Decision-feedback Prediction Channel Estimation for MIMO Cooperative Transmission

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***Abstract*— In this paper, the channel estimation (CE) scheme suitable for MIMO cooperative transmission using time division duplex (TDD) in a high mobility environment is considered. We propose a decision-feedback linear prediction (DF-LP) CE scheme, which is used during data transmission. The uncoded bit error rate (BER) performances of OFDM downlink using maximal ratio transmit diversity (MRTD) and single-carrier (SC) uplink using minimum mean square error combining diversity (MMSECD) are evaluated by computer simulation. It is confirmed that the proposed DF-LP CE scheme can significantly improve the BER performance in a very high mobility environment.**

Keywords—5G; MIMO; cooperative transmission; channel estimation; decision-feedback; linear prediction

I. INTRODUCTION

The development of the 5[th] generation (5G) mobile communications system is now underway, aiming at starting its communications services in around 2020 [1]. The multi-input and multi-output transmission technology [2-5] is a promising technique for enhanced mobile broadband communications. In this paper, we consider maximal-ratio transmit diversity (MRTD) [6] for orthogonal frequency division multiplexing (OFDM) downlink transmissions and the frequency-domain maximal-ratio combining diversity (MRCD)/minimum mean square error combining diversity (MMSECD) [7] for single-carrier (SC) uplink transmissions. Transmit power allocation (PA) is also considered. A conceptual structure of the OFDM downlink transmission system using MRTD is illustrated in Fig. 1. A large number of transmit/receive antennas are used at a base station (BS) while a few antennas are used at a user equipment (UE). Prior to data transmission, a predetermined number of BS antennas are selected. MIMO channel state information (CSI) must be shared between BS and UE prior to MIMO cooperative data transmission. This is done by MIMO channel estimation (CE). If the time-division duplex (TDD) is used, no CSI feedback is necessary due to the channel reciprocity. A subframe structure [8], consisting of uplink pilot time slot (UpPTS) and downlink PTS (DwPTS) followed by 12 data time slots (DTSs), is illustrated in Fig. 2. Frequency-division multiplexed (FDM) orthogonal pilots are used for pilot-aided MIMO CE at BS and UE.

The CSI of MIMO channel is estimated by using uplink and downlink pilots transmitted at the beginning of each subframe. Therefore, in a high mobility environment, the estimated CSI becomes outdated during the data slot period and hence, the transmission performance degrades [8]. In this paper, we introduce the linear prediction (LP) [9] into the CE algorithm and propose a decision-feedback LP (DF-LP) CE for CSI updating during a time interval of 12 DTSs.

This paper is organized as follows. Section II describes the pilot-aided CE prior to the data transmission. The DF-LP CE during data transmission is proposed in Sect. III. In Sect. IV, the uncoded bit error rate (BER) performance achievable with the proposed DF-LP CE in a high mobility environment is evaluated by computer simulation. Finally, Sect. V offers concluding remarks and a future study.

Throughout the paper, the numbers of BS antennas and UE antennas for data transmission are denoted by N_{bs} and N_{ue}, respectively. N_{bs} antennas are selected from $N_{bs'}$ ($\geq N_{bs}$) antennas based on CSI estimate obtained by using the uplink pilot. Sets of BS antennas, selected BS antennas, and UE antennas are denoted by $\mathbf{N}_{bs'}$, $\mathbf{N}_{bs}$, and $\mathbf{N}_{ue}$, respectively. The total number of subcarriers is denoted by N_c.

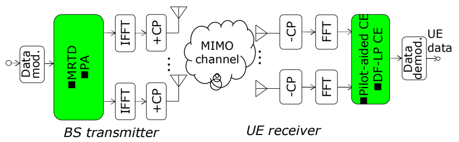


Fig. 1. OFDM downlink transmission using MRTD.

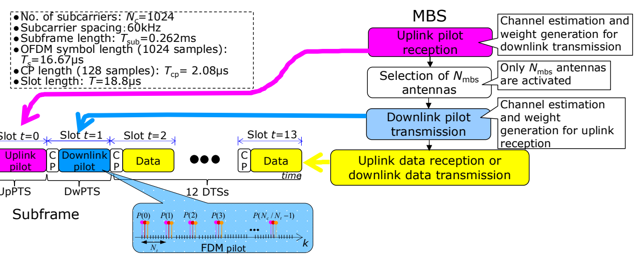


Fig. 2. Subframe structure and flowchart of pilot-aided MIMO CE before data transmission.

II. PILOT-AIDED MIMO CE

A flowchart of pilot-aided MIMO CE is illustrated in Fig. 2. Prior to the data transmission (slot time t=0 and 1), UE transmits orthogonal FDM pilots from N_{ue} antennas for BS to obtain the CSI estimate of $N_{bs'}\times N_{ue}$ and to select N_{bs} ($\leq N_{bs'}$) BS antennas from N_{bs} BS antennas. Then, BS transmits orthogonal FDM pilots from N_{bs} selected BS antennas for UE to obtain the CSI estimate. By doing so, BS and UE share the CSI of $N_{bs}\times N_{ue}$ MIMO channel due to the channel reciprocity when TDD is used. An equally spaced pilot subcarrier arrangement is employed for orthogonal FDM pilots. Knowing the fact that

N_{ue} is generally much smaller than $N_{\mathrm{bs'}}$, transmitting the uplink pilots is done first.

During UpPTS period (slot time t=0), FDM orthogonal uplink pilots are transmitted from N_{ue} UE antennas. By receiving those FDM uplink pilots, MBS estimates the CSI $\{H_{mn}(k;t=0); m\in \mathbf{N}_{\mathrm{bs'}}, n\in \mathbf{N}_{\mathrm{ue}}\}$ of $N_{\mathrm{bs'}}\times N_{\mathrm{ue}}$ MIMO channel between $N_{\mathrm{bs'}}$ BS antennas and N_{ue} UE antennas. The CSI estimate is denoted as $\{\hat{H}_{mn}(k;t=0); m\in \mathbf{N}_{\mathrm{bs'}}, n\in \mathbf{N}_{\mathrm{ue}}\}$. Then, N_{bs} BS antennas having the largest received pilot powers are selected for data transmission during 12 DTSs period.

MBS computes the MRT weights $\{W_m(k); m\in \mathbf{N}_{\mathrm{bs}}\}$ and the PA weight $\Omega(k)$ for downlink transmission and the MRCD/MMSECD weight $\{W_m(k); m\in \mathbf{N}_{\mathrm{bs}}\}$ for uplink reception, based on CSI estimate $\{\hat{H}_{mn}(k;t=0); m\in \mathbf{N}_{\mathrm{bs}}, n\in \mathbf{N}_{\mathrm{ue}}\}$ of $N_{\mathrm{bs}}\times N_{\mathrm{ue}}$ MIMO channel.

During DwPTS period (slot time t=1), BS transmits orthogonal FDM pilots from the selected N_{bs} BS antennas, which are received by UE to obtain the CSI estimate $\{\hat{H}_{mn}(k;t=1); m\in \mathbf{N}_{\mathrm{bs}}, n\in \mathbf{N}_{\mathrm{ue}}\}$ of $N_{\mathrm{bs}}\times N_{\mathrm{ue}}$ MIMO channel. Then, UE computes the equivalent channel gain $H_e(k;t)$ if the current subframe is used for downlink data transmission while it computes the PA weight $\Omega(k)$ if the current subframe is used for uplink data transmission. The equivalent channel gain is described in Section III.

III. DF-LP CE for CSI Updating

Here, N_{ue}=1 antenna is assumed for the simplicity. $N_{\mathrm{bs}}\times 1$ MRTD and PA are done at BS for OFDM downlink transmission while PA at UE and $1\times N_{\mathrm{bs}}$ MRCD/MMSECD are done at BS for SC uplink transmission. A flowchart of DF-LP CE is illustrated in Fig. 3. During downlink data transmission (slot time t=2~13), the equivalent channel gain at UE is updated while keeping the MRT weight and PA weight at BS unchanged. On the other hand, during uplink data transmission (slot time t=2~13), the MRCD/MMSECD weight is updated while keeping the PA weight unchanged.

A. OFDM downlink using MRTD

At slot time t (=2~13), BS transforms an N_c-symbol data block $\{d(n;t); n=0\sim N_c-1\}$ to be transmitted into the frequency-domain signal $\{d(k;t); k=0\sim N_c-1\}$ after serial-to-parallel conversion and then, multiplies the MRT weight $W_m(k)$ and the PA weight $\Omega(k)$ to it, where variable k denotes the k(=0~N_c-1)th subcarrier. The resultant frequency-domain signal $\left\{\sqrt{2S\cdot\Omega(k)}W_m(k)d(k;t); k=0\sim N_c-1, m\in \mathbf{N}_{\mathrm{bs}}\right\}$ is transformed back to the time-domain OFDM signal by an N_c-point inverse discrete Fourier transform (IDFT) for downlink transmission over an $N_{\mathrm{bs}}\times 1$ MISO channel, where S denotes the average transmit power per subcarrier and $E[|d(k;t)|^2]=1$. BS computes $W_m(k)$ and $\Omega(k)$ using $N_{\mathrm{bs}}\times 1$ MISO channel estimate $\{\hat{H}_m(k;t=0); m\in \mathbf{N}_{\mathrm{bs}}\}$ obtained by the uplink pilot at slot time t=0, as follows.

$$W_m(k)=\frac{\hat{H}_m^*(k;t=0)}{\sqrt{\sum_{m\in \mathbf{N}_{\mathrm{bs}}}\left|\hat{H}_m(k;t=0)\right|^2}}\quad \text{MRT weight}\ , \tag{1}$$

$$\Omega(k)=\frac{1}{\sum_{m\in \mathbf{N}_{\mathrm{bs}}}\left|\hat{H}_m(k;t=0)\right|^2}\Bigg/\frac{1}{N_c}\sum_{k=0}^{N_c-1}\left(\frac{1}{\sum_{m\in \mathbf{N}_{\mathrm{bs}}}\left|\hat{H}_m(k;t=0)\right|^2}\right). \tag{2}$$

The above $W_m(k)$ and $\Omega(k)$ are used over slot time t=2~13.

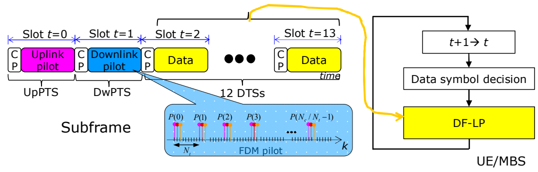


Fig. 3. A flowchart of DF-LP CE during data transmission.

The received OFDM signal at UE is transformed by an N_c-point DFT to the frequency-domain signal $\{R(k;t); n=0\sim N_c-1\}$ and then, data symbol decision is carried out. The received kth subcarrier component $R(k;t)$ can be written as

$$R(k;t)=\sqrt{2S}H_e(k;t)d(k;t)+N(k;t)\,, \tag{3}$$

where $H_e(k;t)$ is the equivalent channel gain given as

$$\begin{aligned}H_e(k;t)&=\sqrt{\Omega(k)}\left(\sum_{m\in \mathbf{N}_{\mathrm{bs}}}W_m(k)H_m(k;t)\right)\\&=\begin{cases}\sum_{m\in \mathbf{N}_{\mathrm{bs}}}\hat{H}_m^*(k;t=0)H_m(k;t)\Bigg/\sqrt{\sum_{m\in \mathbf{N}_{\mathrm{bs}}}\left|\hat{H}_m(k;t=0)\right|^2} & \text{w/o PA}\\ \dfrac{\sum_{m\in \mathbf{N}_{\mathrm{bs}}}\hat{H}_m^*(k;t=0)H_m(k;t)}{\sum_{m\in \mathbf{N}_{\mathrm{bs}}}\left|\hat{H}_m(k;t=0)\right|^2}\Bigg/\sqrt{\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}\left(\dfrac{1}{\sum_{m\in \mathbf{N}_{\mathrm{bs}}}\left|\hat{H}_m(k;t=0)\right|^2}\right)} & \text{w/ PA}\end{cases}\end{aligned}\ . \tag{4}$$

Here $E[|H_m(k;t)|^2]=1$ and $N(k;t)$ represents the zero-mean complexed-valued Gaussian noise component having variance of $2N_0/T$ with N_0 being the single-sided power spectrum density of additive white Gaussian noise (AWGN) and T being the DFT block length in time.

Symbol decision is done as

$$\hat{d}(k;t)=\min_{d\in \mathbf{D}}\left|R(k;t)-\sqrt{2S}\hat{H}_e(k;t)d\right|,\quad 2\le t\le 13\,, \tag{5}$$

where $\hat{H}_e(k;t)$ represents the equivalent channel gain estimate and d and $\mathbf{D}$ denote the symbol candidate and set of candidate symbols, respectively. Finally, UE outputs the received symbol block $\{\hat{d}(n;t); n=0\sim N_c-1\}$ after parallel-to-serial conversion. The equivalent channel gain estimate $\hat{H}_e(k;t=2)$ to be used for symbol decision at slot time t=2 can be computed from the CSI obtained using the uplink pilot, as

$$\hat{H}_e(k;t=2) = \begin{cases} \dfrac{\sum_{m\in\mathbf{N}_{bs}} \hat{H}_m^*(k;t=1)\hat{H}_m(k;t=1)}{\sqrt{\sum_{m\in\mathbf{N}_{bs}} \left|\hat{H}_m(k;t=1)\right|^2}} & \text{MRTD w/o PA} \\ \dfrac{\sum_{m\in\mathbf{N}_{mbs}} \hat{H}_m^*(k;t=1)\hat{H}_m(k;t=1)}{\sum_{m\in\mathbf{N}_{bs}} \left|\hat{H}_m(k;t=1)\right|^2} \Big/ \sqrt{\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}\left(\dfrac{1}{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t=1)\right|^2}\right)} & \text{MRTD w/ PA} \end{cases} . \quad (6)$$

Although $\hat{H}_e(k;t=2)$ can be used during slot time t=2~13 in a low mobility environment, it may deviate from the actual equivalent channel gain in a high mobility environment. Since the channel changes over a time interval of 12 DTSs, the downlink BER performance degrades. Therefore, we introduce LP of the equivalent channel gain by using both the decision result $\{\hat{d}(n;t-1);n=0\sim N_c-1\}$ and the UE received signal block $\{R(k;t-1)\}$ of the previous slot time t-1. The following simple 1st order and 2nd order LP can be used at UE side.

$$\begin{cases} \hat{H}_e(k;t=3) = \tilde{H}_e(k;t=2) \quad :1^{st} \text{ order LP}, \ t=3 \\ \hat{H}_e(k;t) = \begin{cases} \tilde{H}_e(k;t-1) \quad :1^{st} \text{ order LP} \\ 2\tilde{H}_e(k;t-1) - \tilde{H}_e(k;t-2) \quad , t=4\sim 13 \\ \quad :2^{nd} \text{ order LP} \end{cases} \end{cases} . \quad (7)$$

where $\tilde{H}_e(k;t)$ is the moving-averaged equivalent channel gain estimate, computed using

$$\tilde{H}_e(k;t) = \frac{1}{Q}\sum_{l=-(Q-1)/2}^{+(Q-1)/2} \frac{R(k+l;t)}{\hat{d}(k+l;t)}, \ t=2\sim 13 \cdot \quad (8)$$

There may exist the optimum value in Q, which will be determined by computer simulation.

B. SC uplink using MRCD/MMSECD

Either MRCD or MMSECD can be used at BS. In SC transmission, the inter-symbol interference (ISI) cannot be neglected. When using MRCD, the frequency-selectivity of the equivalent channel after MRCD is enhanced and accordingly, ISI is also enhanced. Therefore, PA needs to be introduced at UE to sufficiently suppress the ISI. On the other hand, when using MMSECD, the frequency-selectivity of the equivalent channel can be weakened and hence, PA is not necessary, i.e., $\Omega(k)=1$.

At slot time t (=2~13), UE transforms an N_c-symbol data block $\{d(n;t);n=0\sim N_c-1\}$ to be transmitted into the frequency-domain signal $\{D(k;t);k=0\sim N_c-1\}$ by an N_c-point DFT as follows.

$$D(k;t) = \frac{1}{\sqrt{N_c}}\sum_{n=0}^{N_c-1} d(n;t)\exp\left(-j2\pi k\frac{n}{N_c}\right) \cdot \quad (9)$$

Then, after multiplying the PA weight $\Omega(k)$, UE transforms $\{D(k;t)\}$ back to the time-domain signal block by an N_c-point IDFT for uplink transmission over an $1\times N_{bs}$ SIMO channel. Each received signal $\{R_m(k;t);k=0\sim N_c-1\}$ at the mth MBS antenna is transformed by an N_c-point DFT into the frequency-domain signal. After performing MRCD/MMSECD [7], the diversity-combined time-domain received signal $\{r(n;t);n=0\sim N_c-1\}$ is obtained by an N_c-point IDFT.

BS computes the MRCD/MMSECD weight $\{W_m(k);m\in\mathbf{N}_{bs}\}$ using the $1\times N_{bs}$ SIMO channel estimate $\{\hat{H}_m(k;t=0);m\in\mathbf{N}_{bs}\}$ obtained by the uplink pilot at slot time t=0, as follows.

$$W_m(k) = \begin{cases} \dfrac{\hat{H}_m^*(k;t=0)}{\sqrt{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t=0)\right|^2}} & \text{MRC weight} \\ \dfrac{\hat{H}_m^*(k;t=0)}{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t=0)\right|^2 + (S\cdot T/N_0)^{-1}} & \text{MMSEC weight} \end{cases} . \quad (10)$$

On the other hand, UE computes $\Omega(k)$ using the $1\times N_{bs}$ SIMO channel estimate $\{\hat{H}_m(k;t=1);m\in\mathbf{N}_{bs}\}$ obtained by receiving the downlink pilot at slot time t=1, as follows.

$$\Omega(k) = \begin{cases} \dfrac{1}{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t=1)\right|^2} \Big/ \dfrac{1}{N_c}\sum_{k=0}^{N_c-1}\left(\dfrac{1}{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t=1)\right|^2}\right) & \text{for MRCD} \\ 1 & \text{for MMSECD} \end{cases} , \quad (11)$$

which is continuously used for uplink data transmission during slot time t=2~13.

When using PA at UE and MRCD/MMSECD at BS, the frequency-domain received signal $R(k;t)$ after diversity combining at BS can be expressed as

$$\begin{aligned} R(k;t) &= \sum_{m\in\mathbf{N}_{bs}} W_m(k)R_m(k;t) \\ &= \sqrt{2S}H_e(k,t)D(k;t) + \sum_{m\in\mathbf{N}_{bs}} W_m(k)N_m(k;t) \end{aligned} , \quad (12)$$

where $R_m(k;t)$ represents the frequency-domain received signal on the mth BS antenna, $H_e(k;t)$ is the equivalent channel gain, and $D(k;t)$ is the kth subcarrier component of the transmitted data symbol block $\{d(n;t);n=0\sim N_c-1\}$. $H_e(k;t)$ is given as

$$H_e(k;t) = \sqrt{\Omega(k)}\sum_{m\in\mathbf{N}_{bs}} W_m(k)H_m(k;t)$$
$$= \begin{cases} \dfrac{\sum_{m\in\mathbf{N}_{mbs}}\hat{H}_m^*(k;t=0)H_m(k;t)}{\sqrt{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t=0)\right|^2}\sqrt{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t=1)\right|^2}} \Big/ \sqrt{\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}\left(\dfrac{1}{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t=1)\right|^2}\right)} & \text{MRCD(w/PA)} \\ \dfrac{\sum_{m\in\mathbf{N}_{bs}}\hat{H}_m^*(k;t=0)H_m(k;t)}{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t=0)\right|^2 + (S\cdot T/N_0)^{-1}} & \text{MMSECD(w/o PA)} \end{cases} . \quad (13)$$

After performing MRCD/MMSECD, the time-domain received signal block $\{r(n;t); n=0\sim N_c-1\}$ is obtained by applying an N_c-point IDFT to $\{R(k;t); k=0\sim N_c-1\}$. $r(n;t)$ and symbol decision are expressed as

$$r(n;t)=\frac{1}{\sqrt{N_c}}\sum_{k=0}^{N_c-1}R(k;t)\exp\left(j2\pi n\frac{k}{N_c}\right), \tag{14}$$

$$\hat{d}(n;t)=\min_{d\in\mathbf{D}}\left|r(n;t)-\sqrt{2S}\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}H_e(k;t)\right)d\right|,\ t=2\sim 13. \tag{15}$$

As seen from Eq. (15), symbol decision at BS requires the knowledge of the equivalent channel gain $H_e(k;t)$ after MRCD/MMSECD. However, $\Omega(k)$ and $H_m(k;t)$ are unknown to BS. Therefore, firstly, $\hat{H}_m(k;t=0)$ is substituted into Eq. (11) instead of $\hat{H}_m(k;t=1)$ to obtain the estimate of $\Omega(k)$. Then, by substituting the resultant $\Omega(k)$ and $\hat{H}_m(k;t=0)$ instead of $H_m(k;t)$ into Eq. (13), the equivalent channel gain estimate $\hat{H}_e(k;t)$ to be used for symbol decision at slot time t=2 is obtained, as

$$\hat{H}_e(k;t=2)=\begin{cases}1\Bigg/\sqrt{\dfrac{1}{N_c}\displaystyle\sum_{k=0}^{N_c-1}\left(\dfrac{1}{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t=0)\right|^2}\right)^2} & \text{MRCD(w/PA)}\\[3ex] \dfrac{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t=0)\right|^2}{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t=0)\right|^2+(S\cdot T/N_0)^{-1}} & \text{MMSECD(w/oPA)}\end{cases}. \tag{16}$$

Similar to OFDM downlink transmission, although $\{\hat{H}_m(k;t=0); m\in\mathbf{N}_{bs}\}$ and $\hat{H}_e(k;t=2)$ can be used during slot time t=2~13 in a low mobility environment, they may deviate from the actual ones in a high mobility environment. Therefore, we introduce the following simple 1st order and 2nd order LP of $1\times N_{bs}$ SIMO channel $\{H_m(k;t); m\in\mathbf{N}_{bs}\}$ as follows.

$$\begin{cases}\hat{H}_m(k;t=3)=\tilde{H}_m(k;t=2)\ \ :1^{st}\text{ order LP},\ t=3\\ \hat{H}_m(k;t)=\begin{cases}\tilde{H}_m(k;t-1) & :1^{st}\text{ order LP}\\ 2\tilde{H}_m(k;t-1)-\tilde{H}_m(k;t-2) & \\ :2^{nd}\text{ order LP}\end{cases},\ t=4\sim 13\end{cases}, \tag{17}$$

where $\tilde{H}_m(k;t)$ is the equivalent channel gain estimate obtained using both MMSE CE [10] and frequency-domain moving average as

$$\tilde{H}_m(k;t)=\frac{1}{Q}\sum_{l=-(Q-1)/2}^{+(Q-1)/2}\frac{R(k+l;t)\hat{D}^*(k+l;t)}{\left|\hat{D}(k+l;t)\right|^2+(E_s/N_0)^{-1}} \tag{18}$$

with $\hat{D}(k;t)$ being the kth subcarrier component of hard-decision symbol block $\{\hat{d}(n;t); n=0\sim N_c-1\}$ obtained by an N_c-point DFT. MMSE CE is introduced since the SC signal spectrum varies over the signal bandwidth. Similar to the OFDM downlink transmission, the optimum Q will be determined by computer simulation.

BS computes the MRCD/MMSECD weight $\{W_m(k); m\in\mathbf{N}_{bs}\}$ by replacing $\hat{H}_m(k;t=0)$ with $\hat{H}_m(k;t)$ in Eq. (10) and the equivalent channel gain $\hat{H}_e(k;t)$ by replacing $\hat{H}_m(k;t=0)$, $\hat{H}_m(k;t=1)$, and $H_m(k;t)$ with $\hat{H}_m(k;t)$ in Eq. (13), as follows.

$$W_m(k)=\begin{cases}\dfrac{\hat{H}_m^*(k;t)}{\sqrt{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t)\right|^2}} & \text{MRC weight}\\[3ex] \dfrac{\hat{H}_m^*(k;t)}{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t)\right|^2+(S\cdot T/N_0)^{-1}} & \text{MMSEC weight}\end{cases}, \tag{19}$$

$$\hat{H}_e(k;t)=\begin{cases}1\Bigg/\sqrt{\dfrac{1}{N_c}\displaystyle\sum_{k=0}^{N_c-1}\left(\dfrac{1}{\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t)\right|^2}\right)^2} & \text{MRCD (w/ PA)}\\[3ex] \sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t)\right|^2\Big/\sum_{m\in\mathbf{N}_{bs}}\left|\hat{H}_m(k;t)\right|^2+(S\cdot T/N_0)^{-1} & \text{MMSECD (w/o PA)}\end{cases}. \tag{20}$$

Using $\hat{H}_e(k;t)$ of Eq. (20) and the MRCD/MMSECD output $R(k;t)$ obtained by substituting $\{W_m(k); m\in\mathbf{N}_{bs}\}$ into Eq. (12), symbol decision is carried out as shown in Eq. (15).

IV. COMPUTER SIMULATION

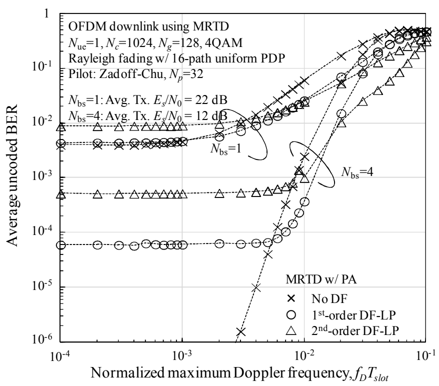
Here, a preliminary computer simulation is carried out assuming $N_{bs'}$=N_{bs} (=1~4) (i.e., no BS antenna selection is considered) and N_{ue}=1 to examine the uncoded BER performances of OFDM downlink transmission using $N_{bs}\times1$ MRTD and SC uplink transmission using $1\times N_{bs}$ MMSECD. The computer simulation parameters are summarized in Table 1. The Zadoff-Chu sequence [11] is employed as the pilot. 4QAM data modulation is assumed. MMSE CE for SC uplink requires the knowledge of the received E_s/N_0 as seen from Eq. (18), however, the perfect knowledge of received E_s/N_0 is assumed.

TABLE I. COMPUTER SIMULATION PARAMETERS

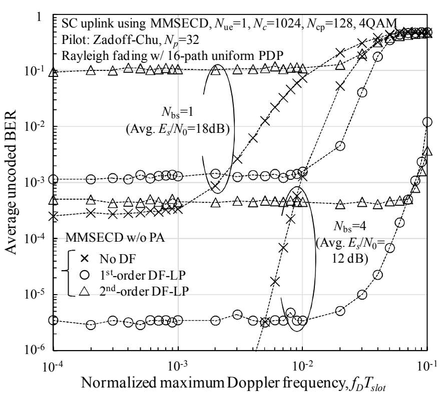
Subframe structure (14 slots)	No. of subcarriers	N_c=1024
	CP length	N_{cp}=128 samples
	Uplink/downlink pilots	Zadoff-Chu
	Data (12 slots)	4QAM
Fading channel	Type of fading	L=16-path block Rayleigh assuming Jakes model
	Power delay profile	Uniform
	Maximum delay time	16 samples
$N_{bs}\times1$ MRTD (downlink)/ $1\times N_{bs}$ MRCD/MMSECD (uplink)	No. of BS antennas	N_{bs}=1~4
	No. of UE antennas	N_{ue}=1

Firstly, we examined the impact of the moving average window size Q on the achievable BER performance and found that Q=11 is near the optimum. Then, we examined the uncoded average BER performance in a high mobility environment. We observed that the use of PA improves the BER performance when the dominant cause of decision error is the AWGN (i.e., in a relatively low E_s/N_0 region) while almost no impact is obtained when the dominant cause of decision error is the random phase variation due to the time-varying channel (i.e., in a high E_s/N_0 region). The BER performance is plotted as a function of the normalized maximum Doppler frequency $f_D T_{slot}$ in Fig. 4, where f_D and T_{slot} denote the Rayleigh fading maximum Doppler frequency and the time slot length, respectively (in the final paper, the BER performance of SC uplink using MRCD (with PA) will be included and discussed). It can be seen from the figure that in a high mobility environment such as $f_D T_{slot} > 0.01$, DF-LP together with pilot-aided CE provides better BER performance than the use of pilot-aided CE only (i.e., no DF). When N_{bs}=4, the use of the 1st order DF-LP increases the allowable maximum $f_D T_{slot}$ (required for keeping BER < 10^{-3}) from 0.008 to 0.01 for OFDM downlink using MRTD (with PA) and from 0.01 to 0.07 for SC uplink using MMSECD (without PA). Assuming the subcarrier spacing of 60 kHz and the carrier frequency of 28 GHz, the traveling speeds of up to 21 km/h and 144 km/h may be allowed for OFDM downlink and SC uplink, respectively.

It should be noted that the BER performance is degraded due to decision error propagation in a low E_s/N_0 region. Therefore, an adaptive switching between pilot-aided CE only and pilot-aided CE + DF-LP is necessary. This is left as our future study.



(a) OFDM downlink using MRTD (w/ PA)



(b) SC uplink using MMSECD (w/o PA)

Fig. 4. Uncoded BER versus $f_D T$.

V. CONCLUSION

In this paper, decision-feedback prediction channel estimation (DF-LP CE) for MIMO cooperative transmission was proposed. The effect of the proposed DF-LP CE on OFDM downlink transmission using MRTD and SC uplink transmission using MMSECD was confirmed by computer simulation. Instead of LP, the use of recursive least square algorithm (RLS) [12] can improve the tracking ability against time-varying channel while suppressing the noise enhancement. This is our interesting future study. In the computer simulation, the number of BS antennas before antenna selection and that of UE antennas were assumed as $N_{bs'}=N_{bs}$ (=1~4) and N_{ue}=1, respectively. Further performance improvement can be obtained due to increased diversity gain by increasing $N_{bs'}$ and N_{ue} for the given N_{bs}. Also assumed in the computer simulation was the block Rayleigh fading. However, this assumption may not hold in a very high mobility environment such as $f_D T_{slot}$=0.1. A further careful examination is necessary. They are left as our important future studies.

ACKNOWLEDGMENT

This paper includes a part of results of “The research and development project for realization of the fifth-generation mobile communications system” commissioned to Tohoku University by The Ministry of Internal Affairs and Communications (MIC), Japan.

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