An effective channel estimation scheme for bi-directional two-timeslot OFDM relay transmission using analog network coding

Yuta Koshimizu and Eiji Okamoto
Department of Computer Science and Engineering, Graduate school of Engineering, Nagoya Institute of Technology
Gokiso-cho, Showa-ku, Nagoya-shi, 466-8555, Japan

Abstract*—Recently, relay transmission schemes have been much studied according to increasing demands for ad-hoc wireless communication. In bi-directional relay transmission, the transmission efficiency has been decreased due to the waiting time. To tackle this problem, network coding (NC) has been proposed. The orthogonal frequency division multiplexing (OFDM)-based relay transmission applying analog network coding (ANC) can reduce the number of timeslots per one sequence compared to that without NC and it increases the transmission capacity. However, in the bi-directional relay transmission with ANC, the relay node receives the signals of each transmitting nodes simultaneously and the signal detection must be conducted. To do that, the channel state information of each nodes is needed, and the rate-efficient and accurate channel estimation should be applied for the bi-directional relay, which is not much considered so far. Therefore, in this paper, we propose a new channel estimation scheme for bi-directional two-timeslot OFDM relay transmission where the transmitters send sparse Walsh-code pilots and the receiver detects and estimates each channel information by the correlation processing. The proposed scheme can estimate the channels by less pilot symbols sparsely allocated in OFDM frame. In addition, an accurate interpolation scheme with two-dimensional fast Fourier transform (FFT) is utilized. The performances of the proposed scheme are evaluated by computer simulations.

I. INTRODUCTION

Recently, according to the increasing demands for ad-hoc network communication, relay transmission has attracted much attention and many schemes have been proposed. One of those schemes is a network coding (NC) [1-3]. In a bi-directional relay transmission, a waiting time is needed for two-way transmission, which results in a throughput degradation. NC has been proposed to solve this problem. In the conventional three-node bi-directional relay transmission, four timeslots are needed to finish one sequence, while it can be decreased to two timeslots by applying an analog network coding (ANC), which greatly improves the throughput performance [3]. However, in the bi-directional relay transmission with ANC, Timeslot 1 becomes a multiple input single output (MISO) transmission where the relay node receives the mixed signals of each transmit node. Thus, the signal detection and separation are needed not only for data transmission but also for channel estimation. In the conventional bi-directional transmission schemes, the channel is often assumed as known perfectly at the receiver. In [4], the channel estimation scheme for bi-directional relay transmission has been proposed in which an accurate estimation of each channels are achieved by orthogonal division of each transmission link with phase-shifted pilots. However, to obtain the orthogonality this scheme needs the block pilot frame as shown in Fig. 1(a). This means that whole one orthogonal frequency division multiplexing (OFDM) frame must be used for pilots, resulting in decreased throughput performance.

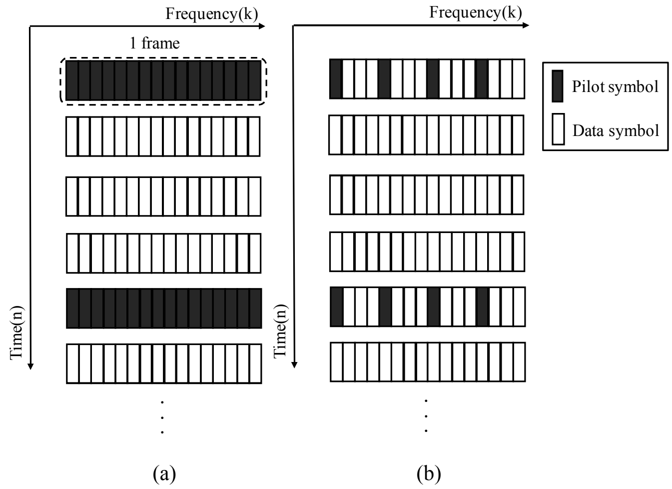

Fig. 1. Pilot frame structure: (a) block type, (b) sparse type.

Therefore, in this paper we propose a new channel estimation scheme for bi-directional two-timeslot OFDM relay transmission with a reduced redundancy as shown in Fig. 1(b). In the proposed scheme, Walsh codes are adopted as the pilot [5-6] and after channel decoupling by correlation processing, an accurate channel estimation is conducted. The interpolation between sparse pilots are carried out by two-dimensional fast Fourier transform (FFT)-based scheme [7]. This two-dimensional FFT-based interpolation just needs domain transformation and zero insertion, and performs well. We evaluate the performances of the proposed scheme by computer simulations.

This paper is organized as follows. Bi-directional network model is described in Section II. The proposed scheme and the

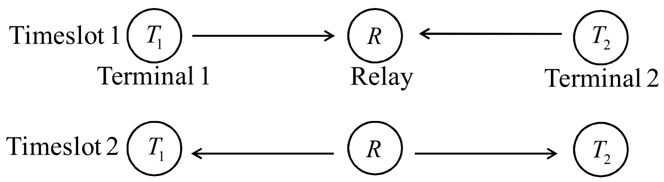

Fig. 2. Bi-directional relay network model.

two-dimensional FFT-based interpolation scheme are presented in Section III and IV, respectively. Numerical results are shown in Section V and the conclusions are drawn in Section VI.

II. NETWORK MODEL

Fig. 2 shows the network model considered in this paper. Node T_1 and T_2 transmit data each other via relay node R using OFDM frame. It is assumed that each terminal equips one antenna. At Timeslot 1, node T_1 and T_2 send data to relay R, which can be regarded as MISO transmission. As described in Section I, the received signal at R becomes MISO mixed signal where each transmitted signal should be detected and separated. Also in the channel estimation, this signal processing must be done. At Timeslot 2, an amplify-and-forward (AF) relaying from R to T_1 and T_2 is conducted. Different from Timeslot 1, Timeslot 2 relaying becomes two parallel single input single output (SISO) transmission without co-channel interference. Then, the channel estimation is relatively easy to conduct compared to Timeslot 1. In the following, operations at each timeslot are described.

At Timeslot 1, T_1 and T_2 send data to R at the same time. Let $S_i(k)$ ($k = 0, \cdots, N-1$) as the OFDM transmit symbol of node T_i ($i = 1, 2$) after first modulation at subcarrier k, where N is the number of subcarriers. The N-symbol time sequence of OFDM is composed by inverse fast Fourier transform (IFFT) of $S_i(k)$ at each T_i. After the addition of N_{GI}-symbol guard interval (GI), the composed OFDM signal is transmitted. The received symbol at relay node R after removal of GI and FFT is described by

$$R_1(k) = \sum_{i=1}^{2} H_{1,i}(k) S_i(k) + N_1(k) \quad (k = 0, \cdots, N-1) \tag{1}$$

where $H_{1,i}(k)$ is the channel coefficient between T_i and R at Timeslot 1, and $N_1(k)$ is zero-mean Gaussian noise. At Timeslot 2, relay node R re-transmits the signal. The forwarding signal $\hat{R}_1(k)$ is generated by

$$\hat{R}_1(k) = \frac{R_1(k)}{\sqrt{\left|H_{1,1}(k)\right|^2 + \left|H_{1,2}(k)\right|^2 + \sigma_1^2}} \quad (k = 0, \cdots, N-1) \tag{2}$$

where σ_1^2 is noise variance at relay node R and the transmit power is normalized to 1. This $\hat{R}_1(k)$ is transmitted to each node T_i after IFFT and GI insertion. The received symbol $R_{2,i}(k)$ at T_i after GI removal and FFT is described by

$$R_{2,i}(k) = H_{2,i}(k) \hat{R}_1(k) + N_{2,i}(k) \quad (k = 0, \cdots, N-1) \tag{3}$$

where $H_{2,i}(k)$ and $N_{2,i}(k)$ are the channel coefficient between R and T_i and zero-mean Gaussian noise at T_i, respectively, at Timeslot 2. To obtain the data of other node, T_i subtracts his own transmit data from $R_{2,i}(k)$ first as follows.

$$\hat{R}_{2,i}(k) = \sqrt{\left|H_{1,1}(k)\right|^2 + \left|H_{1,2}(k)\right|^2 + \sigma_1^2}\, R_{2,i}(k) - H_{1,i}(k) H_{2,i}(k) S_i(k) \quad (k = 0, \cdots, N-1) \tag{4}$$

Then, the received data $\tilde{R}_{2,i}(k)$ is obtained by the equalization of $\hat{R}_{2,i}(k)$ as follows.

$$\tilde{R}_{2,i}(k) = \frac{\hat{R}_{2,i}(k)}{H_{1,j}(k) H_{2,i}(k)} \quad (k = 0, \cdots, N-1) \tag{5}$$

Here, (5) is the zero-forcing (ZF) equalization and j is calculated by

$$j = (3 - i) = \begin{cases} 1 & \text{if } i = 2 \\ 2 & \text{if } i = 1 \end{cases} \tag{6}$$

After the execution of (5) at each node T_i, the bi-directional relay transmission is terminated.

As described above, it is needed that relay node R estimates the channel $H_{1,i}(k)$ at Timeslot 1 and T_i estimates the channel $H_{2,i}(k)$ at Timeslot 2.

III. PROPOSED CHANNEL ESTIMATION SCHEME

In this section we describe the proposed channel estimation scheme at Timeslot 1 which is MISO transmission requiring channel detection and separation. Timeslot 2 is the SISO transmission and the conventional link-by-link channel estimation can be applied [8]. The characteristics of the proposed scheme at Timeslot 1 are

(1) Walsh code is used as the pilot symbols at transmitter T_i.
(2) Each channel is extracted by the correlation processing.
(3) Sparse pilot symbol allocation is available.

We use 2-by-2 Walsh code as shown below as the pilots.

$$\boldsymbol{W}_2 = \begin{bmatrix} W_{11} & W_{12} \\ W_{21} & W_{22} \end{bmatrix} = \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} \tag{7}$$

The column 1 is allocated to node T_1 and column 2 is allocated to node T_2. The transmit frame structure of T_1 and T_2 consisting of multiple OFDM frame are shown in Fig. 3. This Walsh-code pilots of Fig. 3 enable the orthogonal division of T_1 and T_2 channel coefficients after correlation processing at relay node R. The channel coefficients are obtained as follows. In Fig. 3, the pilot symbols $P_i(k,n)$ of node T_i are described by

$$\begin{gathered} P_i(k,n) = W_{xi},\ x = \mathrm{mod}\left(n/N_t, 2\right) + 1 \\ (k = 0, N_f, 2N_f, \cdots), (n = 0, N_t, 2N_t, \cdots) \end{gathered} \tag{8}$$

where k, n are the frequency index and time index, respectively, mod(a,b) are the remainder of a divided by b, and N_t and N_f are the periods of pilot symbol in the time and frequency directions, respectively. Here, the OFDM frame

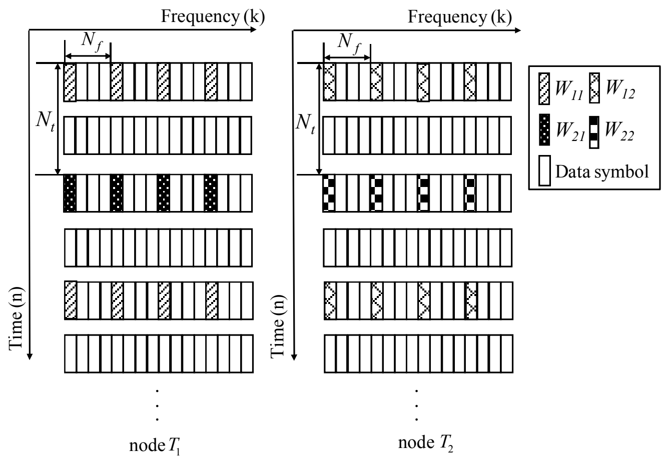

Fig. 3. Frame structure at Timeslot 1.

with pilots is considered. The received symbol $R(k,n)$ at relay node R is obtained by

$$R(k,n)=\sum_{i=1}^{2}H_i(k,n)P_i(k,n)+N(k,n)$$
$$(k=0,N_f,2N_f,\cdots),(n=0,N_t,2N_t,\cdots) \tag{9}$$

where $H_i(k,n)$ is the channel coefficient between T_i and R at the index of (k,n) and $N(k,n)$ is the Gaussian noise. As (9), the pilot symbols are received as overlapped signals from T_1 and T_2 . To obtain the separated channel $\hat{H}_i(k,n)$ from $R(k,n)$, the correlation processing is executed based on Walsh-code orthogonality in relay node, which is given by

$$\hat{H}_i(k,n)=\frac{1}{2}\left(\begin{array}{l}\frac{1}{2}\sum_{l=-1}^{0}R(k,n+lN_t)P_i(k,n+lN_t)\\+\frac{1}{2}\sum_{l=0}^{1}R(k,n+lN_t)P_i(k,n+lN_t)\end{array}\right) \tag{10}$$

Here, we confirm the extraction of $\hat{H}_1(k,n)$ from $R(k,n)$ using (10) when $\bmod\left(n/N_t,2\right)=0$. Eq. (10) is expanded as follows.

$$\begin{aligned}\hat{H}_1(k,n)&=\frac{1}{4}\left(H_1(k,n-N_t)+2H_1(k,n)+H_1(k,n+N_t)\right)\\&+\frac{1}{4}\left(-H_2(k,n-N_t)+2H_2(k,n)-H_2(k,n+N_t)\right)\\&+\frac{1}{4}\left(N(k,n-N_t)+2N(k,n)+N(k,n+N_t)\right)\\&=H_1(k,n)+\Delta_1+\Delta_2+\Delta_3\end{aligned} \tag{11}$$

where

$$\begin{cases}\Delta_1=\frac{1}{4}\left(H_1(k,n-N_t)-2H_1(k,n)+H_1(k,n+N_t)\right)\\\Delta_2=\frac{1}{4}\left(-H_2(k,n-N_t)+2H_2(k,n)-H_2(k,n+N_t)\right)\\\Delta_3=\frac{1}{4}\left(N(k,n-N_t)+2N(k,n)+N(k,n+N_t)\right)\end{cases} \tag{12}$$

The objective coefficient is $H_1(k,n)$, Δ_1 and Δ_2 are the interference due to the time variance of channel, and Δ_3 is the Gaussian noise. These interferences Δ_1 and Δ_2 are mitigated

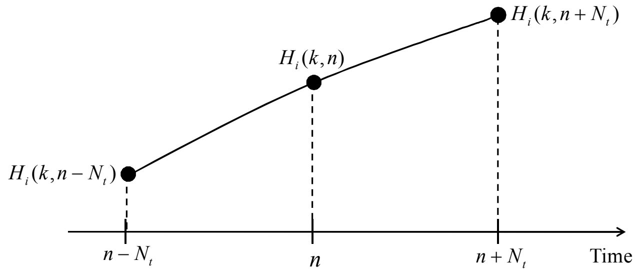

Fig. 4. Channel fluctuation.

when the channel $H_i(k,n)$ is static or proportionally changes at the interval of $\left[n-N_t,n+N_t\right]$ as shown in Fig. 4, and they become $\Delta_1\approx\Delta_2\approx 0$ that means the accurate channel estimation is obtained. Furthermore, since three independent terms are averaged in Δ_3, the Gaussian noise is also mitigated.

IV. Channel Interpolation Using Two-dimensional FFT

As described in Section III, the discrete channel coefficients at pilot symbol indices are obtained. To obtain the all channel coefficients of (k,n), interpolation should be applied. We adopt the channel interpolation scheme using two-dimensional FFT (2D-FFT) [9-10]. Here the 2D-FFT scheme is briefly reviewed and the details are written in [7].

First, the discrete channels of $\hat{H}_i(k,n)$ are collected to $\bar{H}_i(k,n)$ as follows

$$\bar{H}_i(k,n)=\hat{H}_i(N_f k,N_t n)$$
$$(k=0,1,\cdots,N_{pf}-1),\ (n=0,1,\cdots,2N_{pt}-1) \tag{13}$$

where N_{pf} and N_{pt} are the number of reference pilot symbols in the frequency and time direction, respectively. When these parameters are large, more exact interpolation is obtained with increased calculation complexity and delay. The interpolation is carried out with applying a window function to raise the accuracy. The time-domain window function $W(n)$ is multiplied to $\bar{H}_i(k,n)$ as

$$G_i(k,n)=\bar{H}_i(k,n)W(n)\ \ (n=0,1,\cdots,2N_{pt}-1) \tag{14}$$

As the window function, the Blackman function is used in this paper as follows

$$W(n)=0.423-0.498\cos\left(\frac{\pi n}{N_{pt}}\right)+0.0792\cos\left(\frac{2\pi n}{N_{pt}}\right)$$
$$(n=0,1,\cdots,2N_{pt}-1) \tag{15}$$

Next, $G_i(k,n)$ is transformed to transpose domain [10] by 2D-FFT which is described by

$$g_i(m,l)=\sum_{k=0}^{N_{pf}-1}\left\{\sum_{n=0}^{2N_{pt}-1}G_i(k,n)\exp\left(\frac{-j\pi l(n-N_{pt})}{N_{pt}}\right)\right\}\cdot\exp\left(\frac{-j2\pi m(k-N_{pf}/2)}{N_{pf}}\right) \tag{16}$$
$$(m=-\frac{N_{pf}}{2},-\frac{N_{pf}}{2}+1,\cdots,\frac{N_{pf}}{2}-1),\ (l=-N_{pt},-N_{pt}+1,\cdots,N_{pt}-1)$$

where $g_i(m,l)$ is the transformed function of $G_i(k,n)$ by 2D-FFT and m , l are the symbol indices after transformed corresponding to k , n . Then, the interpolated function $g_i'(m',l')$ is generated by insertion of zero sequences outside of $g_i(m,l)$ as (17). This operation corresponds to the interpolation between pilot symbol in (k,n)-domain.

$$g_i'(m',l') = \begin{cases} N_{pf} N_{pt} g_i(m',l') & \left[-\frac{N_{pf}}{2} \le m' \le \frac{N_{pf}}{2} - 1, -N_{pt} \le l' \le N_{pt} - 1 \right] \\ 0 & \text{others} \end{cases}$$
$$(m' = -\frac{N_f N_{pf}}{2}, -\frac{N_f N_{pf}}{2} + 1, \cdots, \frac{N_f N_{pf}}{2} - 1),$$
$$(l' = -N_t N_{pt}, -N_t N_{pt} + 1, \cdots, N_t N_{pt} - 1) \qquad (17)$$

Here, m' and l' are the symbol indices after interpolation corresponding to m and l , respectively. The interpolated channel $G_i'(k',n')$ can be obtained by transforming the transpose-domain function $g_i'(m',l')$ into frequency-time domain by 2D-IFFT, which is described by

$$G_i'(k',n') = \frac{1}{N_f N_{pf}} \sum_{m'=-(N_f N_{pf})/2}^{(N_f N_{pf})/2} \left\{ \frac{1}{2N_t N_{pt}} \sum_{l'=-N_t N_{pt}}^{N_t N_{pt}-1} g_i'(m',l') \exp\left(\frac{j\pi l'(n' - N_t N_{pt})}{N_t N_{pt}} \right) \right\} \cdot \exp\left(\frac{-j2\pi m'(k' - N_f N_{pf}/2)}{N_f N_{pf}} \right)$$
$$(k' = 0, 1, \cdots, N_f N_{pf} - 1), (n' = 0, 1, \cdots, 2N_t N_{pt} - 1) \qquad (18)$$

where k' and n' are the symbol indices after transformed corresponding to frequency and time directions as well as k and n .

Finally, after dividing time-domain window function and eliminating degraded part of both ends due to alias effect, the estimated channel coefficients $\hat{H}_i(k',n')$ are obtained, which is given by

$$\hat{H}_i(k', n' + N_t N_{pt}/2) = \frac{G_i'(k', n' + N_t N_{pt}/2)}{W'(n' + N_t N_{pt}/2)}$$
$$(k' = 0,1,\cdots,N_f N_{pf} - 1), (n' = 0,1,\cdots,N_t N_{pt} - 1)$$
$$W'(n) = 0.423 - 0.498\cos\left(\frac{\pi n}{N_t N_{pt}}\right) + 0.0792\cos\left(\frac{2\pi n}{N_t N_{pt}}\right) \qquad (19)$$
$$(n = 0,1,\cdots,2N_t N_{pt} - 1)$$

Note that N_t , N_f , N_{pt} , and N_{pf} should be the power of two to utilize the 2D-FFT and that a noise reduction technique using zero padding which adopted in [4] is utilized for a fair comparison.

V. NUMERICAL RESULTS

We evaluate performances of the proposed scheme by computer simulations. Tab. I shows the simulation conditions. The frame structure at Timeslot 1 is shown in Fig. 3 as described in Section III, and the frame structure at Timeslot 2 is shown in Fig. 5. It is assumed that the estimated channels at Timeslot 1 on relay node R are perfectly fed back to each node by ideal feedback channel [4]. This channel information is used in subtraction of own information on NC in (4) and in equalization in (5). In practice, if the channel is quasi-static, this assumption is satisfied by using some sparsely periodic feedback signal from R to T_i , or if the channel is dynamically changed, some quantized channel index should be inserted in data every time. However, these implementation is not considered in this paper because of a space limit.

TABLE I
SIMULATION CONDITIONS

Transmission scheme	OFDM
Modulation	QPSK
Number of subcarriers	64
Number of GI	10
Timeslot 1 pilot symbol period (N_t, N_f)	(1,4), (2,4), (4,4)
Timeslot 2 pilot symbol period (N_t', N_f')	(4,4)
(N_{pt}, N_{pf}) in (13)	(4,16)
Channel	i.i.d 9-path 3dB decayed Rayleigh fading
Relay scheme	Amplify-and-forward
Equalization	Zero-forcing weight

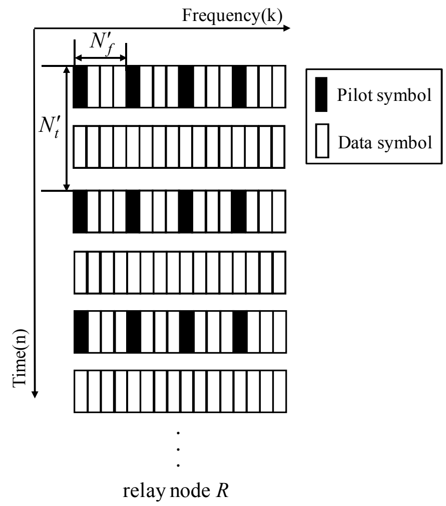

Fig. 5. Frame structure at Timeslot 2.

Fig. 6 shows the bit error rate (BER) performance versus E_b / N_0 when a normalized Doppler frequency is $f_d T_s = 10^{-4}$. The average BER of both end terminals T_1 and T_2 are plotted. As the conventional schemes, we compare the performance with the second-degree Gaussian interpolation scheme [11] with channel estimation of [4]. Note that in the conventional estimation scheme of [4], the ratio of pilot symbols becomes $1/2$ at $N_t = 2$, $1/4$ at $N_t = 4$, while in the proposed scheme it becomes $1/4$, $1/8$, and $1/16$ at $(N_t, N_f) = (1,4)$, $(2,4)$, and $(4,4)$, respectively. In Fig. 6, "Ideal" is the BER with perfect channel estimation. It is confirmed that the proposed scheme has near performance to the conventional scheme in spite of sparse pilot symbols, and even if the ratio of pilot symbol is decreased in the time direction, BER is not degraded. This means that the channel fluctuation is almost linear

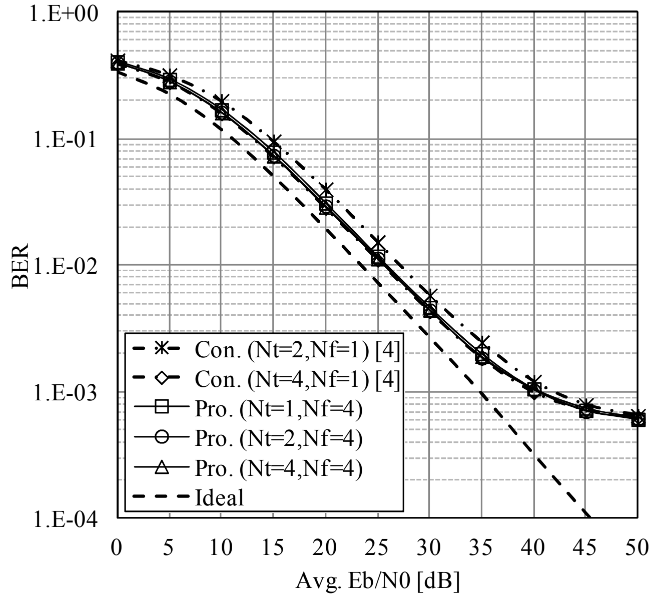

Fig. 6. BER performances versus E_b / N_0 at $f_dT_s = 10^{-4}$.

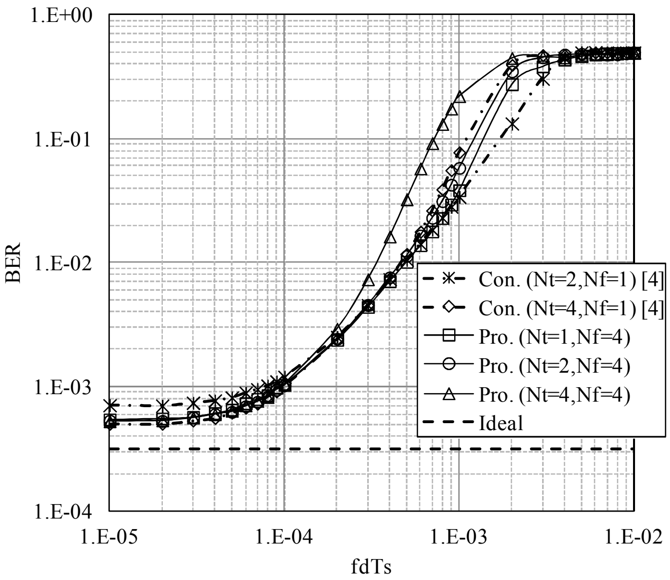

Fig. 7. BER performances versus f_dT_s at $E_b / N_0 = 40$ dB .

during $[n-N_t, n+N_t]$ at $f_dT_s = 10^{-4}$. However, there is an error floor at BER=6×10^{-4} in the proposed scheme due to a severe frequency selectivity of the channel whose condition is listed in Tab. I. In this channel, the received power of some subcarriers become fallen and the channel equalization accuracy decreases in those subcarriers, resulting in the error floor.

Fig. 7 shows the BER versus f_dT_s at $E_b / N_0 = 40$dB . Below $f_dT_s < 10^{-4}$ the BER of proposed scheme is lower than 10^{-3} and almost stable regardless of pilot insertion ratio. Compared with the conventional scheme, the proposed scheme achieves similar performance with less pilot symbols. When the fading becomes fast such as $f_dT_s \geq 10^{-4}$, the BER of the proposed scheme is degraded and that of sparse pilot setting is worse. This is because the interval $[n-N_t, n+N_t]$ becomes long with a large N_t and the linear condition of fading in this interval is not satisfied. In this case, the interference coefficients Δ_1 and Δ_2 become larger with the large period N_t , and the performance is quickly degraded.

VI. CONCLUSION

We proposed a new channel estimation scheme for bi-directional OFDM relay transmission with ANC. Applying the Walsh code as pilot symbols in the transmitter, the relay node can detect and separate the channels of two terminals at the same time by conducting correlation signal processing. Conventional schemes use a long pilot sequences in the bi-directional relay transmission, while we achieve a less redundant pilot frame. In addition, from numerical results it has been confirmed that the better performance is obtained in the proposed scheme, especially, at $f_dT_s \leq 10^{-4}$, BER is not degraded even if the interval N_t is enlarged.

In future studies, an enhanced scheme in which the interference coefficients Δ_1 and Δ_2 are mitigated will be considered.

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