Joint Channel Parameter Estimation and Signal Detection for Downlink MIMO DS-CDMA Systems*

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SUMMARY This paper proposes two space-time joint channel parameter estimation and signal detection algorithms for downlink DS-CDMA systems with multiple-input-multiple-output (MIMO) wireless multipath fading channels. The proposed algorithms initially use the space-time MUSIC to estimate the DOA-delays of the multipath channel. Based on these estimated DOA-delays, a space-time channel decoupler is developed to decompose the multipath downlink channel into a set of independent parallel subchannels. The fading amplitudes of the multipath can then be estimated from the eigen space of the output of the space-time channel decoupler. With these estimated channel parameters, signal detection is carried out by a maximal ratio combiner on a pathwise basis. Computer simulations show that the proposed algorithms outperform the conventional space-time RAKE receiver while having the similar performance compared with the space-time minimum mean square error receiver.

key words: signal detection, DS-CDMA, MIMO, downlink channels, software radio

1. Introduction

attention in recent years due to their capability of integrating different communication standards to provide diverse communication services on a software platform [1], [2]. Also, it is one of the major goals for future wireless communications engineering, especially for 3-G or beyond in which high transmission data rate and multiuser communication services are indispensable. With antenna arrays employed at both the transmitter and the receiver, multiple-input-multiple-output direct sequence code division multiple access (MIMO-DS-CDMA) systems are emerging to achieve high transmission data rate in a multiuser system.

Diversity techniques are often adopted to enhance the channel reliability in wireless MIMO systems. Several space-time codecs, such as the BLAST [3] architecture and various space-time trellis coding schemes [4], have been developed to pursue the performance limit of MIMO systems. However, almost all these pioneering works assume that the channel state information (CSI) is perfectly estimated (or tracked) and is known to the transmitters and/or the receivers. With a large amount of high-dimensional CSI existing in a MIMO system, simple and accurate channel estimation becomes crucial for high data rate wireless systems.

Radio signals are often characterized by multipath propagation in wireless communication channels. A multipath channel can be parameterized by propagation directions, propagation delays, and fading amplitudes of multipath, due to signal reflection and diffraction from terrestrial objects and mobility of channel environments. Various algorithms for the space-time parameter estimation have been proposed to estimate the parameters of a wireless channel [6]–[11]. Among them, [6], [7] develop a series of MUSIC-based algorithms to estimate the path propagation delays. [8] propose to utilize the Root-MUSIC and some modified mean square error (MSE) methods, in which a preamble is sent periodically from the transmitter, to jointly estimate the path delays and carrier phases of the wireless channel. However, [6], [7] assume that each user possesses only a single propagation path, which is not practical for wireless applications. On the other hand, [8] employ a preamble sequence to train the receiver so that the frequency efficiency is degraded. In addition, [9]–[11], due to restricting to single user scenarios, they are not applicable to the multiuser scenarios.

This paper presents a class of blind space-time joint channel parameter estimation and signal detection algorithms for downlink MIMO DS-CDMA systems with multipath fading channels. The algorithms are blind in a sense that, for a downlink DS-CDMA channel, the only information known to each mobile receiver is its own spreading code. The proposed algorithms first exploit the space-time MUSIC algorithm (ST-MUSIC) for direction of arrivals (DOAs) and propagation delays joint estimation. Then, based on the DOA-delay estimates, a space-time channel decoupler is developed to suppress the multiple access interference (MAI) and to isolate each independent multipath simultaneously. As a result, the multipath channel is decomposed into a set of independent parallel subchannels. Taking over the subchannel signals, fading amplitudes can be estimated as the significant eigen vector of the covariance matrix of the outputs of the space-time channel decoupler. Thereafter, signal detection can be performed by a maximal ratio combiner. According to different multiple access interference (MAI) suppression techniques, two space-time signal detection algorithms referred to as space-time complementary orthogonal projection (ST-COP) and the space-time minimum variance distortionless response (ST-MVDR) are
proposed, respectively.

The rest of this paper is organized as follows. Section 2 introduces the parameterized system model for downlink MIMO DS-CDMA systems with multipath fading channels. Section 3 describes the proposed algorithms. Section 4 conducts some computer simulations to assess the proposed algorithms, and Sect. 5 concludes the paper.

2. System Model

A downlink MIMO DS-CDMA system with $M_r$ transmit antennas and $M_R$ receive antennas is shown in Fig. 1. At the base station, each transmit antenna simultaneously sends $K$ data streams at rate $\frac{1}{T}$ (symbols/sec), respectively, for $K$ users. The baseband signal of the $m$th transmit antenna is thus given by

$$z_m(t) = \sum_{i=-\infty}^{\infty} \sum_{k=1}^{K} d_{mk}(i) \sum_{n=0}^{N-1} s_{mk}(n) g(t - (iN + n) T_c),$$

(1)

where $d_{mk}(i)$ denotes the $i$th transmit data symbol for the $k$th user of the $m$th transmit antenna, $\{s_{mk}(0), \ldots, s_{mk}(N-1)\}$ is the $N$-chip spreading sequence identical for all data symbol of the $k$th user of the $m$th transmit antenna, $N$ is the spreading gain, $T_c = \frac{1}{T}$ is the DS-CDMA chip period, and $g(t)$ is the normalized pulse-shaping waveform. The support of $g(t)$ plus the maximum delay of the multipaths is assumed to extend an $L$-chip period, and $g(t) = 0$ if $t < 0$. Furthermore, throughout this paper, we assume the data symbol $d_{mk}(i)$ is $M$-PSK ($M$-phase shift keying) differentially-encoded. Sampled $\rho$ times per chip period $T_c$, the $q$th sample of the $i$th symbol in (1) can be represented as

$$z_m(\frac{qT_c}{\rho}) = \sum_{k=1}^{K} d_{mk}(i) \sum_{n=0}^{\rho N-1} s'_{mk}(n) g\left((q - n) \frac{T_c}{\rho}\right),$$

(2)

where $q' = (iN\rho + q)$, $\{s'_{mk}(n)\}$ denotes the interpolated version of the spreading sequence $\{s_{mk}(n)\}$ with $\rho - 1$ zeros inserted between $s_{mk}(n)$ and $s_{mk}(n-1)$. The summation $\sum_{n=0}^{\rho N-1} s'_{mk}(n) g\left((q - n) \frac{T_c}{\rho}\right)$ above is equivalent to the discrete time convolution between the spreading sequence $\{s'_{mk}(n)\}$ and the sampled pulse-shaping waveform $g\left((n \frac{T_c}{\rho})\right)$. By collecting samples of $Q$ symbols periods, the discrete time baseband signal of the $m$th transmit antenna, starting from the $i$th symbol, can be expressed as

$$z_m^Q(i) = \begin{bmatrix} z_m(iN T_c) & z_m((iN + 1) T_c) & \ldots, & z_m((i + Q - 1) N - \frac{1}{\rho} T_c) \end{bmatrix} = g^T(0) S_{mk} D_{mk}(i)$$

(3)

where the superscript $(\cdot)^T$ denotes the matrix transpose operation, and the sampled temporal vector $g(t)$ of size $L_p \times 1$ represents the sampled values of $g(t - \tau)$ with sampling rate $\frac{1}{T}$. The matrix $S_{mk}$ denotes the spreading-code matrix for the $m$th transmit antenna of the $k$th user. Due to the convolution operation, $S_{mk}$ is a Toeplitz matrix with

$$e_{mk}^s = \begin{bmatrix} s_{mk}(0), & 0^T_{p-1}, & s_{mk}(N-1), & 0^T_{p-1}, & \ldots, & s_{mk}(N-L+2), & 0^T_{p-1} \end{bmatrix}^T$$

as its first column and

$$r_{mk}^s = \begin{bmatrix} s_{mk}(0), & 0^T_{p-1}, & \ldots, & s_{mk}(N-1), & 0^T_{p-1}, & s_{mk}(0), & \ldots, & s_{mk}(L-1) \end{bmatrix}$$

(5)

as its first row, where $0^T_{p-1}$ is a $(\rho - 1) \times 1$ vector consisting of $\rho - 1$ zero elements. To include the inter symbol interference (ISI) between adjacent symbols, the corresponding data matrix, $D_{mk}(i)$, is also a Toeplitz matrix with

$$e_{mk}^d = \begin{bmatrix} d_{mk}(i), & 0^T_{N_p-1}, & d_{mk}(i-1), & 0^T_{N_p-1}, \ldots, & d_{mk}(i+Q-1), & 0^T_{N_p-1} \end{bmatrix}^T$$

(6)

as its first column and

$$r_{mk}^d = \begin{bmatrix} d_{mk}(i), & 0^T_{N_p-1}, & \ldots, & d_{mk}(i+Q-1), & 0^T_{N_p-1} \end{bmatrix}$$

(7)

as its first row.

Due to multipath propagation, signals experience fading effects, delay spreads and angle spreads when they go through a wireless channel. Suppose that, at the $h$th user, $P$ incoming rays arrive at the receiver. After coherent frequency down-conversion, the complex baseband equivalent representation of the receive signals can be expressed as

$$x^D(t) = \sum_{w=1}^{W} \sum_{p=1}^{P} a_{Rw} (\phi^D_{R,p}(w)p^D_{p}(t))$$

$$= \sum_{m=1}^{M} \sum_{n=1}^{N} a_{T,m} (\phi^D_{T,p}(t)) z_m^Q(t - t^D_{m,n}(t)) + n(t),$$

(8)

where $a_{Rw} (\phi^D_{R,p}(w))$ denotes the receive array response for a path arriving from the DOA, $\phi^D_{R,p}(w)$. Assume that a uniform linear array (ULA) is employed both at the receiver and the transmitter, $a_{Rw} (\phi^D_{R,p}(w)) = \frac{1}{\sqrt{M_r}} \begin{bmatrix} e^{j\pi \sin(w_1)}, & \ldots, & e^{j\pi (M_r - 1) \sin(w_1)} \end{bmatrix}^T$. The first column and
where \( G_{m,k}(\tau) = \left[ S_{m,k}^T \cdot g(t_1), \ldots, S_{m,k}^T \cdot g(t_P) \right] \) with dimension \((N - \kappa) P \times P\), and \( \mathbf{N}(i) \) is the truncated version of \( \mathbf{N}(i) \).

### 3. The Proposed Approach

In obtaining a set of diversity channels, conventional spatial diversity techniques generally assume that the signal fading at the receive antenna elements are uncorrelated and the optimum diversity can be achieved by the elementwise maximal ratio combining. However, in real propagation environments, especially when the receive antenna elements are spaced insufficiently far apart, fading effects among them are no longer uncorrelated [14]. Under this circumstance, the diversity combining should be investigated among the multipaths with independent fadings rather than among the receive antenna elements. Therefore, in this section, we design the diversity combiners based on the signals received from each individual path instead of combining the signals received from the antenna elements.

Shown in Fig. 2 is the structure of the proposed algorithms which consist of 1) a channel DOA-delay estimator by using the ST-MUSIC algorithm, 2) a space-time channel decoupler to decompose the multipath multiuser channel into a set of independent single user parallel AWGN subchannels, 3) a fading amplitude estimator, and 4) a pathwise maximal ratio combiner for signal detection. Two space-time signal detection algorithms are developed with different MAI suppression criteria used by the space-time channel decoupler. One is referred to as the ST-complementary orthogonal projection (ST-COP) algorithm, and the other is referred to as the ST-minimum variance distortionless response (ST-MVDR) algorithm. In the following, we start to introduce the ST-MUSIC algorithm.

### 3.1 The ST-MUSIC

It is shown from (10) that the DOA and the delay information are respectively in the column and row spaces of \( \mathbf{X}(i) \). With (10), several parameter estimation algorithms, such as the MUSIC [7], and the ESPRIT [13] algorithms can be utilized for DOA estimation. As for blind delay estimation, the ESPRIT-based algorithm cannot work without the knowledge of the spreading sequences of all active users. In a downlink DS-CDMA scenario, lack of information on the

![Fig. 2](image-url) The structure of the proposed algorithm.
total number of active users and their corresponding spreading sequences renders only the MUSIC algorithm applicable because 1) the MUSIC algorithm does not require the spreading sequences of the other users, and 2) it is also insensitive to signal subspace inflation. With this observation, we herein pick the MUSIC-based algorithm for parameter estimation in the proposed algorithm.

The ST-MUSIC algorithm is an extension of the JADE-MUSIC algorithm [10], which is designed for the DOA-delay joint estimation for TDMA systems with training sequences inserted in each data burst to work as a virtual array. However, in a downlink DS-CDMA system, the spreading sequence of the user of interest can serve the same function.

By stacking all columns of \( X(i) \) in (9), and making use of the rules \( \text{vec}[\text{XYZ}] = (Z^T \otimes X) \cdot \text{vec}[Y] \) and \( \text{vec}[X_1 + X_2] = \text{vec}[X_1] + \text{vec}[X_2] \), where \( \text{vec} (\cdot) \) denotes the columnwised stacking operation applied on the embraced matrix and \( \otimes \) denotes the kronecker product, we obtain the receive space-time vector as

\[
\mathbf{x}(i) = \text{vec}[X(i)] = \sum_{m,k} d_{m,k}(i) \boldsymbol{Q}_{m,k}(\theta_k, \tau) \beta + \mathbf{n},
\]

(11)

where

\[
\boldsymbol{Q}_{m,k}(\theta_k, \tau) = \begin{bmatrix} \boldsymbol{q}_{m,k}(\theta_{k,1}, \tau_1) & \cdots & \boldsymbol{q}_{m,k}(\theta_{k,p}, \tau_p) \end{bmatrix},
\]

(12)

and \( \boldsymbol{q}_{m,k}(\theta_{k,p}, \tau_p) = \mathbf{S}^T_{m,k} \mathbf{g} (\tau_p) \otimes \mathbf{a} (\theta_k, \rho) \) is referred to as the space-time signature of the \( p \)th path contributed from the \( m \)th transmit antenna for the \( k \)th user. \( \mathbf{n} = \text{vec}(\mathbf{N}) \) denotes the stacked noise vector.

To exploit the signals inside the column space of \( \mathbf{x}(i) \) in (11), the sample average space-time covariance matrix is

\[
\mathbf{R}^{\text{ST}} = E \{ \mathbf{x}(i) \mathbf{x}^H(i) \} = \sum_{m,k} \mathbf{Q}_{m,k}(\theta_k, \tau) \mathbf{R}_{\text{BB}} \mathbf{Q}^H_{m,k}(\theta_k, \tau) + \sigma_n^2 \mathbf{I},
\]

(13)

where \((\cdot)^H\) denotes the complex conjugate transpose operation, and \( \mathbf{R}_{\text{BB}} = E \{ \mathbf{g} \mathbf{g}^H \} = \text{diag} \{ \sigma_1^2, \cdots, \sigma_p^2 \} \) denotes the covariance matrix of the multipath fading amplitudes with \( \sigma_n^2 \) representing the variance of the multipath fading amplitude \( \beta_p \). We assume that the \( P \) paths fade independently. The eigen-decomposition of \( \mathbf{R} \) can be expressed as

\[
\mathbf{R}^{\text{ST}} = \mathbf{V}_s \Lambda_s \mathbf{V}_s^H + \sigma_n^2 \mathbf{V}_n \mathbf{V}_n^H
\]

(14)

where \( \mathbf{V}_s \) denotes the space-time signal subspace matrix, consisting of the \( M_T K_P \) dominant eigenvectors of \( \mathbf{R}^{\text{ST}} \), \( \Lambda_s \) is a diagonal matrix with the eigenvalues corresponding to \( \mathbf{V}_s \) as its diagonal elements, and \( \mathbf{V}_n \) is the complementary orthogonal matrix of \( \mathbf{V}_s \). The main idea behind the ST-MUSIC algorithm is that \( \boldsymbol{Q}_{m,k}(\theta, \tau) \) shares the same column space with \( \mathbf{V}_s \). The joint DOA and delay estimation problem can thus be described as

\[
\left( \hat{\theta}, \hat{\tau} \right) = \arg \min_{(\theta, \tau)} \left\{ \sum_{i=1}^{M_T} \mathbf{q}_{m,1}^H(\theta, \tau) \mathbf{P}^{\text{V}_s}_{m,1}(\theta, \tau) \right\},
\]

(15)

where \( \mathbf{P}^{\text{V}_s}_{m,1} = \mathbf{I} - \mathbf{V}_s \mathbf{V}_s^H \) denotes the projection matrix of the orthogonal complementary subspace of \( \mathbf{V}_s \). Note that summation in (15) is taken only over transmit antenna \( m \), not over the users, since only the spreading sequences of the first user, the user of interest, is known. The number of the eigenvectors required for forming the appropriate signal subspace matrix can be obtained by using either the AIC or MDL methods [15].

It is noteworthy that, with dimension \( M_R (N - \kappa) \rho \times 1 \), the space-time array vector \( \mathbf{q}_{m,1}(\theta, \tau) \) makes the ST-MUSIC algorithm being capable of resolving \( M_R (N - \kappa) \rho - 1 \) multipaths. For the rich scattered environment with large number of multipaths, the resolvability of the ST-MUSIC can be easily extended without any hardware cost by 1) increasing the over-sampling rate \( \rho \), and 2) employing a long spreading sequence to increase \( N \).

In addition, an interesting feature of the ST-MUSIC algorithm is that the dimension of the signal subspace can be inflated without affecting the algorithm performance significantly. It hence makes the ST-MUSIC algorithm insensitive to the natural spreading of the signal subspace, particularly useful for the clustered multipath scenarios due to DOA angular spread and delay temporal spread.

### 3.2 The Space-Time Channel Decoupler

The space-time channel decoupler decomposes the channel according to the space-time signature of each multipath. Based on the space-time signatures of the user of interest, the space-time channel decoupler of the ST-COP algorithm isolates each path by using the COP matrix of an appropriate self-exclusive matrix. To retain the \( p \)th path of the channel and null out the others, we define the space-time self-exclusive matrix as

\[
\mathbf{Q}^*_{p,m} = \left[ \left[ \mathbf{q}_{m,1}(\hat{\theta}_p', \hat{\tau}_p') \right]_{(p',m') \neq (p,m)} \right],
\]

(16)

The corresponding COP matrix is then expressed as

\[
\mathbf{U}_{p,m} = \mathbf{I} - \mathbf{Q}^*_{p,m} \left( \mathbf{Q}^*_{p,m} \right)^H,
\]

(17)

where \((\cdot)^H\) denotes the pseudo inverse of the embraced matrix. Furthermore, in addition to isolating each subchannel, the space-time channel decoupler serves to help suppressing the MAIs and scalarizing the vector signal in (11). To do this, in conjunction with the space-time signature of the user of interest, the weight vectors of the ST-COP space-time channel decoupler are given as

\[
\mathbf{w}^{\text{COP}}_{p,m} = \mathbf{U}_{p,m} \mathbf{q}_{m,1}(\hat{\theta}_p', \hat{\tau}_p'),
\]

(18)

and thus the outputs of these decouplers are
\[ u_{p,m}^{\text{COP}} (i) = \left( w_{p,m}^{\text{COP}} \right)^H \hat{x} (i) \approx d_{m,1} (i) \beta_p + n_{p,m}. \] (19)

Alternatively, the space-time channel decoupler can also be implemented based on the MVDR filters. The MVDR filters have been deployed in many signal processing applications, such as spatial beamformers and frequency filters, which seek to find an optimum weight vector so as to minimize both the interference and the noise at the filter outputs. To design an MVDR space-time channel decoupler for our purpose, we define the corresponding constraint matrix and response vector, respectively, as

\[ C_{p,m} = \left[ q_{m,1} \theta_{p,\bar{r}} \hat{\tau}_p + q_{m,1} \theta_{p,\bar{r}} \hat{\tau}_p \right]_{p \times m}, \]

\[ f = \begin{bmatrix} 1 & 0 & \cdots & 0 \end{bmatrix}^T. \] (21)

The MVDR space-time channel decoupler \( w_{p,m} \) can then be formulated as

\[ w_{p,m} = \arg \min_w \left\{ \left| u_{p,m}^{\text{MVDR}} (i) \right|^2 \right\}, \text{ s.t. } C_{p,m}^H w = f, \] (22)

where \( u_{p,m}^{\text{MVDR}} (i) = w_{p,m}^H \hat{x} (i) \) is the output of the space-time channel decoupler \( w_{p,m} \). We may readily solve (22) by using the constraint optimization method and obtain

\[ w_{p,m} = \left( R_{\text{ST}} \right)^{-1} C_{p,m} \left[ C_{p,m}^H \left( R_{\text{ST}} \right)^{-1} C_{p,m} \right]^{-1} f. \] (23)

The output of the MVDR filters can thus be expressed as

\[ u_{p,m}^{\text{MVDR}} (i) \approx \beta_p d_{m,1} (i) + \bar{n}_p, \] (24)

where \( \bar{n}_p = w_{p,m}^H n (i) \) is the output noise. Note that in (19) and (24), the output signal of the space-time channel decoupler is equivalent to a set of independent parallel sub-channels with additive noises.

### 3.3 Fading Amplitude Estimation and the PMRC

The method of fading amplitude estimation and signal detection employed is identical in both the ST-COP algorithm and the ST-MVDR algorithm. and, therefore, only the ST-COP algorithm is used as demonstrating this method. In addition, we ignore the superscript in (19) for notational simplicity. To estimate the fading amplitude, all the signals in (19) are collected to yield

\[ u_m (i) = \begin{bmatrix} u_{1,m} (i) \\ \vdots \\ u_{p,m} (i) \end{bmatrix} \approx d_{m,1} (i) \beta + n. \] (25)

Consequently, the corresponding signal combining can be executed on a pathwise basis as

\[ \hat{u}_m (i) = \left( \hat{\beta} \right)^H u_m (i) \approx d_{m,1} (i) \left| \beta \right|^2 \alpha + \bar{n}, \] (26)

where \( \bar{n} = \left( \hat{\beta} \right)^H n \) is the output noise. Differential decoding can then be used to extract the phase difference between neighboring \( d_{m,1} \)'s to eliminate the phase ambiguity caused by \( \alpha \).

The main difference between the ST-COP algorithm and the ST-MVDR algorithm is that the former is aiming at only cancelling the inter-path interference (IPI), while the latter is designed for the goal of maximizing the signal-to-interference-plus-noise-ratio (SINR) of the decoupler outputs. Therefore, the ST-MVDR algorithm generally outperforms the ST-COP algorithm due to its capability of MAI cancellation.

However, the DOAs and delays of the channel multipaths may not be perfectly estimated at the receiver. In that case, consistent with the spirit of minimizing filter output powers, the ST-MVDR may cancel the signal of interest together with the MAIs. This self-cancelling effect degrades the performance of the ST-MVDR algorithm, especially in the seriously noise-contaminated environment where the accurate spatial and temporal signatures can hardly be provided.

### 4. Simulations and Discussions

In this section, computer simulations are conducted to assess the proposed algorithms. Consider an \( M_T = 4 \) and \( M_R = 2 \) system with half wavelength spaced ULA at both the transmitter and the receiver. Differentially encoded BPSK (i.e., \( M = 2 \)) data sequences, each spreaded with a gold code sequence of length 31, \( N = 31 \), are transmitted over the channel.

We assume a Rayleigh-fading multipath channel with two groups of totally three paths \((P = 3)\) seen at the receiver for the user of interest. To illustrate that the ST-MUSIC algorithm can resolve the multipaths either with different DOAs or the ones with different delays. The DOAs and delays of the three paths are particularly selected as \( \theta_R = \left[ -31.5^\circ \ -30.1^\circ \ 19^\circ \right] \) and \( \tau = \left[ 0.74 T_c \ 2.82 T_c \ 0.72 T_c \right] \), respectively, where \( T_c = 0.8 \mu s \) is the chip period. Note, the first two rays above have close-by DOAs but have different delays, while the later two paths have different DOAs but have close-by delays. Suppose that the received signals are coherently downconverted, and then sampled with a rate of \( \rho = 2 \) samples/chip. The first five chips in each symbol are ignored to prevent the sampled data from suffering any inter symbol interference (ISI). Five hundred Monte Carlo trials are carried out as the number of active users are set at \( K = 1 \) and \( K = 2 \), respectively. The signal to noise ratio (SNR) defined as \( SNR = \frac{K}{\sigma_s} \).
ranges from $-9$ dB to $6$ dB with $3$-dB step increment, where $\sigma^2_n$ denotes the noise power. Two hundred symbols are processed in each trial. The average fading amplitudes of the multipaths are randomly generated with normalized $0$ dB amplitude and assumed constant for every $20$ symbol periods (i.e., $\lambda = 10$).

Figure 3 shows the pseudo spectrum of the ST-MUSIC algorithm for $\text{SNR} = -3$ dB and $K = 1$. As shown in the figure, the ST-MUSIC algorithm successfully identifies the DOA-delays of the received signals. Figures 4 and 5 illustrate the comparison of the root mean square error (RMSE) of the estimated DOAs and delays with the corresponding Cramer Rao lower bound (CRLB) [11], respectively.

By passing through the estimated DOA-delays to the space-time decouplers for further path isolation and MAI suppression, Figs. 6 and 7 compare the bit error rates (BERs) of the proposed algorithms with the space-time RAKE receivers [16] and the space-time minimum mean square (ST-MMSE) algorithm [17] for $K = 1$ and $K = 2$, respectively. Specifically, with the channel information known to the receiver, the ST-MMSE algorithm [17] is commonly believed the optimal algorithm for blind signal detection and is employed as a target of performance limit pursued by the proposed algorithm herein. When in the single user $K = 1$ case, since there is no MAIs, Fig. 6 shows that all algorithms perform almost the same in the BER performance. In the presence of MAIs, that is, when $K = 2$, Fig. 7 shows that the ST-MVDR algorithm, due to its excellent capability in MAI suppression, has roughly the same BER performance as that of the ST-MMSE and significantly outperforms the others.

Since the algorithms proposed in this paper rely on the estimates of channel parameters, it is useful to analyze the sensitivity of the blind signal detection algorithms to the channel parameter estimation error. Figure 8 illustrates BERs of various ST-based algorithms versus channel parameter estimation error, in the case $\left(\frac{\theta}{M_R}, T_c\right)$ is defined as a space-time beamwidth of the systems with $M_R$-element ULA at the receiver, the DOA-delay estimation error, i.e. the degree of deviation from the true value, of each path is set at fixed values within the range from $0$ % to $50$% of the space-time beamwidth, and $\text{SNR} = 0$ dB. As indicated in the figure, the ST-MVDR algorithm appears more sensitive.

**Fig. 3** The pseudo spectrum of the ST-MUSIC algorithm when $M_T = 4$, and $K = 2$.

**Fig. 4** The RMSE of the estimated DOAs when $M_T = 4$, and $K = 2$.

**Fig. 5** The RMSE of the estimated delays when $M_T = 4$, and $K = 2$.

**Fig. 6** The BERs of the proposed algorithms when $M_T = 4$, and $K = 1$. 
Fig. 7 The BERs of the proposed algorithms when $M_T = 4$, and $K = 2$.

Fig. 8 The BERs of the proposed algorithms with mismatched DOA-delay estimates.

Figures and vulnerable as the mismatch increases.

5. Conclusions

This paper proposes a blind algorithm for blind joint channel parameter estimation and signal detection for wireless DS-CDMA down-link MIMO systems. First, the ST-MUSIC algorithm is developed for estimating the DOA-delays of the received signals. Then, based on the estimated DOA-delays, signals from multipath channels are isolated in the space-time domain with a space-time channel decoupler. Finally, the succeeding maximal ratio combiner is applied to extract the data on a pathwise basis. The ST-MVDR algorithm generally has better BERs than the ST-COP algorithm, since the former is capable of suppressing more MAIs. However, the ST-MVDR algorithm, suffering from the self-cancelling effect, may seriously degrade the system performance when DOA-delays can not be precisely estimated.

References

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