DOWNLINK CHANNELS IDENTIFICATION FOR SPACE-TIME CODED MULTIPLE-INPUT MULTIPLE-OUTPUT MC-CDMA SYSTEMS

Xiaojun Wu [1], *Qinye Yin* [2], *Yanxing Zeng* [2]

[1] Tsinghua University, Beijing 100084, China
[2] Xi' an Jiaotong University, Xi' an 710049, China

ABSTRACT

This paper integrates space-time coding (STC) techniques and the multiple-input multiple-output (MIMO) antenna architecture with multicarrier code-division multiple assess (MC-CDMA) systems. For brevity, we can call such integration as space-time coded MIMO MC-CDMA systems. A numerical model is presented in details for the aforesaid space-time coded MIMO MC-CDMA system Successively, by exploiting subspace decomposition techniques, one novel blind identification method is derived for downlink wireless channels in the space-time coded MIMO MC-CDMA system. Owing to no need of pilot symbols to identification wireless channels between multiple-transmit antennas and multiple-receive ones, our proposed algorithm can finely keep the spectrum efficiency of the space-time coded MIMO MC-CDMA system. Computer simulation results demonstrate the validity and the performance of our proposed algorithm.

1. INTRODUCTION

Multicarrier code-division multiple access (MC-CDMA) systems are widely considered as one promising technique for next-generation (NextG) wireless communications[1]. With cooperation of space-time coding (STC) techniques and multiple-input multiple-output (MIMO)[2], MC-CDMA systems can better meet the demand of NextG wireless communications, such as higher data-rate, larger capacity, and higher fidelity in wireless channels.

Therefore, this paper focuses on aforementioned space-time coded MIMO MC-CDMA systems. We present the numerical model of the space-time coded MIMO MC-CDMA system in details. Successively, we derive one novel blind identification scheme for downlink channels in the space-time coded MIMO MC-CDMA system. With the assistance of computer simulations, we illustrate the different effects of several typical STC mechanisms on the mean-squared error (MSE) performance of our proposed algorithm.

2. MODEL OF SPACE-TIME CODED MIMO MC-CDMA

2.1. Baseband Model

It is assumed that *K* users are randomly distributed around a cell site. All *K* active users share the same set of subcarriers. The number of subcarriers equals to the length of spreading code *G*. Fig. 1 displays the baseband model of the space-time coded MIMO MC-CDMA systems.

As shown in Fig. 1, when MC-CDMA systems over wireless finite impulse response (FIR) channels, a usual approach for combating the resultant inter-block interference (IBI) is via inserting cyclic prefix (ICP) to each transmitted data block. Meanwhile, by removing CP (RCP) at the beginning of each received data block, the IBI can be eliminated.

The MC-CDMA scheme does the spreading spectrum (SS) operation in the frequency domain[1]. We define the *m*-th assigned frequency-domain spreading code for the *k*-th user as a vector $\tilde{\mathbf{c}}_k^{(m)}$ ($k = 1,\cdots,K$; $m = 1,\cdots,M_T$), which can be written as

$$\tilde{\mathbf{c}}_k^{(m)} = \left[\, \tilde{c}_k^{(m)}(1) \quad \tilde{c}_k^{(m)}(2) \quad \cdots \quad \tilde{c}_k^{(m)}(G) \,\right]^{\mathrm{T}} \tag{1}$$

Given the length *G* of frequency-domain spreading codes, the maximum number of active users is therefore determined to be

$$K_{\max} = \lfloor G/M_T \rfloor \tag{2}$$

2.2. Numerical Model

For clarity, we describe the downlink wireless finite impulse response (FIR) channel between the *p*-th ($p = 1,\cdots,M_R$) receive antenna and the *m*-th ($m = 1,\cdots,M_T$) transmit antenna as

$$\mathbf{h}^{(pm)} = \left[h^{(pm)}(0) \quad h^{(pm)}(1) \quad \dots \quad h^{(pm)}(L_{\mathrm{ch}}) \right]^{\mathrm{T}} \tag{3}$$

where L_{ch} expresses the maximum common FIR-channel length of all downlink channels between receive antennas and transmit antennas. Without loss of generality, we assume $L_{\mathrm{ch}} < G$.

The effects of above dispersive channels appear as random frequency-domain attenuations in each subcarrier. We depict the frequency-domain attenuations on all subcarrier channels between the *p*-th receive antenna and the *m*-th transmit antenna as the vector $\boldsymbol{\varsigma}^{(pm)}$ with dimensions $G \times 1$, which can be obtained by performing the discrete Fourier transform (DFT) on the aforementioned time-domain FIR vector $\mathbf{h}^{(pm)}$, that is,

$$\begin{aligned} \boldsymbol{\varsigma}^{(pm)} &= \left[\eta^{(pm)}(1) \quad \eta^{(pm)}(2) \quad \cdots \quad \eta^{(pm)}(G) \right]^{\mathrm{T}} \\ &= \mathbf{F}_{\mathrm{DFT}}(:,1:(L_{\mathrm{ch}}+1))\,\mathbf{h}^{(pm)} = \mathbf{F}_{\mathrm{FRO}}\,\mathbf{h}^{(pm)} \end{aligned} \tag{4}$$

where the matrix $\mathbf{F}_{\mathrm{DFT}}$ with dimensions $G \times G$ is a DFT matrix, and its every entry is $(\mathbf{F}_{\mathrm{DFT}})_{l,l'} = \frac{1}{\sqrt{G}} \exp(-j\frac{2\pi}{G}(l-1)(l'-1))$ ($l, l' = 1,\cdots,G$); the matrix $\mathbf{F}_{\mathrm{FRO}}$ with dimensions $G \times (L_{\mathrm{ch}}+1)$ is named as the frequency response operator (FRO) and is defined as

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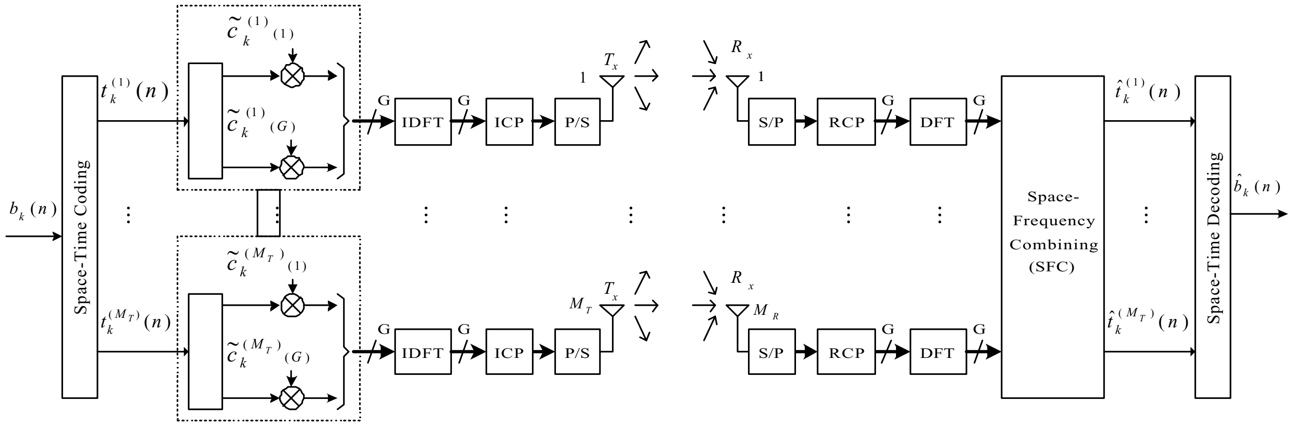

Fig. 1 The baseband model of the space-time coded MIMO MC-CDMA system.

$\mathbf{F}_{\text{FRO}} = \mathbf{F}_{\text{DFT}}(:,1:(L_{\text{ch}}+1))$, which means that the matrix $\mathbf{F}_{\text{FRO}}$ consists of the first $(L_{\text{ch}}+1)$ column vectors of the DFT matrix $\mathbf{F}_{\text{DFT}}$.

At one arbitrary wireless terminal, we define the frequency-domain IBI-free downlink-received data vector at the *p*-th receive antenna from the *m*-th transmit antenna as

$$\tilde{\mathbf{x}}_k^{(pm)} = diag(\tilde{\mathbf{c}}_k^{(m)})\,\text{ç}^{(pm)}\,t_k^{(m)} = \mathbf{Ö}_k^{(m)}\,\text{ç}^{(pm)}\,t_k^{(m)} \quad (5)$$

where $\mathbf{Ö}_k^{(m)}$ is a matrix with dimensions $G\times G$, and is defined as $\mathbf{Ö}_k^{(m)} = diag(\tilde{\mathbf{c}}_k^{(m)})$.

Successively, the frequency-domain IBI-free downlink-received data vector at the *p*-th receive antenna from all transmit antennas is given by

$$\tilde{\mathbf{x}}_k^{(p)} = \sum_{m=1}^{M_T}\tilde{\mathbf{x}}_k^{(pm)} = \sum_{m=1}^{M_T}\mathbf{Ö}_k^{(m)}\,\text{ç}^{(pm)}\,t_k^{(m)} = \mathbf{Ö}_k\,\mathbf{Ç}^{(p)}\mathbf{t}_k \quad (6)$$

where the matrix $\mathbf{Ö}_k$ with dimensions $G\times M_T G$ is defined as $\mathbf{Ö}_k = \left[\mathbf{Ö}_k^{(1)} \;\; \mathbf{Ö}_k^{(2)} \;\; \cdots \;\; \mathbf{Ö}_k^{(M_T)}\right]$; the vector $\mathbf{t}_k$ with dimensions $M_T\times 1$ is defined as $\mathbf{t}_k = \left[t_k^{(1)} \;\; t_k^{(2)} \;\; \cdots \;\; t_k^{(M_T)}\right]$, which describes the space-time coded symbol block of the *k*-th user and consists of M_T symbols from M_T transmit antennas; the matrix $\mathbf{Ç}^{(p)}$ with dimensions $M_T G\times M_T$ is defined as

$$\begin{aligned}\mathbf{Ç}^{(p)} &= diag(\text{ç}^{(p1)},\text{ç}^{(p2)},\cdots,\text{ç}^{(pM_T)}) \\ &= \left[\mathbf{â}^{(p1)} \;\; \mathbf{â}^{(p2)} \;\; \cdots \;\; \mathbf{â}^{(pM_T)}\right]\end{aligned} \quad (7)$$

According to (7), we write the column vector $\mathbf{â}^{(pm)}$ $(p=1,\cdots,M_R;\; m=1,\cdots,M_T)$ with dimensions $M_T G\times 1$ as

$$\begin{aligned}\mathbf{â}^{(pm)} &= \left[\mathbf{0}^{(1)\,\text{T}} \cdots \mathbf{0}^{(m-1)\,\text{T}}\;\text{ç}^{(pm)\,\text{T}}\;\mathbf{0}^{(m+1)\,\text{T}} \cdots \mathbf{0}^{(M_T)\,\text{T}}\right]^{\text{T}} \\ &= \left[\mathbf{O}^{(1)\,\text{T}} \cdots \mathbf{O}^{(m-1)\,\text{T}}\;\mathbf{F}_{\text{FRO}}^{\text{T}}\;\mathbf{O}^{(m+1)\,\text{T}} \cdots \mathbf{O}^{(M_T)\,\text{T}}\right]^{\text{T}}\mathbf{h}^{(pm)} \\ &= \mathbf{F}^{(m)}\,\mathbf{h}^{(pm)}\end{aligned} \quad (8)$$

where all $\mathbf{0}^{(i)}$ $(i=1,\cdots,M_T; i\neq m)$ with dimensions $G\times 1$ are vectors of zeros; all $\mathbf{O}^{(i)}$ $(i=1,\cdots,M_T; i\neq m)$ with dimensions $G\times(L_{\text{ch}}+1)$ are matrices of zeros; the matrix $\mathbf{F}^{(m)}$ $(m=1,\cdots,M_T)$ is with dimensions $M_T G\times(L_{\text{ch}}+1)$.

Substituting (8) into (7), (7) is then rewritten as

$$\mathbf{Ç}^{(p)} = \left[\mathbf{F}^{(1)}\mathbf{h}^{(p1)} \;\; \mathbf{F}^{(2)}\mathbf{h}^{(p2)} \;\; \cdots \;\; \mathbf{F}^{(M_T)}\mathbf{h}^{(pM_T)}\right] \quad (9)$$

Stacking these frequency-domain IBI-free downlink-received data vectors corresponding to all M_R receive antennas, we obtain an extended IBI-free downlink-received data vector with dimensions $M_R G\times 1$ as

$$\tilde{\mathbf{x}}_k = \left[\tilde{\mathbf{x}}_k^{(1)\,\text{T}} \;\; \tilde{\mathbf{x}}_k^{(2)\,\text{T}} \;\; \cdots \;\; \tilde{\mathbf{x}}_k^{(M_R)\,\text{T}}\right]^{\text{T}} = \tilde{\mathbf{C}}_k\,\mathbf{Ç}\,\mathbf{t}_k \quad (10)$$

where the matrix $\tilde{\mathbf{C}}_k$ with dimensions $M_R G\times M_R M_T G$ and the matrix $\mathbf{Ç}$ with dimensions $M_R M_T G\times M_T$ are, respectively,

$$\tilde{\mathbf{C}}_k = \mathbf{I}_{M_R}\otimes\mathbf{Ö}_k \quad (11)$$

$$\begin{aligned}\mathbf{Ç} &= \left[\mathbf{Ç}^{(1)\,\text{T}} \;\; \mathbf{Ç}^{(2)\,\text{T}} \;\; \cdots \;\; \mathbf{Ç}^{(M_R)\,\text{T}}\right]^{\text{T}} \\ &= \begin{bmatrix}\mathbf{F}^{(1)}\mathbf{h}^{(11)} & \mathbf{F}^{(2)}\mathbf{h}^{(12)} & \cdots & \mathbf{F}^{(M_T)}\mathbf{h}^{(1M_T)} \\ \mathbf{F}^{(1)}\mathbf{h}^{(21)} & \mathbf{F}^{(2)}\mathbf{h}^{(22)} & \cdots & \mathbf{F}^{(M_T)}\mathbf{h}^{(2M_T)} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{F}^{(1)}\mathbf{h}^{(M_R1)} & \mathbf{F}^{(2)}\mathbf{h}^{(M_R2)} & \cdots & \mathbf{F}^{(M_T)}\mathbf{h}^{(M_RM_T)}\end{bmatrix}\end{aligned} \quad (12)$$

where $\mathbf{I}_{M_R}$ is an M_R-order identity matrix, operator $\otimes$ depicts the Kronecker product among matrices.

In multiuser MIMO MC-CDMA systems, all K active users within one cell share all subcarriers at the same time. In downlink channels, owing to the fact that space-time coded symbols from different users are synchronized, we give the numerical model for the multiuser MIMO MC-CDMA system by

$$\tilde{\mathbf{x}} = \sum_{k=1}^{K}\tilde{\mathbf{x}}_k = \sum_{k=1}^{K}\tilde{\mathbf{C}}_k\,\mathbf{Ç}\,\mathbf{t}_k \quad (13)$$

Considering the thermal noise, we rewrite (13) as

$$\tilde{\mathbf{y}} = \tilde{\mathbf{x}} + \tilde{\mathbf{n}} \quad (14)$$

where every entry of the thermal noise matrix $\tilde{\mathbf{n}}$ with dimensions $M_R G\times 1$ is the complex zero-mean Gaussian noise with variance σ_n^2.

3. DOWNLINK CHANNELS IDENTIFICATION FOR SPACE-TIME CODED MC-CDMA SYSTEMS

3.1. Subspace Based Channel Identification Scheme

Now, we consider the auto-correlation matrix of $\tilde{\mathbf{x}}$ in (13).

$$\mathbf{R}_{\mathbf{xx}} = E[\tilde{\mathbf{x}}\tilde{\mathbf{x}}^{\mathrm{H}}] = E[(\sum_{k=1}^{K}\tilde{\mathbf{C}}_k \mathbf{Ç}\,\mathbf{t}_k)(\sum_{k=1}^{K}\tilde{\mathbf{C}}_k \mathbf{Ç}\,\mathbf{t}_k)^{\mathbf{H}}]$$
$$= \sum_{k=1}^{K}\tilde{\mathbf{C}}_k \mathbf{Ç}\,\mathbf{Ç}^{\mathbf{H}}\,\tilde{\mathbf{C}}_k^{\mathrm{H}} \quad (15)$$

where $\mathbf{R}_{\mathbf{xx}}$ is of dimensions $M_R G \times M_R G$.

Likewise, the auto-correlation matrix of $\tilde{\mathbf{y}}$ in (14) is

$$\mathbf{R}_{\mathbf{yy}} = E[\tilde{\mathbf{y}}\,\tilde{\mathbf{y}}^{\mathrm{H}}] = \sum_{k=1}^{K}\tilde{\mathbf{C}}_k \mathbf{Ç}\,\mathbf{Ç}^{\mathbf{H}}\,\tilde{\mathbf{C}}_k^{\mathrm{H}} + \sigma_n^2\,\mathbf{I} \quad (16)$$

By doing eigenvalue decomposition on the noisy auto-correlation matrix $\mathbf{R}_{\mathbf{yy}}$, we can obtain

$$\mathbf{R}_{\mathbf{yy}} = \sum_{l=1}^{M_R G}\lambda_l \mathbf{u}_l \mathbf{u}_l^{\mathrm{H}} \quad (17)$$

where λ_l and $\mathbf{u}_l$ are, respectively, eigenvalues and corresponding eigenvectors of $\mathbf{R}_{\mathbf{yy}}$.

We sort the eigenvalues as $\lambda_1 \geq \lambda_2 \geq \cdots \geq \lambda_{M_R G}$. Then, as can be seen, eigenvectors associated with the $M_T K$ nonzero largest eigenvalues span the signal subspace where signals of different users lie, while remaining smallest $(M_R G - M_T K)$ eigenvalues span the noise subspace that is orthogonal to the signal subspace [3].

We define the noise subspace as the matrix $\mathbf{U}_N$ with dimensions $M_R G \times (M_R G - M_T K)$, that is

$$\mathbf{U}_N = \begin{bmatrix}\mathbf{u}_{M_T K+1} & \mathbf{u}_{M_T K+2} & \cdots & \mathbf{u}_{M_R G}\end{bmatrix} \quad (18)$$

By exploiting the orthogonality property between the noise subspace and the signal subspace, the following matrix equations hold.

$$(\tilde{\mathbf{C}}_k \mathbf{Ç})^{\mathrm{H}}\mathbf{U}_N = \mathbf{Ç}^{\mathbf{H}}\,\tilde{\mathbf{C}}_k^{\mathrm{H}}\,\mathbf{U}_N = \mathbf{O} \quad (19)$$

where $\mathbf{O}$ with dimensions $M_T \times (M_R G - M_T K)$ is matrix of zeros.

In above matrix equations, there are $M_R\ M_T\ (L_{\mathrm{ch}}+1)$ unknowns while there are $M_T\ (M_R G - M_T K)$ linear equations. When $M_T\ (M_R G - M_T K) \geq M_R\ M_T\ (L_{\mathrm{ch}}+1)$, the linear equation set defined in (19) becomes overdetermined. Therefore, by solving (19), we can estimate the corresponding downlink channel matrix $\hat{\mathbf{Ç}}$ up to an ambiguous constant.

When $M_T\ (M_R G - M_T K) < M_R\ M_T\ (L_{\mathrm{ch}}+1)$, the linear equation set in either (19) becomes underdetermined. We, therefore, are unable to estimate the downlink channel matrix $\hat{\mathbf{Ç}}$ uniquely.

Under the constraint of the maximum number of active users defined in (2), the minimum number of linear equations is given by

$$\varepsilon_{\min} = M_T\ (M_R G - M_T K_{\max})$$
$$= M_T\ (M_R G - M_T \frac{G}{M_T}) = M_T\ (M_R - 1)G \quad (20)$$

If M_R in (20) is set as $M_R = 1$, the value of $\varepsilon_{\min}$ becomes zero resulting in the matrix equation defined in (19) is meaningless. Therefore, the number M_R of receive antennas must be constantly not less than 2 so that the matrix equation defined in (19) is meaningful when the number of active users reaches its upper limit.

Moreover, to avoid the noise subspace $\mathbf{U}_N$ to collapse to null, there is one essential constraint on the upper limit of the number of active users; that is, the number of active users satisfies the following inequation $(M_R G - M_T K) \geq 1$. Considering the principal constraint of the maximum number of active users defined in (2), the number of active users must satisfy $K \leq \min(\lfloor (M_R G - 1)/M_T \rfloor, \lfloor G/M_T \rfloor)$.

3.2. Solution of Overdetermined Linear Equation Set

In order to solve the overdetermined linear equation set defined in (19), we also can consider the following serials of equations, i.e.,

$$(\mathbf{Ö}_k^{(m)}\ \mathbf{F}_{\mathrm{FRO}}\,\mathbf{h}^{(pm)})^{\mathrm{H}}\,\mathbf{u}_l\,((p-1)G+1:pG)$$
$$= \mathbf{h}^{(pm)\ \mathrm{H}}\ \mathbf{F}_{\mathrm{FRO}}^{\mathrm{H}}\,\mathbf{Ö}_k^{(m)\ \mathrm{H}}\ \mathbf{u}_l\,((p-1)G+1:pG) = 0 \quad (21)$$

for $l = M_T K+1,\cdots,M_R G$; $p = 1,\cdots,M_R$; $m = 1,\cdots,M_T$.

Equations described in (19) are equivalent to the following equations.

$$\left\| \mathbf{h}^{(pm)\ \mathrm{H}}\ \mathbf{F}_{\mathrm{FRO}}^{\mathrm{H}}\,\mathbf{Ö}_k^{(m)\ \mathrm{H}}\ \mathbf{u}_l\,((p-1)G+1:pG) \right\|^2$$
$$= \mathbf{h}^{(pm)\ \mathrm{H}}\ \mathbf{F}_{\mathrm{FRO}}^{\mathrm{H}}\,\mathbf{Ö}_k^{(m)\ \mathrm{H}}\ \mathbf{u}_l\,((p-1)G+1:pG) \quad (22)$$
$$\cdot\,\mathbf{u}_l^{\mathrm{H}}((p-1)G+1:pG)\,\mathbf{Ö}_k^{(m)}\ \mathbf{F}_{\mathrm{FRO}}\ \mathbf{h}^{(pm)}$$
$$= \mathbf{h}^{(pm)\ \mathrm{H}}\ \mathbf{F}_{\mathrm{FRO}}^{\mathrm{H}}\,\mathbf{Ö}_k^{(m)\ \mathrm{H}}\ \mathbf{u}_l^{(p)}\ \mathbf{u}_l^{(p)\ \mathrm{H}}\,\mathbf{Ö}_k^{(m)}\ \mathbf{F}_{\mathrm{FRO}}\ \mathbf{h}^{(pm)} = 0$$

for $l = M_T K+1,\cdots,M_R G$; $p = 1,\cdots,M_R$; $m = 1,\cdots,M_T$. In (22), the vector $\mathbf{u}_l^{(p)}$ with dimensions $G \times 1$ is defined as $\mathbf{u}_l^{(p)} = \mathbf{u}_l\,((p-1)G+1:pG)$ $(p = 1,\cdots,M_R)$.

We can obtain the solution to above equations by solving the following subspace fitting problem..

$$\hat{\mathbf{h}}^{(pm)} = \arg\min_{\|\mathbf{h}^{(pm)}\|^2=1}\left[\sum_{k=1}^{K}\sum_{l=M_T K+1}^{M_R G}\mathbf{h}^{(pm)\ \mathrm{H}}\ \mathbf{F}_{\mathrm{FRO}}^{\mathrm{H}}\mathbf{Ö}_k^{(m)\ \mathrm{H}}\ \mathbf{u}_l^{(p)}\right.$$
$$\left.\cdot\,\mathbf{u}_l^{(p)\ \mathrm{H}}\ \mathbf{Ö}_k^{(m)}\ \mathbf{F}_{\mathrm{FRO}}\,\mathbf{h}^{(pm)}\right] \quad (23)$$
$$= \arg\min_{\|\mathbf{h}^{(pm)}\|^2=1}\left[\mathbf{h}^{(pm)\ \mathrm{H}}\left(\sum_{K=1}^{K}\mathbf{F}_{\mathrm{FRO}}^{\mathrm{H}}\mathbf{Ö}_k^{(m)\ \mathrm{H}}\ \mathbf{U}_N^{(p)}\right.\right.$$
$$\left.\left.\cdot\,\mathbf{U}_N^{(p)\ \mathrm{H}}\ \mathbf{Ö}_k^{(m)}\ \mathbf{F}_{\mathrm{FRO}}\right)\mathbf{h}^{(pm)}\right]$$
$$= \arg\min_{\|\mathbf{h}^{(pm)}\|^2=1}\left[\mathbf{h}^{(pm)\ \mathrm{H}}\ \mathbf{Q}^{(pm)}\ \mathbf{h}^{(pm)}\right]$$

where the matrix $\mathbf{Q}^{(pm)}$ with dimensions $(L_{\mathrm{ch}}+1)\times(L_{\mathrm{ch}}+1)$ is

$$\mathbf{Q}^{(pm)} = \sum_{K=1}^{K}\mathbf{F}_{\mathrm{FRO}}^{\mathrm{H}}\mathbf{Ö}_k^{(m)\ \mathrm{H}}\ \mathbf{U}_N^{(p)}\ \mathbf{U}_N^{(p)\ \mathrm{H}}\ \mathbf{Ö}_k^{(m)}\ \mathbf{F}_{\mathrm{FRO}} \quad (24)$$

Obviously, we can obtain the solution of (23) by the least-squares method. In short words, the estimated channel vector $\hat{\mathbf{h}}^{(pm)}$ between the *p*-th receive antenna and the *m*-th transmit antenna is just the eigenvector of $\mathbf{Q}^{(pm)}$ associated with its smallest eigenvalue.

By constructing $\mathbf{Q}^{(pm)}$ according to (24) for different receive antenna and transmit antenna ($p = 1,\cdots,M_R$; $m = 1,\cdots,M_T$), we can easily estimate the downlink channel in turn.

4. SIMULATION RESULTS

We evaluate the performance of the proposed algorithm through the MSE of channel identification, which is define as

$$MSE = \frac{1}{N_t}\sum_{i=1}^{N_t}\left\|\hat{\mathbf{H}}(i) - \mathbf{H}\right\|_F^2 \Big/ \left\|\mathbf{H}\right\|_F^2 \qquad (25)$$

where N_t is the number of Monte-Carlo trials in the simulation; the channel matrix $\mathbf{H}$ with dimensions $(L_{ch}+1)\times K$ is defined as $\mathbf{H} = [\mathbf{h}_1 \ \cdots \ \mathbf{h}_K]$; the matrix $\hat{\mathbf{H}}(i)$ is the estimate of $\mathbf{H}$ from the *i*-th trial; the operator $\|\bullet\|_F$ depicts the Frobenius norm. Note that, there is an ambiguous complex coefficient between the estimated channel $\hat{\mathbf{h}}_k$ and the original channel $\mathbf{h}_k$. Such ambiguity can be compensated by the method proposed in [4].

The auto-correlation matrix of the noisy extended IBI-free downlink-received data vector $\tilde{\mathbf{y}}$ in (14) can be estimated from the sampled average matrix as $\hat{\mathbf{R}}_{yy} = \frac{1}{L_b}\sum_{n=1}^{L_b}\tilde{\mathbf{y}}\,\tilde{\mathbf{y}}^H$, where L_b depicts the number of symbol blocks.

The four-phase shift keying (4PSK) modulation mode is used in simulations, and 200 Monte-Carlo trials are performed for each simulation. The spreading factor G is fixed to 32. Hadamard multiple access codes are assigned to different users. The channel impulse responses are generated from independent complex Gaussian random variables. The simulation environment with 18 active users is considered. The SNR is fixed to 12dB, and the length of channel is set to 7.

Four STC techniques are applied to the MIMO MC-CDMA system, such as the diagonal Bell Labs layered space-time architecture (D-BLAST)[5], the vertical BLAST (V-BLAST) architecture[6], the space-time block coding (STBC)[7], and the space-time trellis coding (STTC) for three or four transmit antennas[8].

Fig. 2 and Fig. 3 illustrate the MSE performance as a function of the number of symbol blocks L_b for the proposed algorithm. Because the redundant coding information for each STC mechanism is differently distributed across the space-time plane, the fidelity of the sampled average matrix $\hat{\mathbf{R}}_{yy}$ is varying, which results in the different MSE performance for individual STC mechanism.

It is interesting to compare Fig. 2 with Fig.3. With the increasing of the number of transmit antennas, more samples are necessary to provide a high-fidelity estimation of the auto-correlation matrix, which results in, for fixed L_b, performance slightly degrades.

5. CONCLUSIONS

In this paper, we have introduced a novel downlink channel identification scheme appropriate for the space-time coded MIMO MC-CDMA system over the frequency-selective fading channel. Obviously, without using pilot symbols, our proposed blind algorithm can finely keep the spectrum efficiency of the space-time coded MIMO MC-CDMA system. Our proposed algorithm is especially significant for the larger number of antennas.

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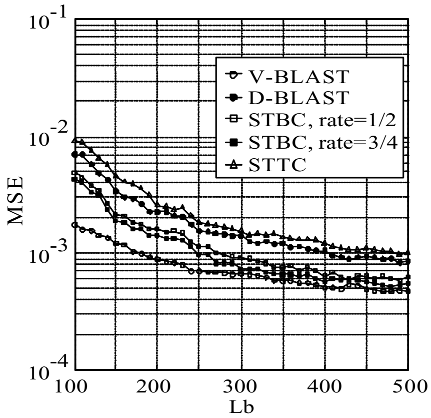

Fig. 2 MSE versus L_b (Tx=3).

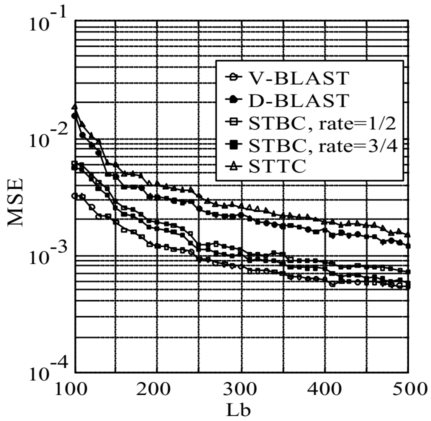

Fig. 3 MSE versus L_b (Tx=4).