bmatrix}. \tag{9}$$

Then, (7) can be rewritten as

$$\mathbf{y}(n) = \sum_{k=1}^{K} \mathbf{\Phi}^{(k)}\mathcal{G}\mathbf{d}^{(k)}(n) + \mathbf{v}(n). \tag{10}$$

2.2. ST encoder

The ST encoder implements the Alamouti's ST coding scheme [21]. Assume that the user symbols $\{b^{(k)}(n)\}$ are drawn from some unit-energy constellation (e.g., PSK). For user k, the ST encoder takes in two consecutive data symbols $b^{(k)}(2n)$ and $b^{(k)}(2n+1)$ and outputs the following code matrix:

$$\mathbf{D}^{(k)}(n) \triangleq \begin{bmatrix} \bar{d}^{(k)}(2n) & \bar{d}^{(k)}(2n+1) \\ \tilde{d}^{(k)}(2n) & \tilde{d}^{(k)}(2n+1) \end{bmatrix}, \tag{11}$$

where

$$\bar{d}^{(k)}(2n) = b^{(k)}(2n), \qquad \bar{d}^{(k)}(2n+1) = -b^{(k)*}(2n+1),$$
$$\tilde{d}^{(k)}(2n) = b^{(k)}(2n+1), \qquad \tilde{d}^{(k)}(2n+1) = b^{(k)*}(2n). \tag{12}$$

The two columns of $\mathbf{D}^{(k)}(n)$ are transmitted in two consecutive time slots, with the first element of each column transmitted from Tx1 and the second one from Tx2, respectively.

The problem of interest to us is to estimate the unknown channel coefficients $\{\bar{h}(i)\}_{i=0}^{M-1}$ and $\{\tilde{h}(i)\}_{i=0}^{M-1}$ from the received signal $\mathbf{y}(n)$ and to recover the user symbols $\{b^{(k)}(n)\}$, $k = 1, \ldots, K$.

3. SUBSPACE-BASED BLIND CHANNEL IDENTIFICATION

The channel status information (CSI) is needed at the receiver for coherent signal detection. Training-assisted channel estimation for STC-MC-CDMA systems was investigated in [34]. However, training-assisted schemes consume precious bandwidth. Alternatively, differential ST coding schemes [35, 36] can be utilized to bypass the need for channel estimation, but at the cost of a loss in SNR (signal-to-noise ratio) of about 3 dB compared to coherent detection [35]. In this section, we present a subspace-based blind channel identification method which utilizes only the second-order statistics of the received signal and, thus, leads to higher bandwidth efficiency than training-assisted schemes.

3.1. Subspace blind channel identification

Define $\mathbf{\Phi} \triangleq [\mathbf{\Phi}^{(1)}, \ldots, \mathbf{\Phi}^{(K)}] \in \mathbb{R}^{P\times 2PK}$ and $\mathbf{d}(n) \triangleq [\mathbf{d}^{(1)T}(n), \ldots, \mathbf{d}^{(K)T}(n)]^T$. Then, (10) can be expressed as

$$\mathbf{y}(n) = \mathbf{\Phi}(\mathbf{I}_K \otimes \mathcal{G})\mathbf{d}(n) + \mathbf{v}(n). \tag{13}$$

Denote by $\mathbf{R}_y$ the covariance matrix of $\mathbf{y}(n)$,

$$\mathbf{R}_y \triangleq E\{\mathbf{y}(n)\mathbf{y}^H(n)\} = \mathbf{\Phi}(\mathbf{I}_K \otimes \mathcal{G})(\mathbf{I}_K \otimes \mathcal{G})^H\mathbf{\Phi}^H + \sigma_v^2\mathbf{I}_P, \tag{14}$$

where σ_v^2 denotes the variance of the noise samples $\{v_p(n)\}$, and we also made the following assumption.

(A1) User symbols $\{b^{(k)}(n)\}$ are i.i.d. (independently and identically distributed) and drawn from a unit-energy constellation so that $E\{\mathbf{d}(n)\mathbf{d}^H(n)\} = \mathbf{I}_{2K}$.

Let R denote the rank of $\mathbf{\Phi}(\mathbf{I}_K \otimes \mathcal{G})$. It is clear that $R \le 2K$ and the equality holds if the $2K$ signature vectors $\{\bar{\boldsymbol{\psi}}^{(k)}, \tilde{\boldsymbol{\psi}}^{(k)}\}_{k=1}^{K}$ are all linearly independent, where the signature vectors are defined as

$$\bar{\boldsymbol{\psi}}^{(k)} \triangleq \bar{\mathbf{\Phi}}^{(k)}\bar{\mathbf{g}} = \bar{\mathbf{\Phi}}^{(k)}\mathcal{F}_M\bar{\mathbf{h}}, \qquad \tilde{\boldsymbol{\psi}}^{(k)} \triangleq \tilde{\mathbf{\Phi}}^{(k)}\tilde{\mathbf{g}} = \tilde{\mathbf{\Phi}}^{(k)}\mathcal{F}_M\tilde{\mathbf{h}}. \tag{15}$$

Therefore, we have the following eigenvalue decomposition (EVD) of $\mathbf{R}_y$:

$$\mathbf{R}_y = \mathbf{\Gamma}_s\Lambda_s\mathbf{\Gamma}_s^H + \sigma_v^2\mathbf{\Gamma}_n\mathbf{\Gamma}_n^H, \tag{16}$$

where $\mathbf{\Lambda}_s = \operatorname{diag}\{\lambda_1, \ldots, \lambda_R\}$ contains the R largest eigenvalues, that is, *signal eigenvalues* of $\mathbf{R}_y$, $\mathbf{\Gamma}_s \in \mathbb{C}^{P\times R}$ contains the

R *signal eigenvectors* which span the *signal subspace*, that is, $\text{ran}\{\boldsymbol{\Gamma}_s\} = \text{ran}\{\boldsymbol{\Phi}(\mathbf{I}_K \otimes \mathcal{G})\}$, and $\boldsymbol{\Gamma}_n \in \mathbb{C}^{P\times(P-R)}$ contains the $(P-R)$ *noise eigenvectors* which span the *noise subspace*. The orthogonality between the noise and signal subspace implies that

$$\boldsymbol{\Gamma}_n^H \boldsymbol{\Phi}(\mathbf{I}_K \otimes \mathcal{G}) = \mathbf{0}. \tag{17}$$

For user k, the above equation is reduced to

$$\boldsymbol{\Gamma}_n^H \bar{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M \bar{\mathbf{h}} = \mathbf{0}, \tag{18}$$

$$\boldsymbol{\Gamma}_n^H \tilde{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M \tilde{\mathbf{h}} = \mathbf{0}. \tag{19}$$

Hence, $\bar{\mathbf{h}}$ and $\tilde{\mathbf{h}}$ are in the null space of $\boldsymbol{\Gamma}_n^H \bar{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M$ and $\boldsymbol{\Gamma}_n^H \tilde{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M$, respectively. If the nullity of these matrices is one, then $\bar{\mathbf{h}}$ and $\tilde{\mathbf{h}}$ can be uniquely and perfectly (up to a scalar) solved from (18) and (19), respectively. The residual scalar ambiguity can be resolved by transmitting a few pilots, similarly to the semiblind schemes developed in [37, 38]. The key idea behind such a semiblind approach is that, in order to resolve the scalar ambiguity, only *a few* pilot symbols are needed, thus leading to a higher spectral efficiency compared to a full training scheme which has to estimate all channel coefficients by training.

The remaining question is under what conditions the nullity of $\boldsymbol{\Gamma}_n^H \bar{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M$ and $\boldsymbol{\Gamma}_n^H \tilde{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M$ is one so that identifiability is guaranteed.

3.2. Blind identifiability

To determine the identification conditions, we first note that the dimensions of $\boldsymbol{\Gamma}_n^H \bar{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M$ and $\boldsymbol{\Gamma}_n^H \tilde{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M$ are $(P-R)\times M$. Since for any $L \times M$ matrix $\mathbf{X}$, $\dim(\text{null}(\mathbf{X})) + \text{rank}(\mathbf{X}) = M$ [29], it is necessary to have

(A2) $P - R \geq M - 1$.

Otherwise, the nullity of both $\boldsymbol{\Gamma}_n^H \bar{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M$ and $\boldsymbol{\Gamma}_n^H \tilde{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M$ will always be greater than one. When the signature vectors $\{\bar{\boldsymbol{\psi}}^{(k)}, \tilde{\boldsymbol{\psi}}^{(k)}\}_{k=1}^K$ are linearly independent of each other (e.g., with independent spreading codes plus mild conditions on the channel), (A2) imposes an upper limit on the number of users allowed in the system

$$K \leq \frac{P - M + 1}{2}. \tag{20}$$

Next, we present the identifiability conditions. Define $\bar{\boldsymbol{\Psi}} \triangleq [\bar{\boldsymbol{\psi}}^{(1)}, \ldots, \bar{\boldsymbol{\psi}}^{(K)}]$ and $\tilde{\boldsymbol{\Psi}} \triangleq [\tilde{\boldsymbol{\psi}}^{(1)}, \ldots, \tilde{\boldsymbol{\psi}}^{(K)}]$ which collect all signature vectors. Let $\bar{\boldsymbol{\Psi}}^{(k)} \in \mathbb{C}^{P\times(K-1)}$ and $\tilde{\boldsymbol{\Psi}}^{(k)} \in \mathbb{C}^{P\times(K-1)}$ be obtained from $\bar{\boldsymbol{\Psi}}$ and $\tilde{\boldsymbol{\Psi}}$, respectively, by deleting the kth column. The following result specifies conditions which guarantee unique and perfect (up to a scalar) estimation of the channel from (18) and (19).

Proposition 1 (Identifiability). *In addition to conditions (A1) and (A2), suppose the following are also satisfied:*

(A3) $\bar{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M$ *and* $\tilde{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M$ *have full column rank;*
(A4) *the following ranges do not intersect:*

$$\text{ran}\,(\bar{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M) \cap \text{ran}\,([\bar{\boldsymbol{\Psi}}^{(k)}, \tilde{\boldsymbol{\Psi}}]) = \varnothing, \tag{21}$$

$$\text{ran}\,(\tilde{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M) \cap \text{ran}\,([\tilde{\boldsymbol{\Psi}}^{(k)}, \bar{\boldsymbol{\Psi}}]) = \varnothing. \tag{22}$$

Then, channel vectors $\bar{\mathbf{h}}$ *and* $\tilde{\mathbf{h}}$ *are uniquely and perfectly (up to a scalar) determined by the solutions of* (18) *and* (19), *respectively.*

Proof. In the sequel, we will only prove that the solution of (18) is unique and perfect. The uniqueness and perfectness of the solution of (19) follow similarly.

The proof of uniqueness goes by contradiction. Assume that there exists another solution of (18)

$$\boldsymbol{\Gamma}_n^H \bar{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M \bar{\mathbf{h}}' = \mathbf{0}. \tag{23}$$

Then, $\bar{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M \bar{\mathbf{h}}' \in \text{ran}([\bar{\boldsymbol{\Psi}}, \tilde{\boldsymbol{\Psi}}])$. That is, there exists nontrivial $[\bar{\boldsymbol{\alpha}}^T, \tilde{\boldsymbol{\alpha}}^T]^T \in \mathbb{C}^{2K\times 1}$ such that

$$\bar{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M \bar{\mathbf{h}}' = \sum_{l=1}^{K} \bar{\alpha}_l \bar{\boldsymbol{\psi}}^{(l)} + \tilde{\boldsymbol{\Psi}} \tilde{\boldsymbol{\alpha}}, \tag{24}$$

where $\bar{\alpha}_l$ denotes the lth element of $\bar{\boldsymbol{\alpha}}$. Since $\bar{\boldsymbol{\psi}}^{(k)} = \bar{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M \bar{\mathbf{h}}$, we rearrange the above equation as follows:

$$\bar{\boldsymbol{\Phi}}^{(k)} \mathcal{F}_M (\bar{\mathbf{h}}' - \alpha_k \bar{\mathbf{h}}) = \sum_{l\neq k} \bar{\alpha}_l \bar{\boldsymbol{\psi}}^{(l)} + \tilde{\boldsymbol{\Psi}} \tilde{\boldsymbol{\alpha}} = \mathbf{0}, \tag{25}$$

where the last equality follows from (21), which necessitates $\tilde{\boldsymbol{\alpha}} = \mathbf{0}$ and $\bar{\alpha}_l = 0$, for $l \neq k$. Meanwhile, assumption (A3) implies that

$$\bar{\mathbf{h}}' = \bar{\alpha}_k \bar{\mathbf{h}}. \tag{26}$$

That is, $\bar{\mathbf{h}}'$ must be collinear with $\bar{\mathbf{h}}$ which proves the uniqueness (up to a scalar).

The perfectness of the solution follows since the true channel vector $\bar{\mathbf{h}}$ is a solution of (18). ☐

We note that assumption (A1) may be relaxed as long as $E\{\mathbf{d}(n)\mathbf{d}^H(n)\}$ has full rank. This is essential for subspace decomposition. Without it, in general, $\text{ran}(\Gamma_s) \neq \text{ran}\{\boldsymbol{\Phi}(\mathbf{I}_K \otimes \mathcal{G})\}$ which makes it impossible to separate the signal and noise subspace. Assumption (A1) is satisfied in practice because information symbols from different users are usually independent of each other. Assumption (A2), as mentioned before, sets a limit on the maximum number of users that can be supported in the system. Assumption (A3) is violated only if spreading codes with at least one zero element are utilized. Since such spreading codes are seldom used in real systems, assumption (A3) is not a restrictive assumption. Assumption (A4) implies that to ensure identifiability, the signatures of all interfering signals, including not only all interfering users but also the interfering signal of the desired user from the transmit antenna which is not being estimated, do not reside in the range of $\bar{\boldsymbol{\Phi}}^{(k)}\boldsymbol{\mathcal{F}}_M$ (or $\tilde{\boldsymbol{\Phi}}^{(k)}\boldsymbol{\mathcal{F}}_M$). We found that with standard spreading codes (e.g., bipolar spreading codes that are linearly independent of one another) and mild conditions on the channel, this assumption is usually satisfied.

3.3. Implementation issues

In practice, $\boldsymbol{\Gamma}_{\mathrm{n}}$ is unknown and needs to be computed from some estimate of $\mathbf{R}_y$. While other estimates are possible, a simple one based on block processing is the sample covariance matrix estimate

$$\hat{\mathbf{R}}_y = \frac{1}{N_y}\sum_{n=0}^{N_y-1}\mathbf{y}(n)\mathbf{y}^H(n), \tag{27}$$

where N_y denotes the block size. An estimate of the noise eigenvectors $\hat{\boldsymbol{\Gamma}}_{\mathrm{n}}$ can be obtained from the EVD of $\hat{\mathbf{R}}_y$. Due to (finite-sample) estimation errors in $\hat{\boldsymbol{\Gamma}}_{\mathrm{n}}$, (18) and (19) will not hold exactly. In this case, we can solve it in the least squares (LS) sense by minimizing the two norm of the vectors on the left side of (18) and (19), respectively.

In some applications, it is also possible that the receiver may have knowledge of some other users' spreading codes (informed by the base station through a control channel). When such additional information is available, it is beneficial to incorporate it in the channel estimation, which in general improves estimation accuracy. To do so, we may utilize the following criteria:

$$\hat{\hat{\mathbf{h}}} = \arg\min_{\mathbf{h}\in\mathbb{C}^{M\times 1}} \mathbf{h}^H\left\{\sum_k \boldsymbol{\mathcal{F}}_M^H\bar{\boldsymbol{\Phi}}^{(k)H}\hat{\boldsymbol{\Gamma}}_{\mathrm{n}}\hat{\boldsymbol{\Gamma}}_{\mathrm{n}}^H\bar{\boldsymbol{\Phi}}^{(k)}\boldsymbol{\mathcal{F}}_M\right\}\mathbf{h}, \tag{28}$$

$$\hat{\hat{\mathbf{h}}} = \arg\min_{\mathbf{h}\in\mathbb{C}^{M\times 1}} \mathbf{h}^H\left\{\sum_k \boldsymbol{\mathcal{F}}_M^H\tilde{\boldsymbol{\Phi}}^{(k)H}\hat{\boldsymbol{\Gamma}}_{\mathrm{n}}\hat{\boldsymbol{\Gamma}}_{\mathrm{n}}^H\tilde{\boldsymbol{\Phi}}^{(k)}\boldsymbol{\mathcal{F}}_M\right\}\mathbf{h}, \tag{29}$$

where the summation (averaging) is with respect to the set of users whose spreading codes are known. When the nullity of the matrices within the curly brackets of (22) and (23), respectively, is greater than one, the channel becomes unidentifiable. It is interesting to note that in such a case, knowing spreading codes of additional users allows further averaging, which helps building up the rank of those matrices and may render the channel identifiable. That is, knowing more spreading codes can relax the identifiability conditions. Although a strict analysis appears difficult, if not impossible, this has been observed in our simulations.

Equations (27), (16), and (28), (29), in that order, summarize the proposed subspace blind channel estimator for the STC-MC-CDMA system. The above block-processing based implementation is, however, known to be computation-wise and complexity-wise inefficient. All subspace-based schemes rely on an estimate of the signal and/or noise eigenvectors. In practice, these eigenvectors can be obtained with significantly reduced complexity by adaptive subspace tracking methods (see, e.g., [39, 40, 41] and references therein). The subspace tracking method in [41], for example, estimates only one noise subspace vector at a time and enjoys such desired properties as fast convergence, no need for exact rank information, and estimating directly the noise subspace without tracking the signal subspace (which is not needed in our channel identification scheme). It should be noted that most of the above subspace tracking algorithms can be applied to the current STC-MC-CDMA systems with little modification. The only notable change is that because of OFDM modulation, noise subspace tracking is now performed in the frequency domain as opposed to the conventional time-domain subspace tracking. Due to space limitation, we will not pursue this topic further in this paper. Interested readers are referred to the above cited references for details about standard subspace tracking methods.

4. RECEIVER DESIGN

In conventional MC-CDMA systems (without transmit diversity), the OFDM demodulated signals are often combined in the frequency domain in order to collect the overall received signal energy scattered on different subcarriers. Typical signal combining schemes include the maximum ratio combining (MRC) and the equal gain combining (EGC) [9, 11]. In this section, we extend these combining schemes to STC-MC-CDMA systems to perform joint combining and decoding. Both MRC and EGC are *single-user* detection schemes based on per-subcarrier combining, that is, the signals at individual subcarrier are independently weighted and summed to generate decision variables. In what follows, we also present a linear multiuser MMSE detector which jointly weights and combines signals on all subcarriers.

4.1. Single-user coherent signal combining and decoding

Without loss of generality, let the first user be the desired user. We define the OFDM demodulated signals corresponding to the pth subcarrier in two consecutive time slots $2n$ and $2n+1$ as $\mathbf{y}_p(n) \triangleq [y_p(2n), y_p^*(2n+1)]^T$, where

$$\begin{aligned}
y_p(2n) \triangleq{}& [\bar{g}(p)\bar{\phi}_p^{(1)}\bar{d}^{(1)}(2n) + \tilde{g}(p)\tilde{\phi}_p^{(1)}\tilde{d}^{(1)}(2n)] \\
&+ \sum_{k=2}^{K}[\bar{g}(p)\bar{\phi}_p^{(k)}\bar{d}^{(k)}(2n) + \tilde{g}(p)\tilde{\phi}_p^{(k)}\tilde{d}^{(k)}(2n)] \\
&+ v_p(2n), \\
y_p(2n+1) \triangleq{}& [\bar{g}(p)\bar{\phi}_p^{(1)}\bar{d}^{(1)}(2n+1) + \tilde{g}(p)\tilde{\phi}_p^{(1)}\tilde{d}^{(1)}(2n+1)] \\
&+ \sum_{k=2}^{K}[\bar{g}(p)\bar{\phi}_p^{(k)}\bar{d}^{(k)}(2n+1) + \tilde{g}(p)\tilde{\phi}_p^{(k)}\tilde{d}^{(k)}(2n+1)] \\
&+ v_p(2n+1),
\end{aligned} \tag{30}$$

and we note that the second term of each equation represents the MUI in the system. Using the ST signal mapping given in (12), we can rewrite (30) as follows:

$$\begin{aligned}
y_p(2n) \triangleq{}& [\bar{g}(p)\bar{\phi}_p^{(1)} b^{(1)}(2n) + \tilde{g}(p)\tilde{\phi}_p^{(1)} b^{(1)}(2n+1)] \\
&+ \sum_{k=2}^{K} [\bar{g}(p)\bar{\phi}_p^{(k)} b^{(k)}(2n) + \tilde{g}(p)\tilde{\phi}_p^{(k)} b^{(k)}(2n+1)] \\
&+ v_p(2n), \\
y_p^*(2n+1) \triangleq{}& [\tilde{g}^*(p)\tilde{\phi}_p^{(1)} b^{(1)}(2n) - \bar{g}^*(p)\bar{\phi}_p^{(1)} b^{(1)}(2n+1)] \\
&+ \sum_{k=2}^{K} [\tilde{g}^*(p)\tilde{\phi}_p^{(k)} b^{(k)}(2n) - \bar{g}^*(p)\bar{\phi}_p^{(k)} b^{(k)}(2n+1)] \\
&+ v_p^*(2n+1).
\end{aligned} \tag{31}$$

In the sequel, we present the MRC and the EGC coherent signal combining schemes for STC-MC-CDMA systems, both treating the MUI in (31) as noise.

4.1.1 Maximum ratio combining

Incorporating the Alamouti decoding [21] while maximizing the output SNR, we obtain the MRC combiner applied on subcarrier p for the detection of $b^{(1)}(2n)$ and $b^{(1)}(2n+1)$, respectively, as follows:

$$\begin{aligned}
\boldsymbol{\alpha}_{p,\mathrm{MRC}}^{(1)} &\triangleq [\bar{g}^*(p)\bar{\phi}_p^{(1)}, \tilde{g}(p)\tilde{\phi}_p^{(1)}]^T, \\
\boldsymbol{\beta}_{p,\mathrm{MRC}}^{(1)} &\triangleq [\tilde{g}^*(p)\tilde{\phi}_p^{(1)}, -\bar{g}(p)\bar{\phi}_p^{(1)}]^T.
\end{aligned} \tag{32}$$

The decision variables are obtained by combining the MRC outputs from all subcarriers

$$\begin{aligned}
\hat{b}^{(1)}(2n) &= \sum_{p=0}^{P-1} \boldsymbol{\alpha}_{p,\mathrm{MRC}}^{(1)T} \mathbf{y}_p(n), \\
\hat{b}^{(1)}(2n+1) &= \sum_{p=0}^{P-1} \boldsymbol{\beta}_{p,\mathrm{MRC}}^{(1)T} \mathbf{y}_p(n).
\end{aligned} \tag{33}$$

4.1.2 Equal gain combining

Likewise, applying the Alamouti decoding along with (coherent) EGC yields

$$\begin{aligned}
\boldsymbol{\alpha}_{p,\mathrm{EGC}}^{(1)} &\triangleq \left[\frac{\bar{g}^*(p)}{g(p)}\bar{\phi}_p^{(1)}, \frac{\tilde{g}(p)}{g(p)}\tilde{\phi}_p^{(1)}\right]^T, \\
\boldsymbol{\beta}_{p,\mathrm{EGC}}^{(1)} &\triangleq \left[\frac{\tilde{g}^*(p)}{g(p)}\tilde{\phi}_p^{(1)}, -\frac{\bar{g}(p)}{g(p)}\bar{\phi}_p^{(1)}\right]^T,
\end{aligned} \tag{34}$$

where $g(p) \triangleq \sqrt{|\bar{g}(p)|^2 + |\tilde{g}(p)|^2}$. The corresponding decision variables are

$$\begin{aligned}
\hat{b}^{(1)}(2n) &= \sum_{p=0}^{P-1} \boldsymbol{\alpha}_{p,\mathrm{EGC}}^{(1)T} \mathbf{y}_p(n), \\
\hat{b}^{(1)}(2n+1) &= \sum_{p=0}^{P-1} \boldsymbol{\beta}_{p,\mathrm{EGC}}^{(1)T} \mathbf{y}_p(n).
\end{aligned} \tag{35}$$

Except for the incorporation of the Alamouti decoding, the MRC and EGC vectors in (32) and (34) are similar to their counterparts for conventional MC-CDMA systems without ST coding [9, 11]. Previous studies have found that these single-user schemes are sensitive to the MUI [11]; also see the numerical results in Section 5. The suboptimal performance of these single-user schemes motivates the pursuit of more sophisticated multiuser detection schemes, such as the linear MMSE detector which is discussed next.

4.2. Linear MMSE multiuser detector

Recently, linear MMSE detection has been considered for MC-CDMA systems without ST coding [42, 43]. In the sequel, we extend it to the current STC-MC-CDMA systems to perform joint combining and decoding over all subcarriers. To facilitate our presentation, we rewrite (10) in a slightly different but equivalent form

$$\mathbf{y}(n) = \bar{\mathbf{G}}\bar{\boldsymbol{\Psi}}\bar{\mathbf{d}}(n) + \tilde{\mathbf{G}}\tilde{\boldsymbol{\Psi}}\tilde{\mathbf{d}}(n) + \mathbf{v}(n), \tag{36}$$

where

$$\begin{aligned}
\bar{\mathbf{G}} &\triangleq \operatorname{diag}\{\bar{g}(0), \dots, \bar{g}(P-1)\}, \\
\tilde{\mathbf{G}} &\triangleq \operatorname{diag}\{\tilde{g}(0), \dots, \tilde{g}(P-1)\}, \\
\bar{\boldsymbol{\Psi}} &\triangleq [\bar{\boldsymbol{\phi}}^{(1)}, \dots, \bar{\boldsymbol{\phi}}^{(K)}], \\
\tilde{\boldsymbol{\Psi}} &\triangleq [\tilde{\boldsymbol{\phi}}^{(1)}, \dots, \tilde{\boldsymbol{\phi}}^{(K)}], \\
\bar{\mathbf{d}}(n) &\triangleq [\bar{d}^{(1)}(n), \dots, \bar{d}^{(K)}(n)]^T, \\
\tilde{\mathbf{d}}(n) &\triangleq [\tilde{d}^{(1)}(n), \dots, \tilde{d}^{(K)}(n)]^T.
\end{aligned} \tag{37}$$

We consider two OFDM demodulated symbols at a time. Using again the signal mapping in (12), we have

$$\begin{aligned}
\mathbf{y}(2n) &= \bar{\mathbf{G}}\bar{\boldsymbol{\Psi}}\mathbf{b}(2n) + \tilde{\mathbf{G}}\tilde{\boldsymbol{\Psi}}\mathbf{b}(2n+1) + \mathbf{v}(2n), \\
\mathbf{y}(2n+1) &= -\bar{\mathbf{G}}\bar{\boldsymbol{\Psi}}\mathbf{b}^*(2n+1) + \tilde{\mathbf{G}}\tilde{\boldsymbol{\Psi}}\mathbf{b}^*(2n) + \mathbf{v}(2n+1).
\end{aligned} \tag{38}$$

Let $\mathbf{z}(n) \triangleq [\mathbf{y}^T(2n), \mathbf{y}^H(2n+1)]^T$. Then, we have (assuming real-valued spreading codes)

$$\mathbf{z}(n) = \begin{bmatrix} \bar{\mathbf{G}}\bar{\boldsymbol{\Psi}} & \tilde{\mathbf{G}}\tilde{\boldsymbol{\Psi}} \\ \tilde{\mathbf{G}}^*\tilde{\boldsymbol{\Psi}} & -\bar{\mathbf{G}}^*\bar{\boldsymbol{\Psi}} \end{bmatrix} \begin{bmatrix} \mathbf{b}(2n) \\ \mathbf{b}(2n+1) \end{bmatrix} + \begin{bmatrix} \mathbf{v}(2n) \\ \mathbf{v}^*(2n+1) \end{bmatrix}. \tag{39}$$

Let $\mathbf{b}^{(1)}(n) = [b^{(1)}(2n), b^{(1)}(2n+1)]^T$ which contains the information symbols of the first user at time $2n$ and $2n+1$. The MMSE receiver minimizes the following mean square error (MSE) criterion (e.g., [44]):

$$\mathcal{W}_{\mathrm{MMSE}}^{(1)} = \arg \min_{\mathcal{W} \in \mathbb{C}^{2P\times 2}} E\left\{ \left\| \mathcal{W}^H \mathbf{z}(n) - \mathbf{b}^{(1)}(n) \right\|^2 \right\}, \tag{40}$$

the solution of which is given by

$$\mathcal{W}_{\mathrm{MMSE}}^{(1)} = \mathbf{R}_z^{-1}\mathbf{R}_{zb}^{(1)}, \tag{41}$$

where $\mathbf{R}_z \triangleq E\{\mathbf{z}(n)\mathbf{z}^H(n)\}$ and $\mathbf{R}_{zb}^{(1)}$ is the cross covariance between $\mathbf{z}(n)$ and $\mathbf{b}^{(1)}(n)$,

$$\mathbf{R}_{zb}^{(1)} \triangleq E\{\mathbf{z}(n)\mathbf{b}^{(1)H}(n)\} = \begin{bmatrix} \bar{\mathbf{G}}\bar{\boldsymbol{\phi}}^{(1)} & \tilde{\mathbf{G}}\tilde{\boldsymbol{\phi}}^{(1)} \\ \tilde{\mathbf{G}}^*\tilde{\boldsymbol{\phi}}^{(1)} & -\bar{\mathbf{G}}^*\bar{\boldsymbol{\phi}}^{(1)} \end{bmatrix}. \quad (42)$$

The decision variables yielded by the linear MMSE detector are given by

$$\hat{\mathbf{b}}^{(1)}(n) \triangleq \left[\hat{b}^{(1)}(2n), \hat{b}^{(1)}(2n+1)\right]^T = \mathcal{W}_{\mathrm{MMSE}}^{(1)H}\mathbf{z}(n). \quad (43)$$

Note that the covariance matrix $\mathbf{R}_z$ required by the MMSE receiver can be computed either based on block processing or adaptively from the received data. Hence, similar to the MRC and EGC schemes, the MMSE receiver requires only the desired user's spreading codes for detection. As we will see in Section 5.2, the MMSE receiver yields improved performance over the MRC and EGC receivers. This is achieved at the cost of increased complexity. Specifically, while both the MRC and EGC have a linear complexity $\mathcal{O}(P)$, a direct, block-processing based implementation of the MMSE receiver results in a complexity of $\mathcal{O}(P^3)$. Adaptive implementation of the MMSE receiver with some decreased complexity is possible [44]. Even so, the adaptive MMSE receiver would incur more computations than the MRC and EGC receivers.

Finally, the hard estimate $\hat{\hat{b}}^{(1)}(n)$ is obtained by comparing the decision variable $\hat{b}^{(1)}(n)$, obtained by any of the above detection/combining schemes, with every constellation point

$$\hat{\hat{b}}^{(1)}(n) = \arg\min_{b\in\mathcal{B}} \left|\hat{b}^{(1)}(n) - b\right|. \quad (44)$$

5. SIMULATION RESULTS

In this section, we present simulation results to evaluate the performance of the proposed blind channel identification algorithm and the signal detection/combining schemes. In our simulation, user symbols are drawn from a unit-energy BPSK (binary phase shift keying) constellation. Walsh-Hadamard codes with processing gain $P = Q = 32$ are used for spreading. We assume a rich scattering environment and generate the FIR channel coefficients $\{\bar{h}(m)\}_{m=0}^{M-1}$ and $\{\tilde{h}(m)\}_{m=0}^{M-1}$ as i.i.d. complex Gaussian random variables with zero-mean and variance $1/M$. The SNR is defined as $\mathrm{SNR} = 10\log_{10} 1/\sigma_v^2$ in dB.

5.1. Channel identification

The figure of merit used for evaluating the performance of channel identification is the averaged MSE defined as

$$\mathrm{MSE}\,(\hat{\bar{\mathbf{h}}}) = \left(\frac{1}{M-1}\right)\sum_m \mathrm{MSE}\,\{\hat{\bar{h}}(m)\}, \quad (45)$$

where $\hat{\bar{h}}(m)$ denotes the estimate of $\bar{h}(m)$; $\mathrm{MSE}(\hat{\tilde{\mathbf{h}}})$ is similarly defined. The channel estimates $\hat{\bar{\mathbf{h}}}$ and $\hat{\tilde{\mathbf{h}}}$ are normalized with respect to the first element of $\bar{\mathbf{h}}$ and $\tilde{\mathbf{h}}$, respectively, due to the inherent scalar ambiguity in all blind channel estimates. As mentioned before, the residual scalar ambiguity can be removed by differential coding or utilizing (a few) pilot symbols (e.g., [37, 38]). We consider two cases involving $K = 5$ and $K = 10$ users, respectively. A total of 200 independent Monte Carlo trials were conducted to obtain the averaged MSE. Figure 2a depicts the averaged MSE versus the SNR when $N_y = 250$ data samples are used to compute the sample covariance matrix $\hat{\mathbf{R}}_y$ (see (27)) whereas Figure 2b shows the averaged MSE versus N_y when SNR = 20 dB. In both cases, we see that the accuracy of the channel estimates is improved as the SNR and/or the number of data samples increases.

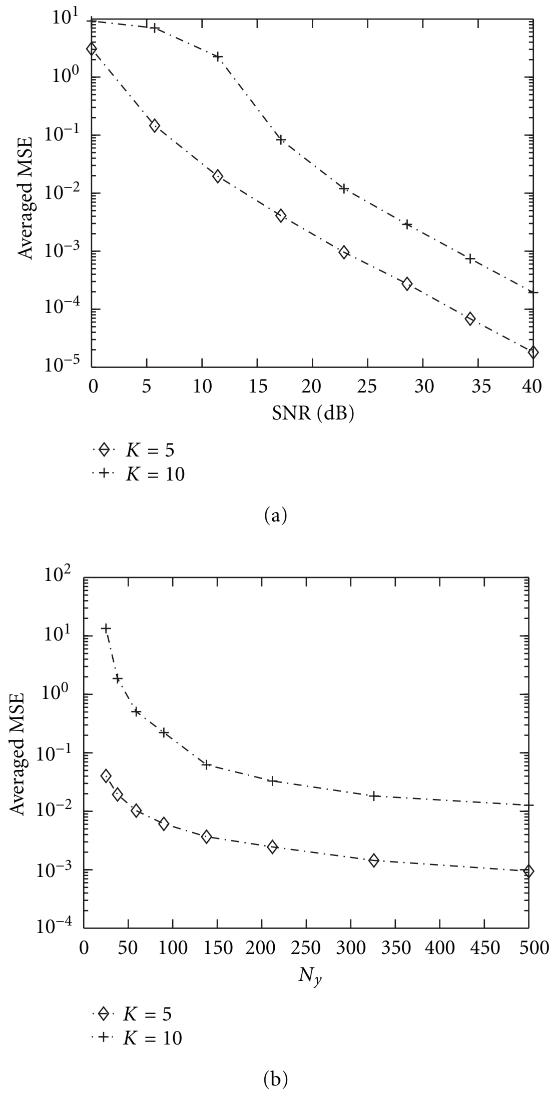
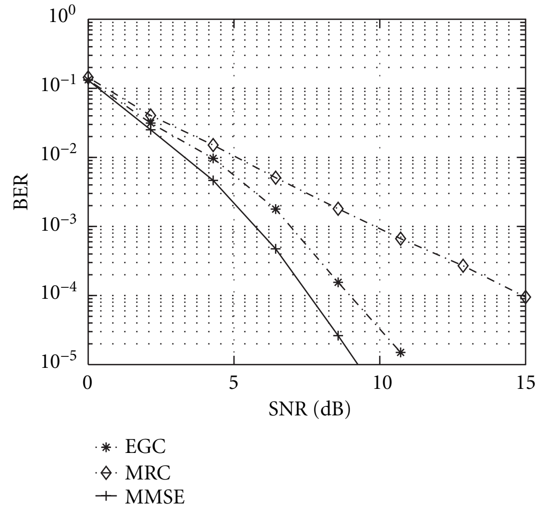


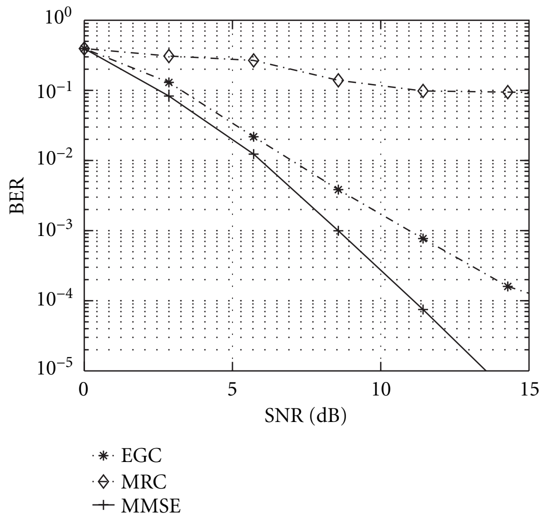
FIGURE 2: Averaged MSE of channel estimation versus (a) SNR when $N_y = 250$ and (b) N_y when SNR = 20 dB.

5.2. Signal detection

Next we examine the performance of the two single-user signal combining schemes, MRC and EGC, and the linear MMSE multiuser detection scheme introduced in Section 4. Channel estimates required by these schemes are obtained by



(a)



(b)

Figure 3: BER versus SNR of the MRC, EGC, and MMSE schemes when (a) $K = 5$ and (b) $K = 10$.

the proposed channel identification algorithm, using $N_y = 500$ samples of data. The BER results presented in the following are averaged over 500 independent channel realizations to emulate a Rayleigh fading environment.

The BER of the three signal detection/combining schemes as a function of the SNR is depicted in Figure 3a when $K = 5$ and Figure 3b when $K = 10$, respectively. The MRC scheme is known to be optimal when there is only one user in the system. It is very sensitive to the presence of MUI which degrades its performance [9]. A comparison of Figure 3a ($K = 5$ users) with Figure 3b ($K = 10$ users) shows that the degradation of MRC becomes more severe as the number of users increases. The EGC and the MMSE receivers are observed to outperform the MRC scheme for both cases, with the MMSE receiver being the best.

The relative performances of the three detection schemes for STC-MC-CDMA systems are similar to those reported in [11] for conventional MC-CDMA systems (i.e., no ST coding) with orthogonal spreading codes. It should be noted that there is no analytical evidence that EGC may always outperform MRC or vice versa. Both are single-user based detection schemes, and may suffer from MUI. The relative performance depends on such factors as spreading codes, user powers, and the underlying channel. When orthogonal spreading codes are used, our simulation results and those in [11] appear to suggest that EGC may be better than MRC in restoring the user orthogonality, to which the performance gain is attributed.

5.3. *Diversity gain*

Finally, we examine the diversity gain offered by ST coding in Rayleigh fading environments. In particular, we compare the BER of the STC-MC-CDMA system and a conventional single-transmit-antenna based MC-CDMA system without ST coding. The transmitted power for the conventional MC-CDMA system is twice that of the STC-MC-CDMA system to ensure that the SNR in both systems are comparable. The three detection/combining schemes, namely MRC, EGC, and MMSE, are employed for detection in the two systems, utilizing both true and estimated channel coefficients. In the latter case, the subspace-based blind channel identification algorithm is used for channel estimation. Figure 4 depicts the BER versus the SNR for the two systems. We note that STC-MC-CDMA performs considerably better than the conventional MC-CDMA system, yielding significant diversity gain (even with estimated channels). We also note that channel estimation for STC-MC-CDMA is less accurate than the single-transmit-antenna based MC-CDMA system, particularly when the input SNR is relatively low. This is not surprising since the number of unknown channel coefficients doubles in STC-MC-CDMA. As the SNR increases, the performance difference between using true and estimated channels is noted to be less than 1 dB for all three detectors.

6. CONCLUSIONS

In this paper, we have presented a subspace-based blind channel identification algorithm for STC-MC-CDMA systems. We have investigated the associated identifiability problem and specified conditions which guarantee unique and perfect (within a scalar ambiguity) channel identification by the proposed algorithm. For receiver design, we have extended the conventional MRC and EGC schemes for single-transmit-antenna based MC-CDMA systems (without ST coding) to STC-MC-CDMA systems. We have also introduced a linear multiuser MMSE detection scheme that performs joint combining and multiuser detection. Simulation results for the proposed channel identification and signal detection/combining schemes have been presented. The diversity advantage of STC-MC-CDMA systems over conventional MC-CDMA systems has also been demonstrated.

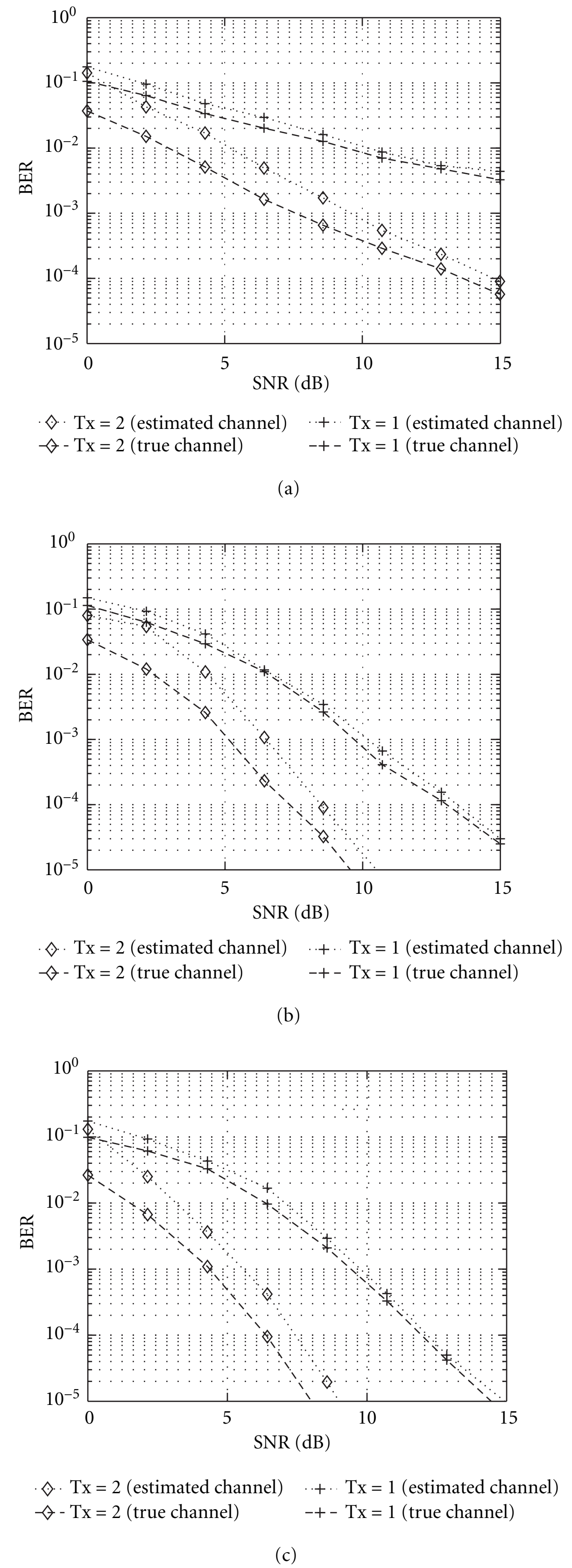


Figure 4: BER versus SNR when $K = 5$. (a) MRC detector; (b) EGC detector; (c) MMSE detector.

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