Channel Estimation for Space-Time Block Coded OFDM Systems using Few Received Symbols

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Abstract: A novel subspace channel estimation is proposed for space-time block coded (STBC) multiple-input multiple–output (MIMO) orthogonal frequency division multiplexing (OFDM) systems. Considering the zero-padding technique, the signal model of the STBC-OFDM system is first studied. The major drawback of the subspace method is the slow convergence which requires a large number of received symbols to compute the noise subspace. The repetition index scheme is developed here to increase the number of equivalent received symbols. Using the space-time coding property, the forward-backward method is presented to improve the convergence speed further. Moreover, the repetition index scheme transforms the white noise into non-white one. The noise prewhitening technique is proposed to reduce the non-white noise effect. Computer simulations show the effectiveness of the proposed forward-backward method and noise prewhitening technique.

1 INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) techniques have been extensively developed in the modern wireless communication systems. The OFDM system utilizes the guard interval such as cyclic prefix (CP) or zero-padding (ZP) to avoid the inter-symbol interference (ISI) and converts the frequency-selective fading channel into a group of narrowband flat-fading channels with the help of discrete Fourier transform (DFT) property. The long term evolution (LTE)-advanced standard adopts the OFDM technique in the fourth-generation (4G) system to achieve a minimum peak data rate and spectral efficiency requirements (Dahlman, 2011). The OFDM with index modulation is further investigated for the next generation communication systems (Wen, 2016). Moreover, the multiple-input multiple-output (MIMO) system is proposed to increase the channel spectral and energy efficiency (Lu, 2014). The space-time coding technique offers both spatial diversity and coding gains (Jafarkhani, 2005). The space-time trellis coding (STTC) was introduced by Tarokh et al. At the same time, Alamouti presented the space-time block coding (STBC) scheme which uses two antennas at the transmitter and a simple maximum likelihood decoding algorithm at the receiver.

The channel estimation is necessarily performed before the receiver design in the coherent communication system. The pilot-based and blind-based channel estimations are the two most popular methods. The spectral efficiency of the pilot-based channel estimation is lower because of the insertion of periodic pilots. On the other hand, the blind subspace method, which requires a large number of received symbols to compute the noise and/or signal subspaces, is converged slowly but no extra bandwidth is needed.

In this paper we propose a fast convergence subspace technique to estimate the channel coefficients of the STBC ZP-OFDM systems. To improve the convergence speed of the blind method, the forward-backward averaging technique is presented in (Yu, 2009) for the MIMO ZP-OFDM system with Alamouti STBC. Using the circular property of a channel matrix, the cyclic repetition method (CRM) is investigated in (Zhang, 2014) for both the CP-OFDM and the ZP-OFDM systems with STBC coding. The repetition index scheme (RIS)

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has been used in the channel estimation of the single-input single–output ZP-OFDM system (Pan, 2013). We extend the RIS method to the STBC ZP-OFDM systems here. Two novel techniques including noise prewhitening and forward-backward techniques are developed to enhance the performance of RIS channel estimation. Simulation results will verify that the proposed techniques effectively improve the performance of the subspace channel estimation.

Notations are defined as follows. Vectors and matrices are denoted by boldface lower and upper case letters, respectively; superscripts of (.)*,(.)T and (.)H, denote the complex conjugate, transpose, and conjugate transpose, respectively; E{.} denotes the statistical expectation; $\mathbf{I}_n$ denotes an n×n identity matrix; ⊗ stands for Kronecker product; $\mathbf{0}_{m\times n}$ denotes a m×n matrix with all zero entries. $\|\cdot\|_F$ denotes the matrix or vector Frobenius norm. Let $\mathbf{V}=[\mathbf{V}^T(0)\cdots\mathbf{V}^T(M-1)]^T$ be a Ma×b matrix and each submatrix $\mathbf{V}$(n) be a a×b matrix, $\mathcal{T}_{M,S}(\mathbf{V})$ and $\mathcal{H}_{M,T}(\mathbf{V})$ denote a (M+S-1)a×Sb block Toeplitz matrix and a Ta × (M – T + 1)b block Hankel matrix, respectively

$$\mathcal{T}_{M,S}(\mathbf{V})=\begin{bmatrix}\mathbf{V}(0) & & \mathbf{0}\\ \vdots & \ddots & \\ \mathbf{V}(M-1) & & \mathbf{V}(0)\\ & \ddots & \vdots\\ \mathbf{0} & & \mathbf{V}(M-1)\end{bmatrix}$$

$$\mathcal{H}_{M,T}(\mathbf{V})=\begin{bmatrix}\mathbf{V}(0) & \mathbf{V}(1) & \cdots & \mathbf{V}(M-T)\\ \mathbf{V}(1) & \mathbf{V}(2) & \cdots & \vdots\\ \vdots & \vdots & \ddots & \vdots\\ \mathbf{V}(T-1) & \mathbf{V}(Q) & \cdots & \mathbf{V}(M-1)\end{bmatrix}$$

2 SYSTEM MODEL

Fig. 1 shows a K-user uplink STBC MIMO-OFDM system with J receive antennas. Let N be the number of subcarriers, $\mathbf{s}_i^{(k)}=\left[s_i^{(k)}(0)\cdots s_i^{(k)}(N-1)\right]^T$ be the OFDM symbol of the k-th user at time i, and $\overline{\mathbf{s}}_i^{(k)}$ and $\overline{\overline{\mathbf{s}}}_i^{(k)}$ be the coded symbols transmitted through the 1st and 2nd antennas, respectively. With two consecutive OFDM symbols $\mathbf{s}_{2i}^{(k)}$ and $\mathbf{s}_{2i+1}^{(k)}$, the modified Alomouti's STBC is given by (Zhang, 2014) $\begin{bmatrix}\overline{\mathbf{s}}_{2i}^{(k)} & \overline{\mathbf{s}}_{2i+1}^{(k)}\\ \overline{\overline{\mathbf{s}}}_{2i}^{(k)} & \overline{\overline{\mathbf{s}}}_{2i+1}^{(k)}\end{bmatrix}=\begin{bmatrix}\mathbf{s}_{2i}^{(k)} & -\mathbf{J}_N\mathbf{s}_{2i+1}^{(k)\,*}\\ \mathbf{s}_{2i+1}^{(k)} & \mathbf{J}_N\mathbf{s}_{2i}^{(k)*}\end{bmatrix}$ where $\mathbf{J}_N=[\boldsymbol{\omega}_1\boldsymbol{\omega}_N\boldsymbol{\omega}_{N-1}\cdots\boldsymbol{\omega}_2]$ and $\boldsymbol{\omega}_i$ is the i-th column of $\mathbf{I}_N$. Let $\mathbf{x}_i^{(k)},\overline{\mathbf{x}}_i^{(k)}$ and $\overline{\overline{\mathbf{x}}}_i^{(k)}$ be the inverse discrete Fourier transforms (IDFT) of $\mathbf{s}_i^{(k)},\overline{\mathbf{s}}_i^{(k)}$ and $\overline{\overline{\mathbf{s}}}_i^{(k)}$, respectively. After the IDFT operations, we have $\begin{bmatrix}\overline{\mathbf{x}}_{2i}^{(k)} & \overline{\mathbf{x}}_{2i+1}^{(k)}\\ \overline{\overline{\mathbf{x}}}_{2i}^{(k)} & \overline{\overline{\mathbf{x}}}_{2i+1}^{(k)}\end{bmatrix}=\begin{bmatrix}\mathbf{x}_{2i}^{(k)} & -\mathbf{x}_{2i+1}^{(k)\,*}\\ \mathbf{x}_{2i+1}^{(k)} & \mathbf{x}_{2i}^{(k)*}\end{bmatrix}$. Let $\mathbf{H}_i(n)\in C^{J\times K}$ be the channel coefficients from the i-th transmit antenna of all users to the receive antennas, $\mathbf{H}_i=\left[\mathbf{H}_i^T(0)\cdots\mathbf{H}_i^T(L_{ZP})\right]^T$, L and L_{ZP} be the lengths of channel impulse response (CIR) and ZP, respectively. Without loss of generality, we assume $L_{ZP}=L$. Collecting the signals among all the receive antennas yields $\mathbf{y}_i(n)=\left[y_i^{(1)}(n)\cdots y_i^{(J)}(n)\right]^T$ and $\mathbf{y}_i=\left[\mathbf{y}_i^T(0)\cdots\mathbf{y}_i^T(P-1)\right]^T$ where P=N+L. Then the two consecutive signal vectors $\mathbf{y}_{2i}$ and $\mathbf{y}_{2i+1}$ are given respectively by

$$\mathbf{y}_{2i}=\mathcal{T}_{L+1,N}(\mathbf{H}_1)\mathbf{x}_{2i}+\mathcal{T}_{L+1,N}(\mathbf{H}_2)\mathbf{x}_{2i+1}+\mathbf{w}_{2i} \quad (1)$$

$$\mathbf{y}_{2i+1}=\mathcal{T}_{L+1,N}(\mathbf{H}_2)\mathbf{x}_{2i}^*-\mathcal{T}_{L+1,N}(\mathbf{H}_1)\mathbf{x}_{2i+1}^*+\mathbf{w}_{2i+1} \quad (2)$$

where $\mathbf{w}_i$ is the additive white Gaussian noise (AWGN) vector, $\mathbf{x}_i=\left[\mathbf{x}_i^T(0)\cdots\mathbf{x}_i^T(N-1)\right]^T$, $\mathbf{x}_i(n)=\left[x_i^{(1)}(n)\cdots x_i^{(J)}(n)\right]^T$. Let $\overline{\mathbf{y}}_i=\left[\overline{\mathbf{y}}_i^T(0)\cdots\overline{\mathbf{y}}_i^T(P-1)\right]^T$, $\overline{\mathbf{y}}_i(m)=\left[\mathbf{y}_{2i}^T(m)\ \ \mathbf{y}_{2i+1}^H(m)\right]^T$, $\overline{\mathbf{x}}_i=\left[\overline{\mathbf{x}}_i^T(0)\cdots\overline{\mathbf{x}}_i^T(N-1)\right]^T$, $\overline{\mathbf{x}}_i(m)=\left[\mathbf{x}_{2i}^T(m)\ \ \mathbf{x}_{2i+1}^T(m)\right]^T$, $\mathbf{H}=[\mathbf{H}^T(0)\cdots\mathbf{H}^T(L)]^T$, and $\mathbf{H}(l)=\begin{bmatrix}\mathbf{H}_1(l) & \mathbf{H}_2(l)\\ \mathbf{H}_2^*(l) & -\mathbf{H}_1^*(l)\end{bmatrix}$. Then integrating (1) and (2) with reshuffling the order of sequences yields $\overline{\mathbf{y}}_i$ as follows,

$$\overline{\mathbf{y}}_i=\mathcal{T}_{L+1,N}(\mathbf{H})\overline{\mathbf{x}}_i+\overline{\mathbf{w}}_i \quad (3)$$

Based on (3), the authors in (Zhang, 2014) developed a subspace method to estimate the channel coefficients. In the following, we will propose a fast convergence subspace method based on the repetition index scheme.

3 PROPOSED METHODS

The repetition index scheme has the following property (Pan, 2013)

$$\begin{bmatrix} \bar{\mathbf{y}}_i \\ \mathbf{0}_{2Jq\times 1} \end{bmatrix} = \mathcal{T}_{L+1,N+q}(\mathbf{H}) \begin{bmatrix} \bar{\mathbf{x}}_i \\ \mathbf{0}_{2Kq\times 1} \end{bmatrix}$$
$$\begin{bmatrix} \mathbf{0}_{2Jq\times 1} \\ \bar{\mathbf{y}}_i \end{bmatrix} = \mathcal{T}_{L+1,N+q}(\mathbf{H}) \begin{bmatrix} \mathbf{0}_{2Kq\times 1} \\ \bar{\mathbf{x}}_i \end{bmatrix} \quad (4)$$

With the repetition index property in (4), we define a composite signal matrix

$$\mathcal{T}_{P,Q}(\bar{\mathbf{y}}_i) = \mathcal{T}_{L+1,N+Q-1}(\mathbf{H})\mathcal{T}_{N,Q}(\bar{\mathbf{x}}_i) + \mathcal{T}_{P,Q}(\bar{\mathbf{w}}_i), \quad (5)$$

where the parameter Q is referred as the repetition index. We can see that there are Q times equivalent signal vectors generated from $\bar{\mathbf{y}}_i$ with identical channel matrix $\mathcal{T}_{L+1,N+Q-1}(\mathbf{H})$. Assuming that the CIR remains unchanged during the transmission of N_S STBC OFDM blocks $\bar{\mathbf{y}}_i$, $i=1,...,N_S$, we can stack the received signals as

$$\begin{aligned}\bar{\mathbf{Y}}_Q^{(N_s)} &= [\mathcal{T}_{P,Q}(\bar{\mathbf{y}}_1)\cdots\mathcal{T}_{P,Q}(\bar{\mathbf{y}}_{N_S})]_{2J(P+Q-1)\times QN_s} \\ &= \mathcal{T}_{L+1,N+Q-1}(\mathbf{H})\bar{\mathbf{X}}_Q^{(N_s)} + \bar{\mathbf{W}}_Q^{(N_s)}\end{aligned} \quad (6)$$

where $\bar{\mathbf{X}}_Q^{(N_s)} = [\mathcal{T}_{N,Q}(\bar{\mathbf{x}}_1)\cdots\mathcal{T}_{N,Q}(\bar{\mathbf{x}}_{N_S})]_{2K(N+Q-1)\times QN_s}$ and $\bar{\mathbf{W}}_Q^{(N_s)} = [\mathcal{T}_{P,Q}(\bar{\mathbf{w}}_1)\cdots\mathcal{T}_{P,Q}(\bar{\mathbf{w}}_{N_S})]_{2J(P+Q-1)\times QN_s}$. Since $\mathcal{T}_{L+1,N+Q-1}(\mathbf{H})$ is a $2J(P+Q-1)\times 2K(N+Q-1)$ block Toeplitz matrix, it is of full column rank if $J \geq K$. Besides, a necessary condition for $\bar{\mathbf{X}}_Q^{(N_s)}$ having full row rank is $N_s \geq 2K(N+Q-1)/Q$. If neglecting the channel noise in (6) with the above two assumptions, we can derive the noise subspace, $\mathbf{U}_n$, from the eigenvalue decomposition (EVD) of the correlation matrix $\bar{\mathbf{Y}}_Q^{(N_s)}\bar{\mathbf{Y}}_Q^{(N_s)H}$ (Yu, 2009), (Zhang, 2014), (Pan, 2013). The noise subspace $\mathbf{U}_n \in \mathbb{C}^{2J(P+Q-1)\times\alpha}$, $\alpha=2J(P+Q-1)-2K(N+Q-1)$, is orthogonal complement to the channel matrix $\mathcal{T}_{L+1,N+Q-1}(\mathbf{H})$ such that $\mathbf{U}_n^H\mathcal{T}_{L+1,N+Q-1}(\mathbf{H}) = \mathbf{0}$. Let $\mathbf{u}_g$ be the g-th column of $\mathbf{U}_n$. Then $\mathbf{u}_g^H\mathcal{T}_{L+1,N+Q-1}(\mathbf{H}) = \mathbf{0}$ can be represented by

$$\mathcal{H}_{L+1,P+Q-1}^H(\mathbf{u}_g)\mathbf{H} = \mathbf{0} \quad (7)$$

When the channel noise is considered, the estimated noise subspace will be deviated from the true one and the homogeneous equation in (7) might be solved by the least square method. Let $\mathbf{H}=[\mathbf{h}_1 \ldots \mathbf{h}_{2K}]$. The channel estimation can be computed by the following optimization problem

$$\begin{aligned}\hat{\mathbf{H}} &= \arg\min_{\|\mathbf{h}_k\|=1} \sum_{g=1}^{\alpha}\sum_{k=1}^{2K}\left\|\mathcal{H}_{L+1,P+Q-1}^H(\mathbf{u}_g)\mathbf{h}_k\right\|^2 \\ &= \arg\min_{\|\mathbf{h}_k\|=1} \sum_{k=1}^{2K}\mathbf{h}_k^H\mathbf{\Psi}\,\mathbf{h}_k\end{aligned} \quad (8)$$

where $\mathbf{\Psi} = \sum_{g=1}^{\alpha}\mathcal{H}_{L+1,P+Q-1}(\mathbf{u}_g)\mathcal{H}_{L+1,P+Q-1}^H(\mathbf{u}_g)$. The solution of the optimization problem in (8) is the eigenvectors of ψ corresponding to the first $2K$ smallest eigenvalues. Because of the essence of blind subspace method, the values of $\hat{\mathbf{H}}$ in (8) differ those of $\mathbf{H}$ from a $2K\times 2K$ ambiguity matrix. For simplicity, we denote the method in (8) as 'ZP-RIS'.

4 FORWARD-BACKWARD TECHNIQUE

The forward-backward technique is developed to increase the equivalent received symbols from (1) and (2) in this section. We first stack $\mathbf{y}_{2i}$ and $\mathbf{y}_{2i+1}$ in an alternative way

$$\begin{bmatrix} \mathbf{y}_{2i+1} \\ -\mathbf{y}_{2i}^* \end{bmatrix} = \begin{bmatrix} \mathcal{T}_{L+1,N}(\mathbf{H}_1) & \mathcal{T}_{L+1,N}(\mathbf{H}_2) \\ \mathcal{T}_{L+1,N}(\mathbf{H}_2^*) & -\mathcal{T}_{L+1,N}(\mathbf{H}_1^*) \end{bmatrix} \begin{bmatrix} -\mathbf{x}_{2i+1}^* \\ \mathbf{x}_{2i}^* \end{bmatrix} + \begin{bmatrix} \mathbf{w}_{2i+1} \\ -\mathbf{w}_{2i}^* \end{bmatrix} \quad (9)$$

The symbol in (9) can be treated as another STBC OFDM received signal. Let $\underline{\mathbf{y}}_i = \left[\underline{\mathbf{y}}_i^T(0)\cdots\underline{\mathbf{y}}_i^T(P-1)\right]^T$, $\underline{\mathbf{y}}_i(m) = \left[\mathbf{y}_{2i+1}^T(m)\ \ \text{-}\mathbf{y}_{2i}^H(m)\right]^T$, $\underline{\mathbf{x}}_i = \left[\underline{\mathbf{x}}_i^T(0)\cdots\underline{\mathbf{x}}_i^T(N-1)\right]^T$, $\underline{\mathbf{x}}_i(m) = \left[-\mathbf{x}_{2i+1}^H(m)\ \ \mathbf{x}_{2i}^H(m)\right]^T$. Then reshuffling the order of sequences in (9) yields $\underline{\mathbf{y}}_i$ as follows,

$$\underline{\mathbf{y}}_i = \mathcal{T}_{L+1,N}(\mathbf{H})\underline{\mathbf{x}}_i + \underline{\mathbf{w}}_i \quad (10)$$

The vector in (10) is referred as the backward STBC symbol. Applying the repetition index technique onto the backward STBC symbol yields

$$\mathcal{T}_{P,Q}(\underline{\mathbf{y}}_i) = \mathcal{T}_{L+1,N+Q-1}(\mathbf{H})\mathcal{T}_{N,Q}(\underline{\mathbf{x}}_i) + \mathcal{T}_{P,Q}(\underline{\mathbf{w}}_i) \quad (11)$$

From (11), we obtain a signal matrix similar to (6)

$$\begin{aligned}\underline{\mathbf{Y}}_Q^{(N_s)} &= [\mathcal{T}_{P,Q}(\underline{\mathbf{y}}_1)\cdots\mathcal{T}_{P,Q}(\underline{\mathbf{y}}_{N_S})]_{2J(P+Q-1)\times QN_s} \\ &= \mathcal{T}_{L+1,N+Q-1}(\mathbf{H})\underline{\mathbf{X}}_Q^{(N_s)} + \underline{\mathbf{W}}_Q^{(N_s)}\end{aligned} \quad (12)$$

Using (6) and (12), the forward-backward technique generates the signal matrix by

$$\mathbf{Y}_Q^{(N_s)} = \left[\bar{\mathbf{Y}}_Q^{(N_s)} \ \underline{\mathbf{Y}}_Q^{(N_s)}\right] = \mathcal{T}_{L+1,N+Q-1}(\mathbf{H})\mathbf{X}_Q^{(N_s)} + \mathbf{W}_Q^{(N_s)} \quad (13)$$

where $\mathbf{X}_Q^{(N_s)} = [\bar{\mathbf{X}}_Q^{(N_s)} \ \underline{\mathbf{X}}_Q^{(N_s)}]$. A necessary condition for $\mathbf{X}_Q^{(N_s)}$ having full row rank is $N_s \geq K(N+Q-1)/Q$. That means the forward-backward technique can converge faster than the methods using $\bar{\mathbf{Y}}_Q^{(N_s)}$ or $\underline{\mathbf{Y}}_Q^{(N_s)}$, respectively. The channel estimation using the forward-backward technique is called 'ZP-RIS-FBM'.

5 NOISE PREWHITENING TECHNIQUE

The noise effect on the repetition index technique is discussed in this section. We observe that the repetition index technique makes the noise matrix $\mathcal{T}_{P,Q}(\bar{\mathbf{w}}_i)$ nonwhite even though the noise vector $\bar{\mathbf{w}}_i$ is white, i.e., $E\{\bar{\mathbf{w}}_i\bar{\mathbf{w}}_j^H\} = \sigma_w^2\delta_{ij}\mathbf{I}_{2PJ}$. Using the definition of the block Toeplitz matrix, the correlation matrix of $\mathcal{T}_{P,Q}(\bar{\mathbf{w}}_i)$ is calculated by

$$E\{\mathcal{T}_{P,Q}(\bar{\mathbf{w}}_i)\mathcal{T}_{P,Q}(\bar{\mathbf{w}}_i)^H\} \triangleq \sigma_w^2\mathbf{R}_{pre} = \sigma_w^2[diag(1,2,\ldots,R-1,\underbrace{R,\ldots,R}_{(|P-Q|+1)'s},R-1,\ldots,2,1)\otimes\mathbf{I}_{2J}] \quad (14)$$

where $R=\min(P,Q)$. Since $\mathbf{R}_{pre}$ is a diagonal matrix, it can be decomposed as $\mathbf{R}_{pre}=\mathbf{L}\mathbf{L}^H$ where $\mathbf{L}$ is given by

$$\mathbf{L} = diag(1,\ldots,\sqrt{R-1},\underbrace{\sqrt{R},\ldots,\sqrt{R}}_{(|P-Q|+1)'s},\sqrt{R-1},\ldots,1)\otimes\mathbf{I}_{2J} \quad (15)$$

From (15), we define the prewhitening matrix as $\mathbf{L}^{-1}$ and a prewhitening signal matrix $\mathcal{T}_{P,Q}^{(L)}(\mathbf{z}_i) = \mathbf{L}^{-1}\mathcal{T}_{P,Q}(\mathbf{z}_i)$ for $\mathbf{z}_i$ is a proper size signal vector. Then signal matrix in (5) can be prewhitened as

$$\mathcal{T}_{P,Q}^{(L)}(\bar{\mathbf{y}}_i) = \mathcal{T}_{L+1,N+Q-1}^{(L)}(\mathbf{H})\mathcal{T}_{N,Q}(\bar{\mathbf{x}}_i) + \mathcal{T}_{P,Q}^{(L)}(\bar{\mathbf{w}}_i) \quad (16)$$

where $\mathcal{T}_{L+1,N+Q-1}^{(L)}(\mathbf{H}) = \mathbf{L}^{-1}\mathcal{T}_{L+1,N+Q-1}(\mathbf{H})$. We can see that $E\{\mathcal{T}_{P,Q}^{(L)}(\bar{\mathbf{w}}_i)\mathcal{T}_{P,Q}^{(L)H}(\bar{\mathbf{w}}_i)\} = \sigma_w^2\mathbf{I}_{2(P+Q-1)J}$. Consequently, the noise subspace calculated from $\mathcal{T}_{P,Q}^{(L)}(\bar{\mathbf{y}}_i)$ will get lower perturbations than that from $\mathcal{T}_{P,Q}(\bar{\mathbf{y}}_i)$. Let $\bar{\mathbf{Y}}_Q^{(N_s,L)} = \mathbf{L}^{-1}\bar{\mathbf{Y}}_Q^{(N_s)}$ and $\mathbf{u}_g^{(L)}$, $g=1,\cdots,\alpha$ be the eigenvectors corresponding to the first α smallest eigenvalues of $\bar{\mathbf{Y}}_Q^{(N_s,L)}\bar{\mathbf{Y}}_Q^{(N_s,L)H}$. Thus we have $\mathbf{u}_g^{(L)H}\mathcal{T}_{L+1,N+Q-1}^{(L)}(\mathbf{H}) = \mathbf{u}_g^{(L)H}\mathbf{L}^{-1}\mathcal{T}_{L+1,N+Q-1}(\mathbf{H}) = \mathbf{0}$ because those eigenvectors are orthogonal to the channel matrix $\mathcal{T}_{L+1,N+Q-1}^{(L)}(\mathbf{H})$. The channel estimate can be obtained from (7) and (8) by letting $\mathbf{u}_g = \mathbf{L}^{-H}\mathbf{u}_g^{(L)}$. Similar method can be applied to $\underline{\mathbf{Y}}_Q^{(N_s)}$ and $\mathbf{Y}_Q^{(N_s)}$ to get improvements in channel estimation.

6 SIMULATION RESULTS

Computer simulations are given to demonstrate the superiority of the proposed subspace channel estimation for STBC ZP-OFDM systems. The number of subcarrier is N = 32 or 64, the CP length is L = 8, the ZP length is L_{zp} = 12, 16-QAM modulation schemes is applied, and the channel impulse responses is assumed to be i.i.d. complex Gaussian random variable with zero mean and unit variance. The signal-to-noise ratio (SNR) is defined as

$$SNR = \frac{E[\|\bar{\mathbf{y}}_i - \bar{\mathbf{w}}_i\|_F^2]}{E[\|\bar{\mathbf{w}}_i\|_F^2]} = \frac{\sigma_s^2 NK(L+1)}{P\sigma_n^2} \quad (17)$$

The performance metric of the channel estimation is the normalized mean-square error (NMSE) is given by

$$NMSE = \frac{1}{N_m}\sum_{i=1}^{N_m}\frac{\|\hat{\mathbf{H}}(i) - \mathbf{H}(i)\|_F^2}{\|\mathbf{H}(i)\|_F^2} \quad (18)$$

where N_m is the number of Monte-Carlo trials, $\mathbf{H}(i)$ and $\hat{\mathbf{H}}(i)$ are the true channel of the i-th trial and the estimated channel after ambiguity correction, respectively. The notations 'ZP-2009' and 'ZP-2009-FBM' denote the channel estimation method developed in (Yu, 2009), 'ZP-CRM' and 'ZP-CFBM' are used to indicate the fast subspace channel estimation in (Zhang, 2014), and 'PreW' represents the noise prewhitening technique. It is noted that the numbers of equivalent STBC-OFDM symbols are N_S, $2N_S$, NN_S and $2NN_S$ for 'ZP-2009', 'ZP-2009-FBM', 'ZP-CRM' and 'ZP-CFBM', and QN_S and $2QN_S$ for 'ZP-RIS' and 'ZP-RIS-FBM', respectively

Fig. 2 shows the NMSE versus the input SNR when N_s = 100, N=32 and Q=8. Since $\bar{\mathbf{y}}_i$ is a $2JN\times1$ vector, the minimum number of STBC-OFDM symbols required for subspace channel estimation is 192. Thus only the method 'ZP-2009' is failed to work but 'ZP-2009-FBM' works very well. The CRM and CFBM using the overlap-and-add (OLA) scheme slowly reduce the NMSE and obtain a higher NMSE than the 'ZP-2009-FBM' as the SNR increases. The proposed RIS and RIS-FBM methods derive much lower NMSE than the compared ones. The effect of the prewhitening technique is not recognizable for Q=8 since the nonwhite noise in (14) is close to white. We further examine the NMSE when N=64, Q=16 and N_s = 100 in Fig. 3. In this scenario, both the 'ZP-2009' and 'ZP-2009-FBM' are failed to work. The effect of nonwhite noise is evident for the 'ZP-RIS' method and the prewhitening technique alleviates the effect considerably. The proposed channel estimations still outperform the compared ones.

Fig. 4 shows the NMSE versus the number of STBC-OFDM symbols when SNR=20dB, N=32 and Q=8. Both the CRM and the CFBM methods reduce the NMSEs very slowly when N_s increases. The 'ZP-2009' and the 'ZP-2009-FBM' decrease the NMSEs very sharply after N_s >180 and N_s >90, respectively. On the other hand, the NMSEs of the proposed RIS and RIS-FBM methods have dropped quickly when N_s >20 and N_s >10, respectively. Fig. 5 shows the NMSE versus the number of STBC-OFDM symbols when SNR=20dB, N=64 and Q=16. We can see that the proposed methods outperform the other ones. When the input SNR is small, the nonwhite noise drastically influences the performance of the subspace channel estimation and prewhitening technique improves this effectively.

7 CONCLUSIONS

In this paper, we proposed a RIS method to enhance the convergence of the performance of subspace channel estimation in STBC ZP-OFDM systems. Based on the STBC property, a FBM method generating twice equivalent STBC OFDM symbols as the RIS is presented. The RIS method turns the white noise into nonwhite. The prewhitening technique is developed to alleviate the nonwhite effect. Computer Simulations showed the proposed methods reduced the NMSEs quickly with few STBC OFDM symbols.

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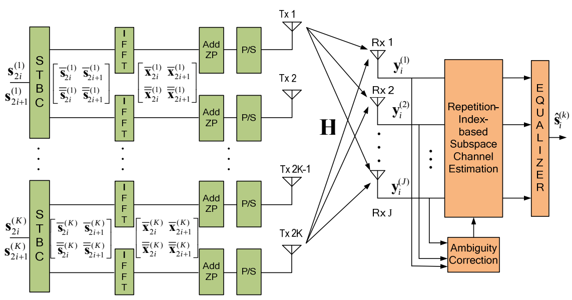

Figure 1: STBC-ZP-OFDM systems.

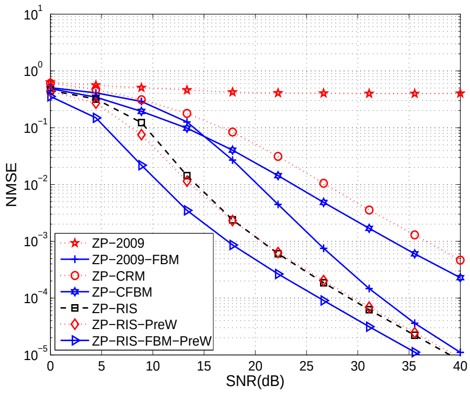

Figure 2: NMSE versus the input SNR when N=32, Q=8, 16-QAM modulation is used, N_S=100 STBC OFDM, K = 2, J = 3, and N_m = 1000.

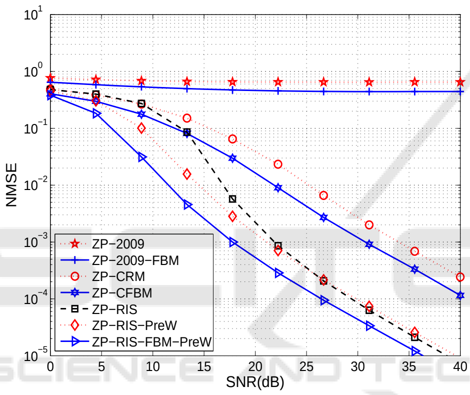

Figure 3: Scenario is the same as that in Fig. 2 except for N=64, Q=16.

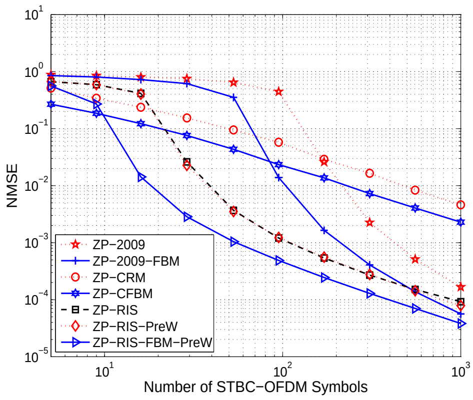

Figure 4: NMSEs versus the number of STBC-OFDM symbols when N=32, Q=8, 16-QAM modulation is used, input SNR=20dB, K = 2, J = 3, and N_m = 1000.

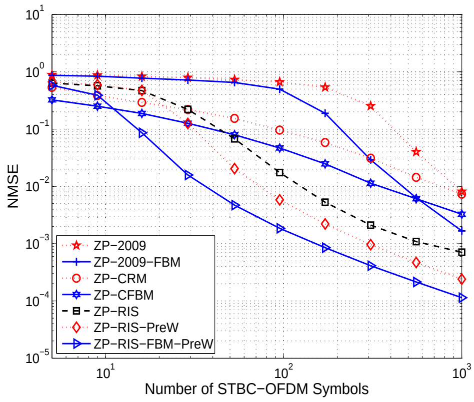

Figure 5: Scenario is the same as that in Fig. 4 except for N=64, Q=16.