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Pilot-Assisted Decision Feedback Channel Estimation for STTD in OFDM Mobile Radio

Koichi ISHIHARA[†a)], **Kazuaki TAKEDA**[†], *Student Members, and* **Fumiyuki ADACHI**[†], *Member*

SUMMARY In this paper, we propose pilot-assisted decision feedback channel estimation (PA-DFCE) for space-time coded transmit diversity (STTD) in orthogonal frequency division multiplexing (OFDM). Two transmit channels are simultaneously estimated by transmitting the STTD encoded pilot. To improve the tracking ability of the channel estimation against fast fading, decision feedback is also used in addition to pilot. For noise reduction and preventing the error propagation, windowing of the estimated channel impulse response in the time-delay domain is applied. The average bit error rate (BER) performance of OFDM with STTD is evaluated by computer simulation. It is found that the use of PA-DFCE can achieve a degradation in the required E_b/N_0 from ideal CE of as small as 0.6 dB for an average $BER = 10^{-3}$ and requires about 2.4 dB less E_b/N_0 compared to differential STTD that requires no CE.

key words: *pilot, decision feedback, channel estimation, OFDM, STTD*

1. Introduction

In the next generation mobile communication systems, very high-speed and high-quality data transmission will be required. However, for very high-speed data transmission, the transmission performance is severely degraded due to severe inter-symbol interference (ISI) resulting from frequency-selective fading. Recently, orthogonal frequency division multiplexing (OFDM) [1], [2] has been attracting much attention. OFDM uses a number of lower rate subcarriers to prevent the ISI and to efficiently utilize the limited frequency band. However, its transmission performance is subjected to frequency-nonselective fading. To improve the performance in frequency-nonselective fading, antenna diversity technique is attractive [3]. Recently, transmit diversity is attracting much attention because it can reduce the number of diversity antennas at a mobile receiver and can alleviate the complexly problem of mobile terminal [4]. Especially, Alamouti's space-time coded transmit diversity (STTD) [5] has been gaining a lot of attention for downlink (base-to-mobile) transmissions [6], [7].

For STTD decoding, the knowledge of channel transfer functions of the two transmit antennas is required. Pilot-assisted channel estimation (CE) can be applied, where CE is carried out using the periodically transmitted known pilot symbols [8], [9]. In this paper, we propose to transmit STTD encoded pilot block for simultaneous estimation of two transmit channels. As the pilot transmission rate increases, more accurate channel estimation is possible, but the power loss due to pilot insertion increases. To avoid this power loss problem, we combine STTD encoded pilot-assisted CE with decision feedback (DF), called pilot-assisted decision feedback CE (PA-DFCE). If STTD encoded pilot block is not available, the previous data decision is feedback and is used as the pilot. To further reduce the noise effect and decision feedback error in DFCE, we apply the time-delay domain windowing which was proposed for CE in OFDM [10], [11].

The remainder of this paper is organized as follows. The transmission system model of OFDM with STTD is presented in Sect. 2. The proposed PA-DFCE is described in Sect. 3. In Sect. 4, the computer simulation results for the BER performance of OFDM with STTD using the proposed PA-DFCE are presented and compared with differential STTD [12] which requires no CE. The paper is concluded in Sect. 5.

2. Transmission System Model of OFDM with STTD

Figure 1 shows the transmitter/receiver structure. We consider OFDM having K orthogonal subcarriers. Throughout the paper, discrete-time representation for the OFDM signal is used. Without loss of generality, we consider the transmission of $2K$ data modulated symbols during two consecutive OFDM signaling intervals, i.e., $0 \le t < 2(K+N_g)$, where N_g is the guard interval (GI).

2.1 STTD Encoding

At the transmitter, the data modulated symbol sequence is divided into a sequence of blocks, each with K symbols. There are two STTD encoding methods; one in the frequency-domain and the other in the time-domain.

Figure 1(a) shows the transmitter structure for STTD encoding in the frequency-domain. The nth even block $\{d_{e,n}(k);\ k = 0 \sim K-1\}$ and odd block $\{d_{o,n}(k);\ k = 0 \sim K-1\}$ are STTD encoded to produce two block sequences. The nth even and odd blocks to be transmitted from the 0th antenna are $\{d_{e,n}(k)\}$ and $\{-d^*_{o,n}(k)\}$, respectively, and those to be transmitted from the 1st antenna are $\{d_{o,n}(k)\}$ and $\{d^*_{e,n}(k)\}$, respectively. The K-point inverse fast Fourier transform (IFFT) is applied to generate the STTD encoded OFDM signals to be transmitted from two antennas. After insertion of an N_g-sample GI, the STTD encoded OFDM signals are transmitted simultaneously from two an-

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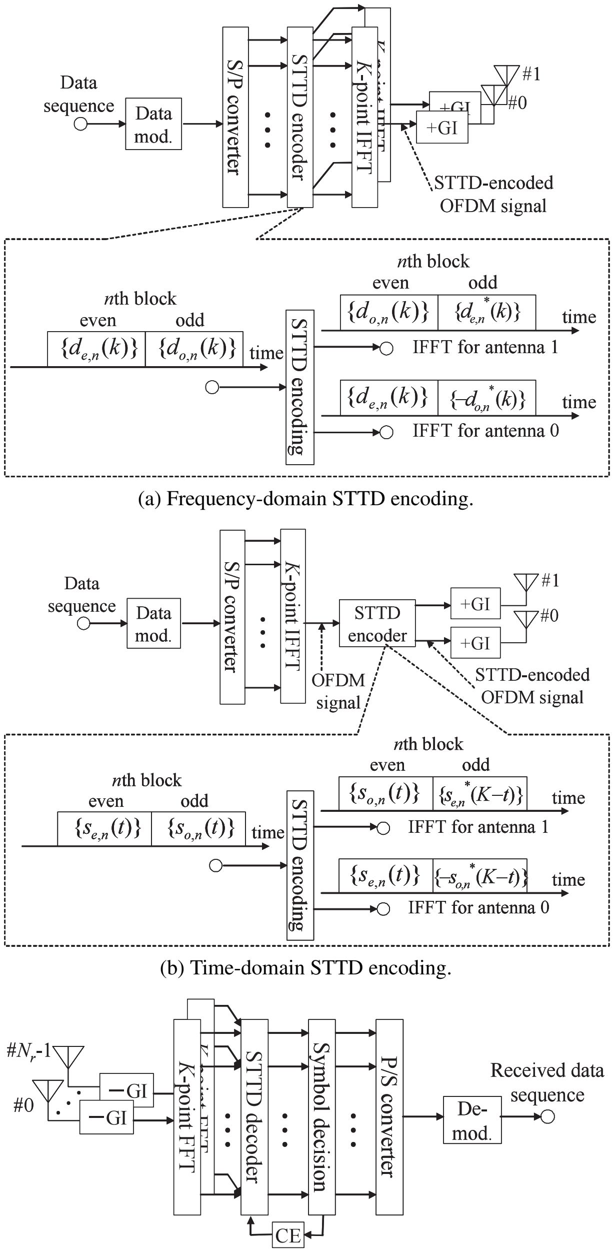
[†]The authors are with the Department of Electrical and Communication Engineering, Graduate School of Engineering, Tohoku University, Sendai-shi, 980-8579 Japan.
a) E-mail: ishihara@mobile.ecei.tohoku.ac.jp



(a) Frequency-domain STTD encoding.

(b) Time-domain STTD encoding.

(c) Receiver.

Fig. 1 Transmitter/receiver structure for OFDM with STTD.

tennas.

From Appendix, we can show that the nth even and odd blocks of STTD encoded OFDM signal to be transmitted from the 0th antenna are $\{s_{e,n}(t);\ t = 0 \sim K-1\}$ and $\{-s^*_{o,n}(K-t);\ t = 0 \sim K-1\}$, respectively, and those to be transmitted from the 1st antenna are $\{s_{o,n}(t);\ t = 0 \sim K-1\}$ and $\{s^*_{e,n}(K-t);\ t = 0 \sim K-1\}$, respectively, where

$$\begin{cases} s_{e,n}(t) = \sqrt{S}\displaystyle\sum_{k=0}^{K-1} d_{e,n}(k)\exp(j2\pi kt/K) \\ s_{o,n}(t) = \sqrt{S}\displaystyle\sum_{k=0}^{K-1} d_{o,n}(k)\exp(j2\pi kt/K) \end{cases} \tag{1}$$

with S being the total transmit power per subcarrier. This suggests that STTD encoding in the time-domain can be implemented as in [13]–[15] that does not require subcarrier-by-subcarrier STTD encoding. K-point IFFT is first applied to generate the nth even and odd blocks of the OFDM signal, $\{s_{e,n}(t);\ t = 0 \sim K-1\}$ and $\{s_{o,n}(t);\ t = 0 \sim K-1\}$, given by Eq. (1). They are STTD encoded in the time-domain as shown in Fig. 1(b).

2.2 Received Signals and STTD Decoding

The STTD encoded OFDM signals transmitted from two transmit antennas are received by N_r receive antennas at the receiver (see Fig. 1(c)). The propagation channel is assumed to be a sample-spaced L-path frequency-selective channel. The lth path gain and time delay of the channel between the 0th (or 1st) transmit antenna and the mth receive antenna are denoted as $\xi^{(m)}_{0,n,l}$ (or $\xi^{(m)}_{1,n,l}$) and τ_l, respectively. The nth even and odd blocks of the received signal on the mth antenna are denoted as $\left\{r^{(m)}_{e,n}(t)\right\}$ and $\left\{r^{(m)}_{o,n}(t)\right\}$, respectively. They are expressed using the equivalent low-pass representation as

$$\begin{cases} r^{(m)}_{e,n}(t) = \displaystyle\sum_{l=0}^{L-1}\left\{\xi^{(m)}_{0,n,l}s_{e,n}(t-\tau_l) + \xi^{(m)}_{1,n,l}s_{o,n}(t-\tau_l)\right\} \\ \qquad\qquad + \eta^{(m)}_{e,n}(t) \\ r^{(m)}_{o,n}(t) = \displaystyle\sum_{l=0}^{L-1}\left\{\xi^{(m)}_{1,n,l}s^*_{e,n}(K-t+\tau_l) - \xi^{(m)}_{0,n,l}s^*_{o,n}(K-t+\tau_l)\right\} \\ \qquad\qquad + \eta^{(m)}_{o,n}(t) \end{cases} \tag{2}$$

for $t = -N_g \sim K-1$, where $\{\eta^{(m)}_{e,n}(t)\}$ and $\{\eta^{(m)}_{o,n}(t)\}$ represent the independent additive white Gaussian noise (AWGN) processes having zero mean and variance $2N_0/T_c$ with N_0 representing the single sided power spectrum density and T_c representing FFT sampling period. After removal of GI, K-point FFT is applied to decompose the received signal into K subcarrier components. The nth even and odd blocks of the subcarrier components, $\{R^{(m)}_{e,n}(k);\ k = 0 \sim K-1\}$ and $\{R^{(m)}_{o,n}(k);\ k = 0 \sim K-1\}$, are given by

$$\begin{cases} R^{(m)}_{e,n}(k) = \dfrac{1}{K}\displaystyle\sum_{t=0}^{K-1} r^{(m)}_{e,n}(t)\exp(-2\pi tk/K) \\ \qquad = \sqrt{S}\left\{H^{(m)}_{0,n}(k)d_{e,n}(k) + H^{(m)}_{1,n}(k)d_{o,n}(k)\right\} + \Pi^{(m)}_{e,n}(k) \\ R^{(m)}_{o,n}(k) = \dfrac{1}{K}\displaystyle\sum_{t=0}^{K-1} r^{(m)}_{o,n}(t)\exp(-2\pi tk/K) \\ \qquad = \sqrt{S}\left\{H^{(m)}_{1,n}(k)d^*_{e,n}(k) - H^{(m)}_{0,n}(k)d^*_{o,n}(k)\right\} + \Pi^{(m)}_{o,n}(k). \end{cases} \tag{3}$$

where $H^{(m)}_{0(\text{or }1),n}(k)$ and $\Pi^{(m)}_{e(\text{or }o),n}(k)$ respectively represent the channel gain and noise due to AWGN at the kth subcarrier and are given by

$$\begin{cases} H^{(m)}_{0(\text{or }1),n}(k) = \sum_{l=0}^{L-1} \xi^{(m)}_{0(\text{or }1),n,l} \exp(-j2\pi\tau_l k/K) \\ \Pi^{(m)}_{e(\text{or }o),n}(k) = \frac{1}{K} \sum_{t=0}^{K-1} \eta^{(m)}_{e(\text{or }o),n}(t) \exp(-j2\pi tk/K) \end{cases}. \quad (4)$$

Subcarrier-by-subcarrier STTD decoding is carried out as follows [5]:

$$\begin{cases} \hat{d}_{e,n}(k) = \sum_{m=0}^{N_r-1} \left\{ \hat{H}^{(m)*}_{0,n}(k) R^{(m)}_{e,n}(k) + \hat{H}^{(m)}_{1,n}(k) R^{(m)*}_{o,n}(k) \right\} \\ \hat{d}_{o,n}(k) = \sum_{m=0}^{N_r-1} \left\{ \hat{H}^{(m)*}_{1,n}(k) R^{(m)}_{e,n}(k) - \hat{H}^{(m)}_{0,n}(k) R^{(m)*}_{o,n}(k) \right\} \end{cases}, \quad (5)$$

where $\hat{d}_{e,n}(k)$ and $\hat{d}_{o,n}(k)$ are the soft decision values and $\hat{H}^{(m)}_{0,n}(k)$ and $\hat{H}^{(m)}_{1,n}(k)$ are the channel estimates of $\sqrt{S}H^{(m)}_{0,n}(k)$ and $\sqrt{S}H^{(m)}_{1,n}(k)$, respectively. Assuming ideal channel estimation (i.e., $\hat{H}^{(m)}_{0(\text{or }1),n}(k) = \sqrt{S}H^{(m)}_{0(\text{or }1),n}(k)$) and substituting Eq. (3) into Eq. (5), we have

$$\begin{cases} \hat{d}_{e,n}(k) = d_{e,n}(k) \sum_{m=0}^{N_r-1} \left\{ \left| \sqrt{S}H^{(m)}_{0,n}(k) \right|^2 + \left| \sqrt{S}H^{(m)}_{1,n}(k) \right|^2 \right\} \\ \quad + \sum_{m=0}^{N_r-1} \left\{ \sqrt{S}H^{(m)*}_{0,n}(k) \Pi^{(m)}_{e,n}(k) + \sqrt{S}H^{(m)}_{1,n}(k) \Pi^{(m)*}_{o,n}(k) \right\} \\ \hat{d}_{o,n}(k) = d_{o,n}(k) \sum_{m=0}^{N_r-1} \left\{ \left| \sqrt{S}H^{(m)}_{0,n}(k) \right|^2 + \left| \sqrt{S}H^{(m)}_{1,n}(k) \right|^2 \right\} \\ \quad + \sum_{m=0}^{N_r-1} \left\{ \sqrt{S}H^{(m)*}_{1,n}(k) \Pi^{(m)}_{e,n}(k) + \sqrt{S}H^{(m)}_{0,n}(k) \Pi^{(m)*}_{o,n}(k) \right\}, \end{cases} \quad (6)$$

where the first term is the desired signal component and the second the noise component. It can be understood that the diversity gain of STTD is equal to that of 2-branch receive antenna diversity using maximal ratio combining (MRC), but with a 3 dB power penalty (the reason for this power penalty is because the transmit power per antenna is halved so that the total transmit power is kept the same).

Symbol decision is carried out as

$$\begin{cases} \bar{d}_{e,n}(k) = \arg\min_{d_{e,n}} \left| \hat{d}_{e,n}(k) - d_{e,n} \sum_{m=0}^{N_r-1} \left\{ \left| \hat{H}^{(m)}_{0,n}(k) \right|^2 + \left| \hat{H}^{(m)}_{1,n}(k) \right|^2 \right\} \right|^2 \\ \bar{d}_{o,n}(k) = \arg\min_{d_{o,n}} \left| \hat{d}_{o,n}(k) - d_{o,n} \sum_{m=0}^{N_r-1} \left\{ \left| \hat{H}^{(m)}_{0,n}(k) \right|^2 + \left| \hat{H}^{(m)}_{1,n}(k) \right|^2 \right\} \right|^2 \end{cases}. \quad (7)$$

Then, after parallel-to-serial (P/S) conversion, data demodulation is performed to recover the transmitted data.

3. Pilot-Assisted Decision Feedback Channel Estimation

Figure 2 shows the block diagram of the DFCE. The simultaneous channel estimation for the two transmit channels is described below. Firstly, we estimate the instantaneous frequency transfer functions of the two transmit channels by applying the reverse modulation and then transform each of them into the instantaneous channel impulse response by applying IFFT. Assuming that the real channel impulse response is present only within the GI, the estimated channel impulse response is replaced by zeros beyond the GI in order to reduce the noise as in [10], [11]. Then, applying FFT, the noise-reduced channel transfer functions associated with two transmit antennas are obtained.

From Eq. (3), the channel transfer functions, $\left\{\sqrt{S}H^{(m)}_{0,n-1}(k)\right\}$ and $\left\{\sqrt{S}H^{(m)}_{1,n-1}(k)\right\}$, can be expressed as

$$\begin{cases} \sqrt{S}H^{(m)}_{0,n-1}(k) = \frac{1}{2} \left\{ R^{(m)}_{e,n-1}(k) d^*_{e,n-1}(k) - R^{(m)}_{o,n-1}(k) d_{o,n-1}(k) \right\} \\ \quad + \frac{1}{2} \left\{ \Pi^{(m)}_{o,n-1}(k) d_{o,n-1}(k) - \Pi^{(m)}_{e,n-1}(k) d^*_{e,n-1}(k) \right\} \\ \sqrt{S}H^{(m)}_{1,n-1}(k) = \frac{1}{2} \left\{ R^{(m)}_{e,n-1}(k) d^*_{o,n-1}(k) + R^{(m)}_{o,n-1}(k) d_{e,n-1}(k) \right\} \\ \quad - \frac{1}{2} \left\{ \Pi^{(m)}_{e,n-1}(k) d^*_{o,n-1}(k) + \Pi^{(m)}_{o,n-1}(k) d_{e,n-1}(k) \right\}. \end{cases} \quad (8)$$

The above suggests that if the past data symbols, $d_{e,n-1}(k)$ and $d_{o,n-1}(k)$, are known, $\sqrt{S}H^{(m)}_{0,n-1}(k)$ and $\sqrt{S}H^{(m)}_{1,n-1}(k)$ can be simultaneously estimated by

$$\begin{cases} \bar{H}^{(m)}_{0,n-1}(k) = \frac{1}{2} \left\{ R^{(m)}_{e,n-1}(k) d^*_{e,n-1}(k) - R^{(m)}_{o,n-1}(k) d_{o,n-1}(k) \right\} \\ \bar{H}^{(m)}_{1,n-1}(k) = \frac{1}{2} \left\{ R^{(m)}_{e,n-1}(k) d^*_{o,n-1}(k) + R^{(m)}_{o,n-1}(k) d_{e,n-1}(k) \right\}. \end{cases} \quad (9)$$

Then, IFFT is applied to $\{\bar{H}^{(m)}_{0,n-1}(k)\}$ and $\{\bar{H}^{(m)}_{1,n-1}(k)\}$ to obtain the instantaneous channel impulse response $\{\bar{h}^{(m)}_{0,n-1}(t)\}$ and $\{\bar{h}^{(m)}_{1,n-1}(t)\}$, respectively. $\{\bar{h}^{(m)}_{0,n-1}(t)\}$ and $\{\bar{h}^{(m)}_{1,n-1}(t)\}$ are the noisy estimates, where the noise component is distributed uniformly in the entire range of time delay (t=0∼ $K-1$). Assuming that the real channel impulse response is present only within the GI, the impulse response estimate beyond the GI can be replaced with zeros (or windowing) as follows [10], [11]:

$$\hat{h}^{(m)}_{0(\text{or }1),n}(t) = \begin{cases} \bar{h}^{(m)}_{0(\text{or }1),n-1}(t), & \text{if } 0 \le t \le N_g - 1 \\ 0, & \text{otherwise.} \end{cases} \quad (10)$$

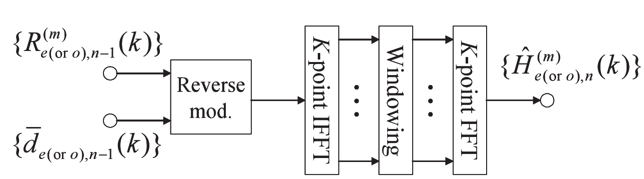


Fig. 2 Structure of DFCE.

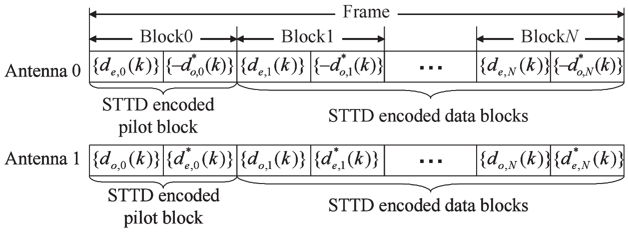


Fig. 3 Frame structure of OFDM with STTD.

Then, FFT is applied to $\{\hat{h}^{(m)}_{0,n}(t)\}$ and $\{\hat{h}^{(m)}_{1,n}(t)\}$ to obtain the channel transfer functions $\{\hat{H}^{(m)}_{0,n}(k)\}$ and $\{\hat{H}^{(m)}_{1,n}(k)\}$, respectively, for STTD decoding of the nth block. Using the above windowing, the noises in $\{\hat{H}^{(m)}_{0,n}(k)\}$ and $\{\hat{H}^{(m)}_{1,n}(k)\}$ can be reduced.

Known pilot can be used as $\{d_{e,n-1}(k)\}$ and $\{d_{o,n-1}(k)\}$. The STTD encoded pilot block can be transmitted for channel estimation of Eq. (9). The frame structure is illustrated in Fig. 3. STTD encoded pilot block is transmitted at the beginning of the frame and followed by N data blocks. If N is large, the tracking ability of the channel estimation against time-varying fast fading tends to be lost. However, the use of small N increases the power loss due to pilot insertion. Therefore, we combine STTD encoded pilot-assisted CE and DF. At the beginning of each frame, channel estimation for STTD decoding of the 1st block is carried out using STTD encoded pilot block. For channel estimation to be used in STTD decoding of the 2nd block onwards, symbol decision of previous STTD encoded block is used as STTD encoded pilot block. Using the above described PA-DFCE, more accurate channel estimation is possible while increasing the tracking ability against fast fading, compared to the use of STTD encoded pilot block only.

Table 1 Simulation conditions.

OFDM	Data modulation	QPSK
	Number of FFT points	K=256
	GI	N_g=32
Fading channel	Channel model	Frequency-selective Rayleigh fading
	Number of paths	L=16
	Power delay profile	Uniform
	Normalized Doppler frequency	f_DT=0.001~0.01
Number of receive antennas		N_r=1, 2

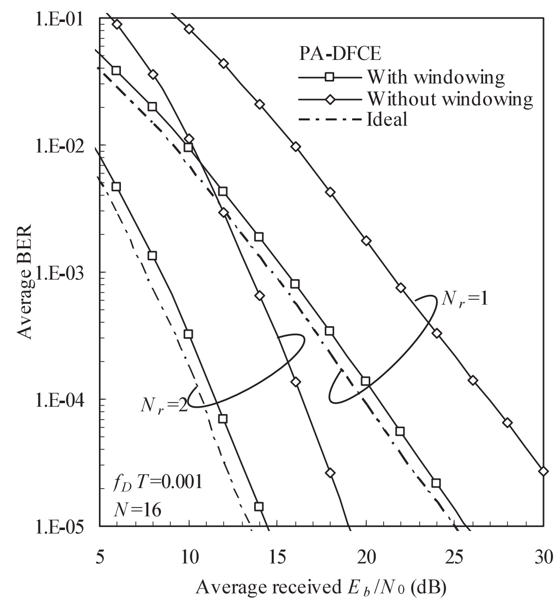


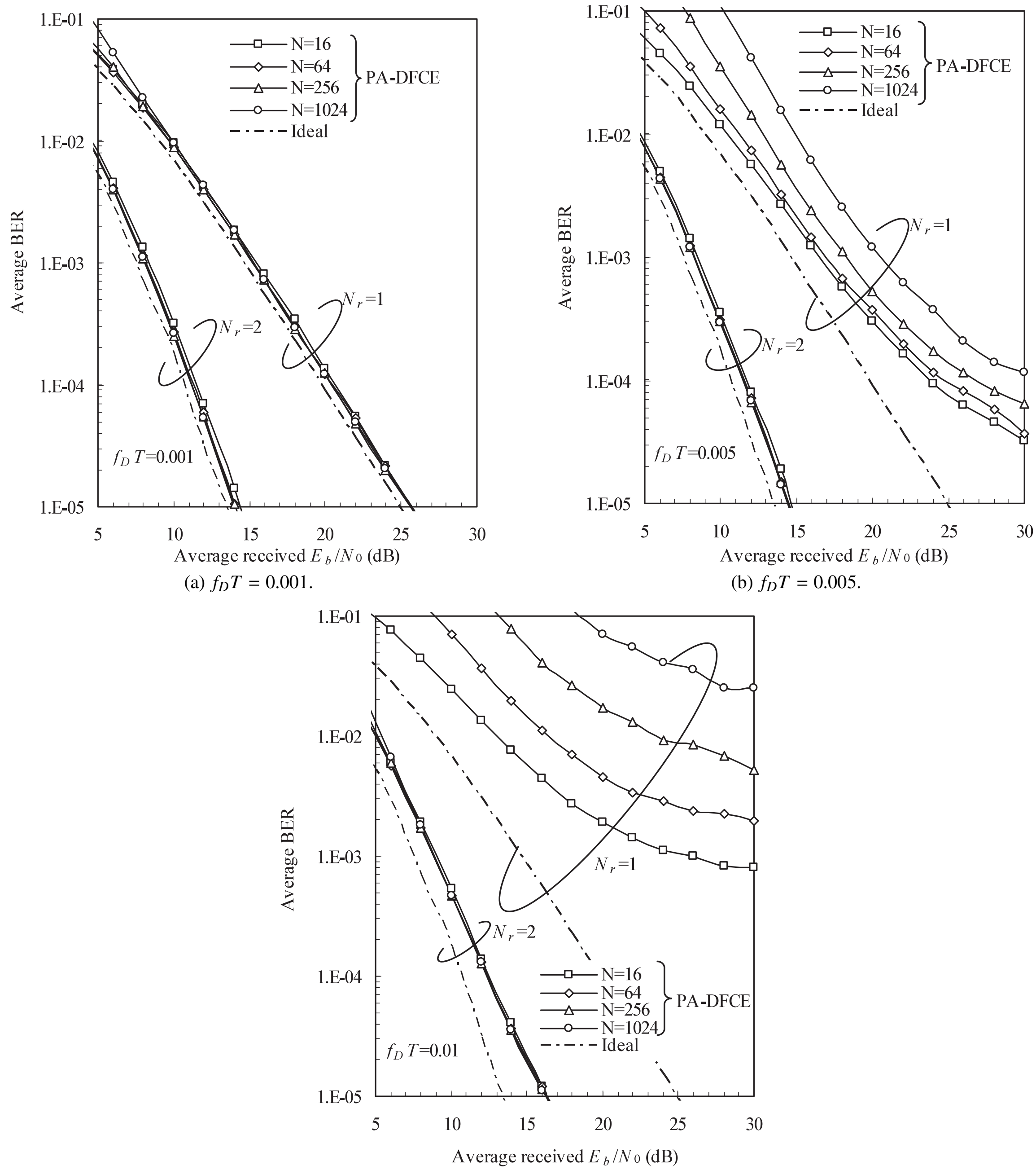
Fig. 4 Average BER performances with and without windowing.

4. Computer Simulation

Table 1 shows the computer simulation conditions. We assume K=256 subcarriers, GI of N_g=32 samples, and quadrature-phase shift keying (QPSK) data modulation. A sample-spaced L=16-path frequency-selective Rayleigh fading channel having a uniform power delay profile is assumed. The normalized maximum Doppler frequency $f_DT = 0.001 \sim 0.01$ is assumed, where T is the OFDM symbol length including GI (i.e., $T = (K + N_g)T_c$).

First, we examine the effect of windowing in the time-delay domain. Figure 4 shows the average BER performances with and without windowing as a function of the average received signal energy per bit-to-the AWGN power spectrum density ratio E_b/N_0 (=0.5 $(ST_c/N_0)(1 + N_g/K)$) when f_DT=0.001 and N=16. It is clearly seen that the use of windowing significantly improves the BER performance. The reason for this significant performance improvement is as follows. When DF and reverse modulation are used, decision errors perturb the estimation of the instantaneous channel transfer function. The instantaneous channel impulse response is obtained by applying FFT to the estimated instantaneous channel transfer function. The perturbation due to decision feedback error is spread over the entire range of time-delay domain and can be effectively reduced by windowing.

Next, we examine the impact of the frame length N. Figure 5 shows the average BER performance with PA-DFCE as a function of the average received E_b/N_0 with N as a parameter when $f_DT = 0.001 \sim 0.01$. When f_DT=0.001, the BER performance of STTD using PA-DFCE is almost insensitive to N. Even with N=1024, almost no performance degradation is seen. The degradation in the required E_b/N_0 for $BER = 10^{-3}$ from the ideal channel estimation case is only about 0.6 dB. Such a small E_b/N_0 degradation from the ideal case is due to noise reduction and prevention of error propagation achieved by windowing in the time-delay domain. For no antenna diversity (N_r=1), the achievable BER performance with PA-DFCE degrades as N increases when f_DT=0.005 and 0.01. However, for two-branch antenna diversity (N_r=2), almost no performance degradation is seen even with N=1024 similar to the case when f_DT=0.001.



(a) $f_D T = 0.001$.

(b) $f_D T = 0.005$.

(c) $f_D T = 0.01$.

Fig. 5 Effect of frame length.

Finally, we compare the BER performances achievable with PA-DFCE and with differential STTD [12] that requires no channel estimation. Figure 6 shows the average BER performances using PA-DFCE and differential STTD with $f_D T$ as a parameter when N=64. The required E_b/N_0 with PA-DFCE for $BER = 10^{-3}$ is smaller by about 2.4 (2.6) dB than with differential STTD for N_r=1 (2) when $f_D T$=0.001. As fading becomes faster, the achievable BER performance with PA-DFCE degrades since the tracking ability of the PA-DFCE tends to be lost. However, without antenna diversity (N_r=1), if $f_D T \leq 0.005$, the STTD with PA-DFCE provides better or at least the same BER performance compared to differential STTD. When two-antenna diversity reception (N_r=2) is used, the proposed channel estimation provides superior BER performance to differential STTD even in a very fast fading environment of $f_D T$=0.01, which is equivalent to a moving speed of 422 km/h for a carrier frequency of 5 GHz and a transmitted data rate of 100 Mbps.

5. Conclusion

In this paper, a pilot-assisted decision feedback channel es-

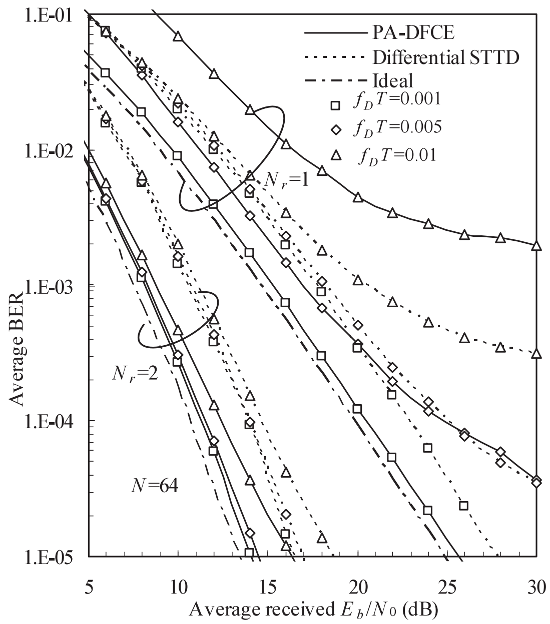


Fig. 6 Performance comparison with differential STTD.

timation (PA-DFCE) suitable for STTD in OFDM signal transmission was proposed. The proposed PA-DFCE can simultaneously estimate the transmit channels by transmitting STTD encoded pilot. The instantaneous channel transfer functions are estimated by reverse modulation using STTD encoded pilot block or decision feedback of the previous decision results. Noise reduction in the channel estimation is achieved by replacing the estimated channel impulse response by zeros beyond GI. Using the decision feedback, very good tracking ability is achieved. It has been confirmed by the computer simulation that STTD using the proposed PA-DFCE provides superior BER performance even in a fast fading environment to differential STTD that requires no channel estimation.

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Appendix

When frequency-domain STTD encoding is used, STTD encoded OFDM signals, $s^0_{e,n}(t)$ and $s^0_{o,n}(t)$, to be transmitted from the 0th antenna are given by

$$\begin{cases} s^0_{e,n}(t) = \sqrt{S}\displaystyle\sum_{k=0}^{K-1} d_{e,n}(k)\exp(j2\pi kt/K), \\ \quad t = 0 \sim K-1 \\ s^0_{o,n}(t) = -\sqrt{S}\displaystyle\sum_{k=0}^{K-1} d^*_{o,n}(k)\exp(j2\pi kt/K), \\ \quad t = K \sim 2K-1, \end{cases} \tag{A$\cdot$1}$$

and STTD encoded OFDM signals, $s^1_{e,n}(t)$ and $s^1_{o,n}(t)$, to be transmitted from the 1st antenna are given by

$$\begin{cases} s^1_{e,n}(t) = \sqrt{S}\displaystyle\sum_{k=0}^{K-1} d_{o,n}(k)\exp(j2\pi kt/K), \\ \quad t = 0 \sim K-1 \\ s^1_{o,n}(t) = \sqrt{S}\displaystyle\sum_{k=0}^{K-1} d^*_{e,n}(k)\exp(j2\pi kt/K), \\ \quad t = K \sim 2K-1. \end{cases} \tag{A$\cdot$2}$$

If the nth even and odd blocks, $s_{e,n}(t)$ and $s_{o,n}(t)$, of OFDM signal before STTD encoding are represented as

$$\begin{cases} s_{e,n}(t) = \sqrt{S}\sum_{k=0}^{K-1} d_{e,n}(k)\exp\left(j2\pi kt/K\right) \\ s_{o,n}(t) = \sqrt{S}\sum_{k=0}^{K-1} d_{o,n}(k)\exp\left(j2\pi kt/K\right) \end{cases}, \qquad \text{(A· 3)}$$

Eq. (A· 1) can be rewritten as

$$\begin{cases} s^0_{e,n}(t) = s_{e,n}(t) \\ s^0_{o,n}(t) = -\sum_{k=0}^{K-1}\left\{\frac{1}{K}\sum_{t'=0}^{K-1} s^*_{o,n}(t')\exp\left(j2\pi kt'/K\right)\right\} \\ \qquad \exp\left(j2\pi kt/K\right) \\ \quad = -\sum_{t'=0}^{K-1} s^*_{o,n}(t')\delta((t'+t) \bmod K) = -s^*_{o,n}(K-t) \end{cases}, \qquad \text{(A· 4)}$$

Similarly, Eq. (A· 2) can be rewritten as

$$\begin{cases} s^1_{e,n}(t) = s_{o,n}(t) \\ s^1_{o,n}(t) = s^*_{e,n}(K-t). \end{cases} \qquad \text{(A· 5)}$$

Eqs. (A· 4) and (A· 5) suggest that STTD encoding in the time-domain, which does not require subcarrier-by-subcarrier STTD encoding, can be implemented as in [13]–[15].

Koichi Ishihara received his B.E. degree in communication engineering from Tohoku University, Sendai, Japan, in 2004. Currently he is a graduate student at the Department of Electrical and Communications Engineering, Tohoku University. His research interests include channel estimation and multiple access interference cancellation techniques for direct sequence and multicarrier CDMA.

Kazuaki Takeda received his B.E. and M.E. degrees in communication engineering from Tohoku University, Sendai, Japan, in 2003 and 2004, respectively. Currently he is pursuing his Ph.D. degree at the Department of Electrical and Communications Engineering, Tohoku University. His research interests include frequency-domain equalization for direct sequence CDMA and transmit/receive diversity techniques. He was the recipient of the 2004 active research award in radio communication systems from IEICE in 2004.

Fumiyuki Adachi See this issue, p.499.