Adaptive Two-Dimension MIMO Channel Estimation for Single-Carrier STBC Time-Division Duplex Transmission

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Abstract— **Recently, we proposed a multi-block frequency-domain equalization (MB-FDE) for single-carrier space-time block coded time-division duplex (SC-STBC TDD) transmission. In this paper, we propose an adaptive two-dimension (2D) multi-input multi-output (MIMO) channel estimation (CE) for SC-STBC TDD transmission in a high mobility environment. In the proposed adaptive 2D-MIMO-CE, the channel over consecutive data blocks is directly estimated by using jointly optimized multiple 2D filters, each associated with each pilot block, based on minimum mean square error (MMSE) criterion. We evaluate, by computer simulation, the BER performance when using SC-STBC TDD transmission with MB-FDE and the adaptive 2D MIMO-CE and showed that the adaptive 2D MIMO-CE can achieve better BER performance than when using MMSE-CE and MMSE prediction.**

Keywords-component; Channel estimation, space-time block coding, single-carrier transmission

I. INTRODUCTION

In the next generation wireless communications, broadband data services are demanded even in a high mobility environment. Single-carrier transmission has an advantage of lower peak-to-average power ratio (PAPR) compared to orthogonal frequency-domain multiplexing (OFDM) transmission and the use of frequency-domain equalization (FDE) achieves a good BER performance [1-3]. Space-time block coding (STBC) transmit diversity [4-9] is also an effective scheme to improve the BER performance. There are two types of STBC transmit diversity: space-time transmit diversity (STTD) [5] and space-time block coded joint transmit/receive diversity (STBC-JTRD) [6]. The STBC-JTRD requires the channel state information (CSI) at the transmitter side while the STTD requires it at the receiver side.

Recently, we proposed SC-STBC time-division duplex (TDD) transmission system [7], in which the frequency-domain (FD)-STTD [8] and FD-STBC-JTRD [9] are used for the uplink and downlink transmissions, respectively. Only the base station (BS) requires the CSI and hence, the complexity problem of mobile terminal (MT) can be alleviated. Since the same frequency is used in TDD for the uplink and downlink, the CSI estimate for the uplink reception can be reused for the downlink transmission. No CSI feedback is required between MT and BS. However, when the channel varies within a STBC codeword, the BER performance significantly degrades due to the interference caused by the orthogonality distortion of the STBC codeword [10]. To remedy this problem, recently, we proposed multi-block (MB)-FDE for SC-STBC-TDD transmission [11,12]. Multiple FDE weight matrices used in MB-FDE, each associated with each data block in a STBC codeword, are jointly optimized based on MMSE criterion by taking into account the channel variation in a STBC codeword. We showed that SC-STBC TDD transmission with MB-FDE can achieve a good BER performance even in a high mobility environment. However, the past study of SC-STBC TDD transmission with MB-FDE assumed the perfect knowledge of channel state information (CSI).

In a high mobility environment, channel estimation (CE) accuracy significantly degrades. Recently, 2-step MMSE prediction-based CE (2-step MMSE-CE) was studied [13]. 2-step MMSE-CE estimates the channel at the pilot block in the first step and then, estimates the channel at the data block by using the idea of MMSE prediction [14]. However, channel tracking ability is still degraded due to CSI error caused in the first step.

In this paper, we propose adaptive two-dimension multi-input multi-output channel estimation (2D-MIMO-CE) for SC-STBC TDD transmission in a high mobility environment. In the proposed adaptive 2D-MIMO-CE, the channel over the consecutive data blocks is directly estimated by using jointly optimized multiple 2D filters, each associated with each pilot block, based on MMSE criterion. Furthermore, iterative processing is applied to achieve higher tracking ability [15]. We evaluate, by computer simulation, that the BER performance when using SC-STBC TDD transmission with the iterative adaptive 2D MIMO-CE and showed that the proposed iterative adaptive 2D MIMO-CE can achieve better BER performance than when using 2-step MMSE-CE.

II. SC-STBC TDD TRANSMISSION

In this paper, we consider SC-STBC TDD transmission. We assume that BS equips with N_{BS} antennas and MT has N_{MT} antennas, respectively. Fig. 1 illustrates the frame structure considered in this paper. Fig. 2 and 3 show the transmitter/receiver structures. In SC-STBC TDD transmission, both CE and FDE are performed at BS in order to make MT structure simple and eliminate CSI feedback. The uplink and downlink data frames consist of N_B data blocks and the pilot frame consists of N_p pilot blocks, respectively. For uplink transmission, MT transmits the uplink data frame and two pilot frames to BS. After estimating the channel at the uplink data frame by applying the iterative adaptive 2D MIMO-CE, BS detects the uplink data signal by performing the receive MB-FDE and

STBC decoding. Then, BS estimates the channel at the downlink data frame by using the adaptive 2D MIMO-CE. After STBC encoding and the transmit MB-FDE, BS transmits the downlink data frame to MT.

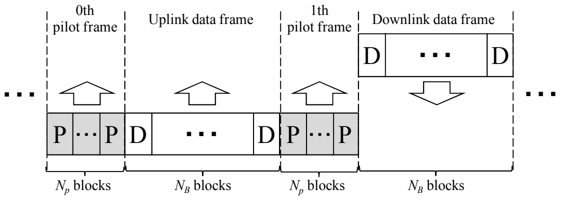
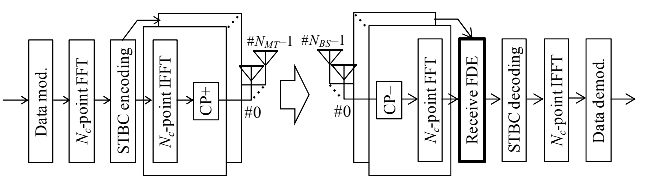
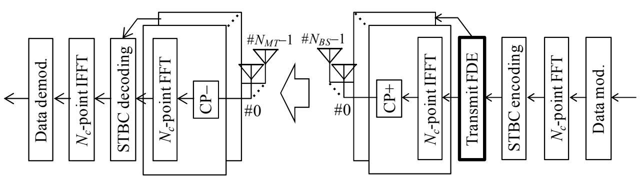


Fig. 1. Frame structure



(a) Mobile terminal (b) Base station

Fig. 2. Uplink transmitter/receiver structure



(a) Mobile terminal (b) Base station

Fig. 3. Downlink transmitter/receiver structure

A. Received signal reprensentation

In this subsection, the behavior of BS and MT and the received signal representation in SC-STBC TDD transmission are only introduced due to page limitation. The detail of the transmit signal representation and MB-FDE are referred in our previous study [11,12].

(a) Pilot frame

Before and after transmitting the uplink data frame, MT transmits the pilot frame to BS. The n_pth frequency-domain received pilot signal vector in the pth pilot frame, $\mathbf{R}_{P,p,n_p}(k)=\left[R_{P,p,n_p}(0,k),\ \cdots,\ R_{P,p,n_p}(N_{BS}-1,k)\right]^T$, can be expressed as

$$\mathbf{R}_{P,p,n_p}(k)=\mathbf{H}_{P,p,n_p}(k)\mathbf{P}_{P,p,n_p}(k)+\mathbf{N}_{P,p,n_p}(k), \qquad (1)$$

where $\mathbf{P}_{P,p,n_p}(k)=\left[P_{P,p,n_p}(0,k),\ \cdots,\ P_{P,p,n_p}(N_{MT}-1,k)\right]^T$ is the n_pth frequency-domain transmit pilot signal vector in the pth pilot frame. $\mathbf{H}_{P,p,n_p}(k)=\left[\mathbf{H}_{P,p,n_p}(0,k),\ \cdots,\ \mathbf{H}_{P,p,n_p}(N_{MT}-1,k)\right]$ with $\mathbf{H}_{P,p,n_p}(n_{MT},k)=\left[H_{P,p,n_p}(0,n_{MT},k),\ \cdots,\ H_{P,p,n_p}(N_{BS}-1,n_{MT},k)\right]^T$ is the $N_{BS}\times N_{MT}$ channel transfer function matrix at the n_pth received pilot block in the pth pilot frame. $\mathbf{N}_{P,p,n_p}(k)=\left[N_{P,p,n_p}(0,k),\ \cdots,\ N_{P,p,n_p}(N_{BS}-1,k)\right]^T$ is the noise vector and $N_{P,p,n_p}(n_{BS},k)$ denotes the zero mean additive white Gaussian noise (AWGN) having variance $2N_0/T_s$ with N_0 and T_s being the single-sided power spectrum density of AWGN and the symbol duration, respectively.

(b) Uplink data frame

At the MT transmitter, the $J\times N_c$ data modulated symbols are divided into the a sequence of J block of N_c symbol each. The J transmit signal blocks are transformed into the frequency-domain signal by the N_c-point fast Fourier transform (FFT). A sequence of J frequency-domain signals are encoded into N_{MT} streams of Q STBC coded frequency-domain single blocks each. The frequency-domain STBC codeword is transformed back to the time-domain STBC codeword by applying N_c-point inverse FFT (IFFT). After insertion of cyclic prefix (CP) into the beginning the each block, the $N_U(=N_B/Q)$ STBC codewords are transmitted to BS during N_B time-slot.

At the BS receiver, a super position of N_{MT} transmitted signal is received by N_{BS} antennas. After CP removal, the received signal is transformed into the frequency-domain signal by N_c-point FFT. The qth frequency-domain received signal vector, $\mathbf{Y}_{U,u,q}(k)=[Y_{U,u,q}(0,k),\ldots,Y_{U,u,q}(N_{BS}-1,k)]^T$, in the uth STBC codeword can be expressed as

$$\mathbf{Y}_{U,u,q}(k)=\mathbf{H}_{U,u,q}(k)\mathbf{X}_{U,u,q}(k)+\mathbf{N}_{U,u,q}(k), \qquad (2)$$

where $\mathbf{X}_{U,u,q}(k)=[X_{U,u,q}(0,k),\ldots,X_{U,u,q}(N_{MT}-1,k)]^T$ is the qth frequency-domain transmit signal vector in the uth STBC codeword. $\mathbf{H}_{U,u,q}(k)=[\mathbf{H}_{U,u,q}(0,k),\ldots,\mathbf{H}_{U,u,q}(N_{MT}-1,k)]$ with $\mathbf{H}_{U,u,q}(n_{MT},k)=[H_{U,u,q}(n_{MT},0,k),\ldots,H_{U,u,q}(n_{MT},N_{BS}-1,k)]^T$ is the $N_{BS}\times N_{MT}$ channel transfer function matrix at the qth received signal block in the uth STBC codeword. $\mathbf{N}_{U,u,q}(k)=[N_{U,u,q}(0,k),\ldots,N_{U,u,q}(N_{BS}-1,k)]^T$ is the noise vector and $N_{U,u,q}(n_{BS},k)$ is AWGN having variance $2N_0/T_s$. At BS receiver, the channel at the uplink data frame is estimated by using the adaptive 2D MIMO-CE (the detail is introduced in next section). Then, BS performs the received MB-FDE and STBC decoding. The frequency-domain STBC decoded signal is transformed back to the time-domain signal by N_c-point IFFT, and finally, data demodulation is carried out.

(c) Downlink data frame

At the BS transmitter, the $J\times N_c$ data modulated symbols are divided into the a sequence of J block of N_c symbol each. The J transmit signal blocks are transformed into the frequency-domain signal by the N_c-point FFT. A sequence of J frequency-domain signals are encoded into N_{MT} streams of Q STBC coded frequency-domain single blocks each. After STBC encoding, the transmit MB-FDE is performed. The frequency-domain STBC codeword after the transmit MB-FDE is transformed back to the time-domain STBC codeword by applying N_c-point IFFT. After insertion of CP into the beginning the each block, the $N_D(=N_B/Q)$ STBC codewords are transmitted to MT during N_B time-slot.

At the MT receiver, a super position of N_{BS} transmitted signal is received by N_{MT} antennas. After CP removal, the received signal is transformed into the frequency-domain signal by N_c-point FFT. The qth frequency-domain received signal vector $\mathbf{Y}_{D,d,q}(k)=[Y_{D,d,q}(0,k),\ldots,Y_{D,d,q}(N_{MT}-1,k)]^T$, in the dth STBC codeword can be expressed as

$$\mathbf{Y}_{D,d,q}(k)=\mathbf{H}_{D,d,q}(k)\mathbf{X}_{D,d,q}(k)+\mathbf{N}_{D,d,q}(k), \qquad (3)$$

where $\mathbf{X}_{D,d,q}(k)=[X_{D,d,q}(0,k),...,X_{D,d,q}(N_{BS}-1,k)]^T$ denotes the qth frequency-domain transmit signal vector in the dth STBC codeword. $\mathbf{H}_{D,d,q}(k)=\left[\mathbf{H}_{D,d,q}(0,k),\ \cdots,\ \mathbf{H}_{D,d,q}(N_{BS}-1,k)\right]$ with $\mathbf{H}_{D,d,q}(n_{BS},k)=[H_{D,d,q}(0,n_{BS},k),\ldots,H_{D,q}(N_{MT}-1,n_{BS},k)]^T$ is the $N_{MT}\times N_{BS}$ channel transfer function matrix at the qth received signal block in the dth STBC codeword. $\mathbf{N}_{D,d,q}(k)=[N_{D,d,q}(0,k),\ldots,N_{D,d,q}(N_{MT}-1,k)]^T$ is the noise vector and $n_{D,q}(n_{MT},k)$ is AWGN having variance $2N_0/T_s$. STBC decoding is performed to obtain the STBC decoded frequency-domain signal. The STBC decoded frequency-domain signal is transformed back to the time-domain signal by N_c-point IFFT, and finally, the data demodulation is carried out.

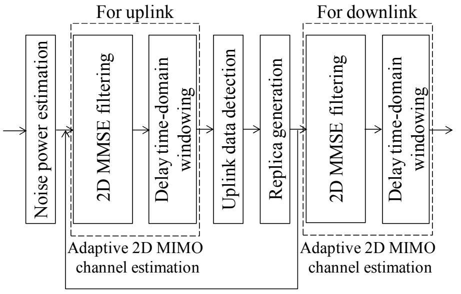


Fig. 4. Flowchart of the iterative adaptive 2D MIMO-CE

III. ITERATIVE ADAPTIVE 2D MIMO-CE

In this paper, we propose the iterative adaptive 2D MIMO-CE for SC-STBC TDD transmission in a high mobility environment. In the proposed iterative adaptive 2D MIMO-CE, the channel at the uplink and downlink data frames are directly estimated by using jointly optimized multiple 2D filters, each associated with each pilot frame, based on MMSE criterion. Furthermore, iterative processing is applied to achieve higher tracking ability [15].

Fig. 4 shows the flowchart of the iterative adaptive 2D MIMO-CE. After estimating the noise variance by applying least square (LS)-CE [16] to pilot frames, BS estimates the channel at the uplink data frame by applying 2D filtering to pilot frames and the delay time-domain windowing [17]. BS performs the received MB-FDE and STBC decoding by using CSI estimates. Then BS computes the log-likelihood ratio (LLR) of the uplink transmit signal and generates the uplink transmit signal replicas [18]. Then, viewing the uplink transmit signal replicas as pilot signals, BS re-estimates the channel at the uplink data frame by applying 2D filtering to both the pilot frames and the uplink data frame and the delay time-domain windowing. The above processing are repeated by I iterations. After I iterations, BS de-modulates the uplink transmit signal. Then, viewing the de-modulated and re-modulated uplink transmit signals as pilot signals, BS estimates the channel at the downlink data frame by applying 2D filtering to both the pilot frames and the uplink data frame and the delay time-domain windowing.

A. Channel estimation for uplink data frame

(a) In the 0th iteration (First channel estimation)

In the 0th iteration, the channel at the uplink data frame is estimated by applying 2D filtering to two pilot data frames. The CSI estimates, $\hat{\mathbf{H}}^{(0)}_{U,u,q}(k)$, for the qth received signal block in the uth STBC codeword are given as

$$\hat{\mathbf{H}}^{(0)}_{U,u,q}(k)=\sum_{p=0}^{1}\tilde{\mathbf{R}}_{P,p}(k)\mathbf{W}^{(0)}_{P,p}(k), \tag{4}$$

where $\tilde{\mathbf{R}}_{P,p}(k)=\left[\mathbf{R}_{P,p,0}(k)\ \cdots\ \mathbf{R}_{P,p,N_p-1}(k)\right]$ is the $N_{BS}\times N_p$ received pilot signal matrix. $\mathbf{W}^{(0)}_{P,p}(k)$ is the $N_p\times N_{MT}$ 2D filter for the pth pilot frame in the 0th iteration. They are jointly optimized so as to minimize the mean square error (MSE) between the actual channel and the CSI estimate given as

$$\left\{\mathbf{W}^{(0)}_{P,0}(k)\ \ \mathbf{W}^{(0)}_{P,1}(k)\right\}=\arg\min E\left[\left\|\mathbf{H}_{U,u,q}(k)-\hat{\mathbf{H}}^{(0)}_{U,u,q}(k)\right\|^2\right]. \tag{5}$$

From the orthogonal projection theorem [14], the adaptive MMSE 2D filters satisfy the following equation given as

$$\tilde{\mathbf{C}}_{(U,u,q)\leftrightarrow(P,p')}\tilde{\mathbf{P}}_{P,p'}(k)=\sum_{p=0}^{1}\left(\mathbf{W}^{(0)}_{P,p}(k)\right)^H\begin{bmatrix}\tilde{\mathbf{P}}^H_{P,p}(k)\tilde{\mathbf{C}}_{(P,p)\leftrightarrow(P,p')}\tilde{\mathbf{P}}_{P,p'}(k)\\+2\sigma^2\mathbf{Z}_{(P,p)\leftrightarrow(P,p')}\end{bmatrix},$$

$$\text{for } p'=0,1 \tag{6}$$

where $\tilde{\mathbf{P}}_{P,p}(k)=diag\left\{\mathbf{P}_{P,p,0}(k)\ \cdots\ \mathbf{P}_{P,p,N_p-1}(k)\right\}$ is the $N_{MT}N_p\times N_p$ transmit pilot signal matrix. $\tilde{\mathbf{C}}_{(U,u,q)\leftrightarrow(P,p')}=E\left[\mathbf{H}^H_{U,u,q}(k)\tilde{\mathbf{H}}_{P,p'}(k)\right]$ is the $N_{MT}\times N_{MT}N_p$ expand cross–correlation matrix and $\tilde{\mathbf{C}}_{(P,p)\leftrightarrow(P,p')}=E\left[\tilde{\mathbf{H}}^H_{P,p}(k)\tilde{\mathbf{H}}_{P,p'}(k)\right]$ is the $N_{MT}N_p\times N_{MT}N_p$ expand self-correlation matrix of channel, respectively, with denoting $\tilde{\mathbf{H}}_{P,p}(k)=\left[\mathbf{H}_{P,0}(k)\ \cdots\ \mathbf{H}_{P,N_p-1}(k)\right]$ as the $N_{BS}\times N_{MT}N_p$ expand channel transfer function matrix at the pth pilot frame. $\mathbf{Z}_{(P,p)\leftrightarrow(P,p')}$ is the noise correlation matrix and given as

$$\mathbf{Z}_{(P,p)\leftrightarrow(P,p')}=\begin{cases}2\sigma^2\mathbf{I}_{N_p} & \text{if} \quad p'=p\\ \mathbf{0} & \text{otherwise}\end{cases}, \tag{7}$$

where $2\sigma^2$ is the noise variance and $\mathbf{I}_N$ denotes the $N\times N$ identity matrix. In the iterative adaptive 2D-MIMO-CE, the noise variance is estimated by applying LS-CE [16] to two pilot frames. By solving (6), the adaptive MMSE 2D filters, $\hat{\mathbf{W}}^{(0)}(k)=\left[\left(\mathbf{W}^{(0)}_{P,0}(k)\right)^T\ \ \left(\mathbf{W}^{(0)}_{P,1}(k)\right)^T\right]^T$, in the 0th iteration are obtained as

$$\hat{\mathbf{W}}^{(0)}(k)=\left[\hat{\mathbf{P}}^H(k)\mathbf{A}^{(0)}\hat{\mathbf{P}}(k)+2\sigma^2\mathbf{I}_{2N_p}\right]^{-1}\hat{\mathbf{P}}^H(k)\mathbf{B}^{(0)}, \tag{8}$$

where

$$\begin{cases}\hat{\mathbf{P}}(k)=diag\left[\tilde{\mathbf{P}}_{P,0}(k)\ \tilde{\mathbf{P}}_{P,1}(k)\right]\\ \mathbf{A}^{(0)}=\begin{bmatrix}\tilde{\mathbf{C}}_{(P,0)\leftrightarrow(P,0)} & \tilde{\mathbf{C}}_{(P,0)\leftrightarrow(P,1)}\\ \tilde{\mathbf{C}}_{(P,1)\leftrightarrow(P,0)} & \tilde{\mathbf{C}}_{(P,1)\leftrightarrow(P,1)}\end{bmatrix}\\ \mathbf{B}^{(0)}=\left[\tilde{\mathbf{C}}_{(U,u,q)\leftrightarrow(P,0)}\ \ \tilde{\mathbf{C}}_{(U,u,q)\leftrightarrow(P,1)}\right]^T\end{cases}. \tag{9}$$

(b) In the ith (i=1,...,I–1) iteration

In the ith (i=1,...,I–1) iteration, viewing the transmit signal replicas generated in the (i–1) iteration as pilot signals, the channel at the uplink data frame is re-estimated by applying 2D filtering to both the pilot frames and the uplink data frame. The CSI estimates, $\hat{\mathbf{H}}^{(i)}_{U,u,q}(k)$, for the qth received signal block in the uth STBC codeword is given as

$$\hat{\mathbf{H}}^{(i)}_{U,u,q}(k)=\sum_{p=0}^{1}\tilde{\mathbf{R}}_{P,p}(k)\mathbf{W}^{(i)}_{P,p}(k)+\sum_{u=0}^{N_U-1}\tilde{\mathbf{Y}}_{U,u}(k)\mathbf{W}^{(i)}_{U,u}(k), \tag{10}$$

where $\tilde{\mathbf{Y}}_{U,u}(k)=\left[\mathbf{Y}_{U,u,0}(k)\ \cdots\ \mathbf{Y}_{U,u,Q-1}(k)\right]$ is the $N_{BS}\times Q$ received signal matrix. $\mathbf{W}^{(i)}_{P,p}(k)$ is the $N_p\times N_{MT}$ 2D filter for the pth pilot frame and $\mathbf{W}^{(i)}_{U,u}(k)$ is the $Q\times N_{MT}$ 2D filter for the uth STBC codeword, respectively. They are jointly optimized so as to minimize MSE between the actual channel and the channel estimates given as

$$\left\{\begin{array}{c}\mathbf{W}^{(i)}_{P,0}(k)\ \mathbf{W}^{(i)}_{P,1}(k)\\ \mathbf{W}^{(i)}_{U,0}(k)\cdots\mathbf{W}^{(i)}_{U,N_U-1}(k)\end{array}\right\}=\arg\min E\left[\left\|\mathbf{H}_{U,u,q}(k)-\hat{\mathbf{H}}^{(i)}_{U,u,q}(k)\right\|^2\right]. \tag{11}$$

From the orthogonal projection theorem [14], the adaptive MMSE 2D filters satisfy the following equation given as

$$\begin{cases}\tilde{\mathbf{C}}_{(U,u,q)\leftrightarrow(P,p')}\tilde{\mathbf{P}}_{P,p'}(k)=\displaystyle\sum_{p=0}^{1}\left(\mathbf{W}^{(i)}_{P,p}(k)\right)^H\begin{bmatrix}\tilde{\mathbf{P}}^H_{P,p}(k)\tilde{\mathbf{C}}_{(P,p)\leftrightarrow(P,p')}\tilde{\mathbf{P}}_{P,p'}(k)\\ +2\sigma^2\mathbf{Z}_{(P,p)\leftrightarrow(P,p')}\end{bmatrix}\\ \qquad\qquad+\displaystyle\sum_{u=0}^{N_U-1}\left(\mathbf{W}^{(i)}_{U,u}(k)\right)^H\left[\left(\tilde{\mathbf{X}}^{(i)}_{U,u}(k)\right)^H\tilde{\mathbf{C}}_{(U,u)\leftrightarrow(P,p')}\tilde{\mathbf{P}}_{U,p'}(k)\right]\\ \tilde{\mathbf{C}}_{(U,u,q)\leftrightarrow(U,u')}\tilde{\mathbf{X}}^{(i)}_{U,u'}(k)=\displaystyle\sum_{u=0}^{N_U-1}\left(\mathbf{W}^{(i)}_{U,u}(k)\right)^H\begin{bmatrix}\left(\tilde{\mathbf{X}}^{(i)}_{U,u}(k)\right)^H\tilde{\mathbf{C}}_{(U,u)\leftrightarrow(U,u')}\tilde{\mathbf{X}}^{(i)}_{U,u'}(k)\\ +2\sigma^2\mathbf{Z}_{(U,u)\leftrightarrow(U,u')}\end{bmatrix}\\ \qquad\qquad+\displaystyle\sum_{p=0}^{1}\left(\mathbf{W}^{(0)}_{P,p}(k)\right)^H\left[\tilde{\mathbf{P}}^H_{P,p}(k)\tilde{\mathbf{C}}_{(P,p)\leftrightarrow(U,u')}\tilde{\mathbf{X}}^{(i)}_{U,u'}(k)\right]\end{cases}$$

for p'=0,1, u'=0,…,N_U–1, (12)

where $\tilde{\mathbf{X}}^{(i)}_{U,u}(k)=diag\left\{\overline{\mathbf{X}}^{(i)}_{U,u,0}(k)\ \cdots\ \overline{\mathbf{X}}^{(i)}_{U,u,Q-1}(k)\right\}$ is the $N_{MT}Q\times Q$ transmit signal replica matrix with denoting $\overline{\mathbf{X}}^{(i)}_{U,u,q}(k)$ as the qth transmit signal replica in the uth STBC codeword. They are generated by applying STBC encoding to the soft-decision symbol replica computed from LLR in the (i–1)th iteration [18]. $\tilde{\mathbf{C}}_{(U,u,q)\leftrightarrow(U,u')}=E\left[\mathbf{H}^H_{U,u,q}(k)\tilde{\mathbf{H}}_{U,u'}(k)\right]$ is the $N_{MT}\times N_{MT}Q$ expand cross–correlation matrix and $\tilde{\mathbf{C}}_{(U,u)\leftrightarrow(U,u')}=E\left[\tilde{\mathbf{H}}^H_{U,u}(k)\tilde{\mathbf{H}}_{U,u'}(k)\right]$ is the $N_{MT}Q\times N_{MT}Q$ expand self-correlation matrix of channel, respectively, with denoting $\tilde{\mathbf{H}}_{U,u}(k)=\left[\mathbf{H}_{U,u,0}(k)\ \cdots\ \mathbf{H}_{U,u,Q-1}(k)\right]$ as the $N_{BS}\times N_{MT}Q$ expand channel transfer function matrix at the uth STBC codeword. $\mathbf{Z}_{(U,u)\leftrightarrow(U,u')}$ is the noise correlation matrix and given as

$$\mathbf{Z}_{(U,u)\leftrightarrow(U,u')}=\begin{cases}2\sigma^2\mathbf{I}_Q & \text{if}\quad u'=u\\ \mathbf{0} & \text{otherwise}\end{cases}. \tag{13}$$

By solving (12), the adaptive MMSE 2D filters, $\hat{\mathbf{W}}^{(i)}(k)=\left[\left(\mathbf{W}^{(i)}_{P,0}(k)\right)^T\ \left(\mathbf{W}^{(i)}_{U,0}(k)\right)^T\ \cdots\ \left(\mathbf{W}^{(i)}_{U,N_U-1}(k)\right)^T\ \left(\mathbf{W}^{(i)}_{P,1}(k)\right)^T\right]^T$, in the ith iteration are obtained as

$$\hat{\mathbf{W}}^{(i)}(k)=\left[\left(\hat{\mathbf{X}}^{(i)}(k)\right)^H\mathbf{A}^{(i)}\hat{\mathbf{X}}^{(i)}(k)+2\sigma^2\mathbf{I}_{(2N_p+N_B)}\right]^{-1}\left(\hat{\mathbf{X}}^{(i)}(k)\right)^H\mathbf{B}^{(i)}, \tag{14}$$

where

$$\begin{cases}\hat{\mathbf{X}}^{(i)}(k)=diag\left[\tilde{\mathbf{P}}_{P,0}(k)\ \tilde{\mathbf{X}}^{(i)}_{U,0}(k)\ \cdots\ \tilde{\mathbf{X}}^{(i)}_{U,N_U-1}(k)\ \tilde{\mathbf{P}}_{P,1}(k)\right]\\ \mathbf{A}^{(i)}=\begin{bmatrix}\tilde{\mathbf{C}}_{(P,0)\leftrightarrow(P,0)} & \tilde{\mathbf{C}}_{(P,0)\leftrightarrow(U,0)} & \cdots & \tilde{\mathbf{C}}_{(P,0)\leftrightarrow(U,N_U-1)} & \tilde{\mathbf{C}}_{(P,0)\leftrightarrow(P,1)}\\ \tilde{\mathbf{C}}_{(U,0)\leftrightarrow(P,0)} & \tilde{\mathbf{C}}_{(U,0)\leftrightarrow(U,0)} & \cdots & \tilde{\mathbf{C}}_{(U,0)\leftrightarrow(U,N_U-1)} & \tilde{\mathbf{C}}_{(U,0)\leftrightarrow(P,1)}\\ \vdots & \vdots & \ddots & \vdots & \vdots\\ \tilde{\mathbf{C}}_{(U,N_U-1)\leftrightarrow(P,0)} & \tilde{\mathbf{C}}_{(U,N_U-1)\leftrightarrow(U,0)} & \cdots & \tilde{\mathbf{C}}_{(U,N_U-1)\leftrightarrow(U,N_U-1)} & \tilde{\mathbf{C}}_{(U,N_U-1)\leftrightarrow(P,1)}\\ \tilde{\mathbf{C}}_{(P,1)\leftrightarrow(P,0)} & \tilde{\mathbf{C}}_{(P,1)\leftrightarrow(U,0)} & \cdots & \tilde{\mathbf{C}}_{(P,1)\leftrightarrow(U,N_U-1)} & \tilde{\mathbf{C}}_{(P,1)\leftrightarrow(P,1)}\end{bmatrix}\\ \mathbf{B}^{(0)}=\left[\tilde{\mathbf{C}}_{(U,u,q)\leftrightarrow(P,0)}\ \tilde{\mathbf{C}}_{(U,u,q)\leftrightarrow(U,0)}\ \cdots\ \tilde{\mathbf{C}}_{(U,u,q)\leftrightarrow(U,N_U-1)}\ \tilde{\mathbf{C}}_{(U,u,q)\leftrightarrow(P,1)}\right]^T\end{cases}. \tag{15}$$

B. Channel estimation for downlink data frame

Viewing the transmit signal replicas generated in the (I–1)th iteration in the uplink channel estimation as pilot signals, the channel at the downlink data frame is estimated by applying 2D filtering to both the pilot frames and uplink data frame. The CSI estimates, $\hat{\mathbf{H}}_{D,d,q}(k)$, for the qth received signal block in the dth STBC codeword is given as

$$\hat{\mathbf{H}}_{D,d,q}(k)=\left(\sum_{p=0}^{1}\tilde{\mathbf{R}}_{P,p}(k)\mathbf{V}_{P,p}(k)+\sum_{u=0}^{N_U-1}\tilde{\mathbf{Y}}_{U,u}(k)\mathbf{V}_{U,u}(k)\right)^T, \tag{16}$$

$\mathbf{V}_{P,p}(k)$ is the $N_p\times N_{MT}$ 2D filter for the pth pilot frame and $\mathbf{V}_{U,u}(k)$ is the $Q\times N_{MT}$ 2D filter for the uth STBC codeword, respectively. They are jointly designed so as to minimize MSE between the actual channel and the channel estimates given as

$$\left\{\begin{array}{c}\mathbf{V}_{P,o}(k)\ \mathbf{V}_{P,1}(k)\\ \mathbf{V}_{U,0}(k)\ \cdots\ \mathbf{V}_{U,N_U-1}(k)\end{array}\right\}=\arg\min E\left[\left\|\mathbf{H}_{D,d,q}(k)-\hat{\mathbf{H}}_{D,d,q}(k)\right\|^2\right]. \tag{17}$$

By solving the above optimization problem similar to the uplink channel estimation, the adaptive MMSE 2D filters, $\mathbf{V}(k)=\left[\left(\mathbf{V}_{P,0}(k)\right)^T\ \left(\mathbf{V}_{U,0}(k)\right)^T\ \cdots\ \left(\mathbf{V}_{U,N_U-1}(k)\right)^T\ \left(\mathbf{V}_{P,1}(k)\right)^T\right]^T$ are obtained as

$$\hat{\mathbf{V}}(k)=\left[\left(\hat{\mathbf{X}}^{(I-1)}(k)\right)^H\mathbf{A}^{(I-1)}\hat{\mathbf{X}}^{(I-1)}(k)+2\sigma^2\mathbf{I}_{(2N_p+N_B)}\right]^{-1}\left(\hat{\mathbf{X}}^{(I-1)}(k)\right)^H\mathbf{D}, \tag{18}$$

where

$$\mathbf{D}=\left[\tilde{\mathbf{C}}_{(D,d,q)\leftrightarrow(P,0)}\ \tilde{\mathbf{C}}_{(D,d,q)\leftrightarrow(U,0)}\ \cdots\ \tilde{\mathbf{C}}_{(D,d,q)\leftrightarrow(U,N_U-1)}\ \tilde{\mathbf{C}}_{(D,d,q)\leftrightarrow(P,1)}\right]^T. \tag{19}$$

$\tilde{\mathbf{C}}_{(D,d,q)\leftrightarrow(P,p)}=E\left[\mathbf{H}^*_{D,d,q}(k)\tilde{\mathbf{H}}_{P,p}(k)\right]$ is the $N_{MT}\times N_{MT}N_p$ expand cross-correlation matrix of channel and $\tilde{\mathbf{C}}_{(D,d,q)\leftrightarrow(U,u)}=E\left[\mathbf{H}^*_{D,d,q}(k)\tilde{\mathbf{H}}_{U,u}(k)\right]$ is the $N_{MT}\times N_{MT}Q$ expand cross-correlation matrix of channel, respectively.

IV. COMPUTER SIMULATION

We consider QPSK data modulation. FFT block size N_c and CP length N_g are set to N_c=128 symbols and N_g=16 samples, respectively. The number of BS and MT antennas is respectively set to N_{BS}=2 and N_{MT}=3 as an example. The channel is assumed to be a time and frequency-selective fading channel having symbol spaced L=16 path uniform power delay profile. We assume the spatial correlation of channel is sufficiently low. Long M sequence is used as pilot signal. The pilot frame length N_p and the data frame length N_B are set to N_p=N_{MT} and N_B=8 blocks, respectively. The number of iterations for the iterative 2D MIMO-CE is set to I=10. In this paper, we assume the time correlation of channel can be perfectly estimated by using Jakes model [19].

A. Uplink BER performance

Fig. 5 shows the uplink BER performance when using SC-STBC TDD transmission with the iterative adaptive 2D MIMO-CE as a function of transmit signal energy per bit-to-AWGN power spectrum density ratio E_b/N_0. The normalized maximum Doppler frequency f_DT_s is assumed to be f_DT_s=5×

10^{-4}. For the comparison, the performances when using 2-step MMSE-CE, which estimates the channel at the pilot frame by using MMSE-CE [16] and then, estimates the channel at the data frame by using MMSE prediction, the iterative 2-step MMSE-CE, which is combination of 2-step MMSE-CE and iterative processing, ideal channel estimation case are also plotted in Fig. 5. Furthermore, the performances when using MB-FDE which is jointly optimized by taking into account the channel variation in a STBC codeword [11] and MB-FDE which is designed based on the assumption of quasi-static fading [12] (below, we call the former as MB-FDE with block optimization and the latter as MB-FDE without block optimization, respectively) are also plotted in Fig. 5. It is seen from Fig. 5 that the performance when using 2-step MMSE-CE has BER error floor and the performance gap between MB-FDE with block optimization and that without block optimization is disappeared. 2-step MMSE-CE estimates the channel at the pilot frame by using MMSE-CE in the first step and then, estimates the channel at the data frame by using the idea of MMSE prediction. Therefore, the channel tracking ability is limited by CSI error in the first step because MMSE-CE requires multiple pilot blocks to estimate a pair of MIMO channel. Furthermore, the performance when using the iterative 2-step MMSE-CE has also BER error floor. This is because 2-step MMSE-CE cannot obtain sufficient accuracy of the transmit signal replica due to CSI error. It is also seen from Fig. 5 that the iterative 2D MMSE-CE can achieve the BER performance similar to the ideal channel estimation case and obtain the effect of block optimization. This is because high tracking ability can be obtained by using jointly optimized multiple 2D filters.

Fig. 6 plots the BER performance when using SC-STBC TDD transmission with MB-FDE and the iterative adaptive MMSE-CE as a function of the normalized maximum Doppler frequency f_DT_s. The transmit E_b/N_0 is set to 8dB. For comparison, the performances when using 2-step MMSE-CE, iterative 2-step MMSE-CE, ideal channel estimation case are also plotted in Fig. 6. It is seen from Fig. 6 that the iterative adaptive 2D MIMO-CE is more robust to channel time-selectively than the 2-step MMSE-CE and the iterative 2-step MMSE-CE. When using MB-FDE with block optimization and the allowable BER is BER=10^{-3}, the iterative 2D MMSE-CE can tolerate about 3 times larger Doppler frequency than the iterative 2-step MMSE-CE. It is seen from Fig. 6 that, when using the iterative adaptive 2D MIMO-CE and the allowable BER is BER=10^{-3}, MB-FDE with block optimization can tolerate about 2.5 times larger Doppler frequency than MB-FDE without block optimization.

B. Downlink BER performance

Fig.7 and 8 plot the downlink BER performance when using SC-STBC TDD transmission with the iterative adaptive 2D MIMO-CE and MB-FDE. The normalized maximum Doppler frequency f_DT_s is assumed to be $f_DT_s=5\times10^{-4}$ in Fig. 7 and the transmit E_b/N_0 is set to 8dB in Fig. 8, respectively. It is seen from Fig. 7 and 8 that the iterative adaptive 2D MIMO-CE can achieve better BER performance than the 2-step MMSE-CE and the iterative 2-step MMSE-CE. However, the performance gap between when using the iterative adaptive 2D MIMO-CE and ideal channel estimation case is still large. This is because the cross-correlation of channels between the pilot frame and the downlink data frame is low and hence, sufficient CE accuracy cannot be obtained even if the iterative adaptive 2D MIMO-CE is used. Therefore, we have to study some techniques to improve the downlink BER performance as our future study.

C. Computational complexity

The complexity of the iterative adaptive 2D MIMO-CE is shown in Table 1. The complexity is defined as the number of complex multiply operations per a frame. The iterative adaptive 2D MIMO-CE need to compute the expand correlation matrices given by (9) and (15) and hence, it requires about 6 times higher complexity than the 2-step MMSE-CE although it does not require MMSE prediction. Improved BER performance of the iterative adaptive 2D MIMO-CE over the iterative 2-step MMSE-CE is obtained at the cost of computational complexity.

TABLE I. COMPUTATIONAL COMPLEXITY

	Iterative adaptive 2D MIMO-CE		Iterative 2-step MMSE-CE	
	Uplink CE	Downlink CE	Uplink CE	Downlink CE
2D filter computation	10,506,240	86,016	806,400	36964
2D filter multiplication	811,008	6240	101376	10752
MMSE prediction			788480	86016
Total	11,409,504		1,829,888	

V. CONCLUSION

In this paper, we proposed an adaptive 2D MIMO-CE for SC-STBC TDD transmission in high mobility environment. In the proposed iterative adaptive 2D MIMO-CE, the channel at the data frame is directly obtained by using jointly optimized multiple 2D filters, each associated with each pilot frame, based on MMSE criterion. It was shown by computer simulation that, when the allowable BER is BER=10^{-3}, the iterative adaptive 2D MIMO-CE can tolerate about 2.5 times higher Doppler frequency than the iterative 2-step MMSE-CE in uplink FD-STTD transmission. However, in downlink FD-STBC-JTRD transmission, the performance gap between when using the iterative adaptive 2D MIMO-CE and ideal channel estimation case is still large. Therefore, we have to study some techniques to further improve the downlink BER performance as our future study. In this paper, we assume the time correlation of channel can be perfectly estimated. The actual time correlation estimation technique is also left as our future work.

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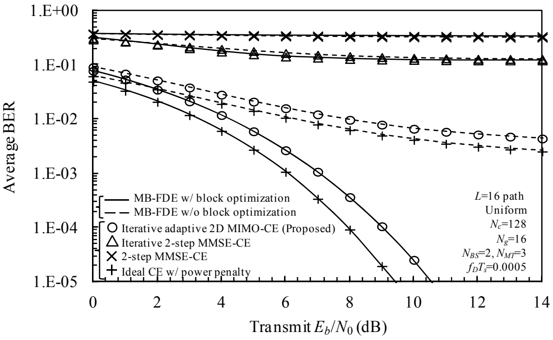

Fig. 5. Transmit E_b/N_0 v.s. uplink BER performance

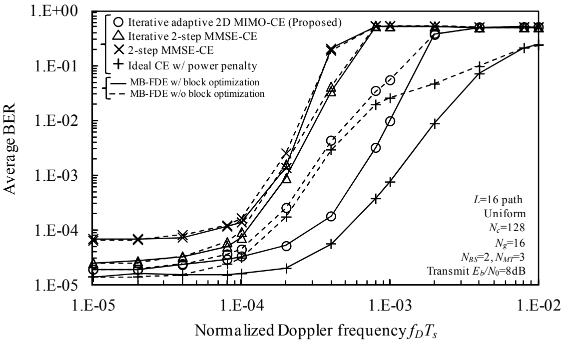

Fig. 6. f_DT_s v.s. uplink BER performance

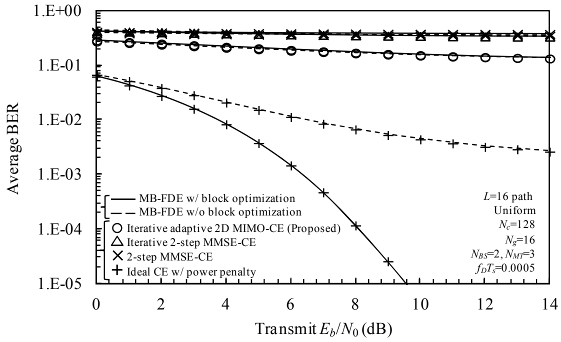

Fig. 7. Transmit E_b/N_0 v.s. downlink BER performance

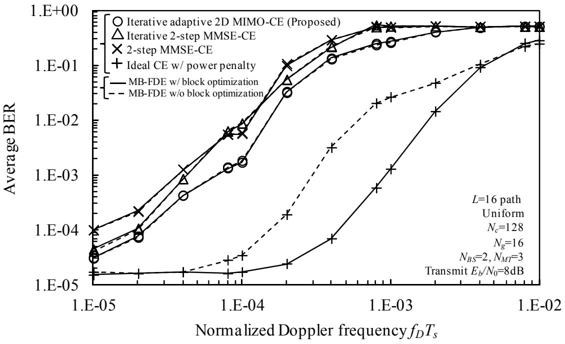

Fig. 8. f_DT_s v.s. downlink BER performance