

PAPER

Iterative Channel Estimation for Frequency-Domain Equalization of DSSS Signals

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SUMMARY As the channel frequency selectivity becomes severer, the bit error rate (BER) performance of direct sequence spread spectrum (DSSS) signal transmission with rake combining degrades due to an increasing inter-path interference (IPI). Frequency-domain equalization (FDE) can replace rake combining with much improved BER performance in a severe frequency-selective fading channel. For FDE, accurate estimation of the channel transfer function is required. In this paper, we propose an iterative channel estimation that uses pilot chips which are time-multiplexed within each chip block for fast Fourier transform (FFT). The pilot acts as a cyclic-prefix of FFT block as well. The achievable BER performance is evaluated by computer simulation. It is shown that the proposed channel estimation has a very good tracking ability against fast fading.

key words: DSSS, frequency-domain equalization (FDE), channel estimation

1. Introduction

In direct sequence spread spectrum (DSSS) wireless transmissions, a coherent rake combining can be employed to exploit the channel frequency-selectivity for improving the transmission performance. Wideband DSSS technique has been adopted in the 3rd generation mobile communication systems for data transmissions of up to a few Mbps [1]. In the next generation mobile communication systems, much higher speed data transmission (e.g., higher than several 10 Mbps) is required. However, for such high-speed data transmission, the channel becomes severely frequency-selective [2] and too many rake fingers (or correlators) are required in DSSS and the transmission performance significantly degrades due to large inter-path interference (IPI) even if coherent rake combining is used.

Recently, orthogonal frequency division multiplexing (OFDM) and multi-carrier code division multiple access (MC-CDMA) [3]–[6] have been attracting much attention. However, OFDM and MC-CDMA signals have large peak-to-average power ratio (PAPR) and thus, a linear transmit power amplifier with large peak power is required. More recently, single-carrier (SC) transmission using one-tap frequency-domain equalization (FDE) has been gaining popularity [7]. We have also shown that the use of FDE can significantly improve the bit error rate (BER) performance of DSSS compared to rake combining [8]–[11]. The DSSS

signal transmission has advantages that the problem of high PAPR can be alleviated and that the computational complexity of FDE does not depend on the degree of the channel frequency-selectivity.

Accurate estimation of the channel transfer function (or the channel gain at each frequency) is necessary for FDE. Many works concerning channel estimation can be found in [11]–[19]. The pilot-assisted channel estimation schemes for DSSS (or DS-CDMA) using rake combining, proposed in [17]–[19], can be applied to DSSS signal transmission using FDE. However, in Refs. [17]–[19], the pilot chip block needs to be transmitted frequently to track against fast fading, and as a result, the transmission efficiency (or the data rate) decreases. Recently, a data-dependent pilot structure has been proposed in [20] for the DS-CDMA downlink with FDE. In a data-dependent pilot structure proposed in [20], pilot chip sequence is inserted into each data block for channel estimation. Then, the cyclic prefix (CP) is added to avoid inter-block interference.

In this paper, we propose a new channel estimation scheme using a guard interval (GI) as a pilot which is suitable for FDE and can achieve a very good tracking ability against fast fading without loss of the transmission efficiency. Although the idea of channel estimation using the GI as the pilot has been suggested for the SC transmission with FDE in [7], [21], [22], the implementation of the channel estimation scheme based on the idea was not presented in [7], [21]. The channel estimation proposed in [22] uses a short pilot chip sequence of $N_g/2$ chips and repeats it twice to form a CP of N_g chips. The first pilot chip sequence of $N_g/2$ chips plays a role of the CP of the second pilot chip sequence of $N_g/2$ chips. Hence, channel estimation is possible only if the maximum time delay difference between the propagation paths is less than $N_g/2$ chips. However, our proposed scheme can be used for a channel having maximum time delay difference up to N_g chips.

The objective of this paper is to propose a new pilot-assisted channel estimation scheme and to show how close is the BER performance of our proposed scheme to the ideal channel estimation performance. Therefore, only the uncoded BER performance is presented.

The remainder of this paper is organized as follows. The transmission system model of DSSS transmission with FDE is presented in Sect. 2. The proposed iterative channel estimation scheme is described in Sect. 3. In Sect. 4, the computer simulation results for the BER performance of DSSS transmission with FDE using the proposed channel

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estimation scheme are presented. The paper is concluded in Sect. 5.

2. Transmission System Model of DSSS with FDE

Figure 1 shows the transmitter/receiver structure. In [8]–[10], the transmit data chip sequence is divided into blocks of N_c chips each and the last N_g chips in each chip block are copied and inserted as the CP into the GI placed at the beginning of each block. However, in this paper, the pilot chip sequence of N_g -chip length for channel estimation is inserted at the end of each chip block and the previous pilot chip sequence is used as the GI for the present block. The chip block structure is shown in Fig. 2. The fast Fourier transform (FFT) window size is N_c chips and the number N_d of data chips in each block is $N_d = N_c - N_g$. The transmission efficiency ζ is given by

$$\zeta = \begin{cases} \frac{1}{1 + \frac{1}{N}} \cdot \frac{1}{1 + \frac{N_g}{N_c}} & \text{for conventional block structure} \\ 1 - \frac{N_g}{N_c} & \text{for the block structure assumed in this paper} \end{cases}, \quad (1)$$

where N is the pilot block insertion period for the conventional block structure (i.e., a pilot block of N_c chips is inserted after every N data chip blocks). The transmission efficiency for the block structure used in this paper is higher than the conventional one if $N < (N_c/N_g)^2 - 1$. Moreover, the proposed channel estimation has a higher tracking ability than the pilot-assisted decision feedback channel estimation (PA-DFCE) [11]. This will be discussed later.

Chip-spaced discrete time representation is used throughout the paper. N_d/SF data-modulated symbols are transmitted in each block, where SF is the spreading factor. Without loss of generality, we consider the transmission of one chip block of N_c chips. The data symbol sequence and

the spreading chip sequence in a block are represented by $\{d(n); n = 0 \sim N_d/SF - 1\}$ and $\{c(t); t = 0 \sim N_d - 1\}$, respectively, with $E[|d(n)|^2] = 1$ and $|c(t)| = 1$; $E[\cdot]$ represents the ensemble average operation. The pilot-inserted (or the GI-inserted) DSSS signal $\{s(t); t = 0 \sim N_c - 1\}$, to be transmitted, can be expressed using the equivalent baseband representation as

$$s(t) = \begin{cases} \sqrt{\frac{2E_c}{T_c}} d\left(\left\lfloor \frac{t}{SF} \right\rfloor\right) \cdot c(t) & \text{for } 0 \leq t \leq N_d - 1 \\ \sqrt{\frac{2E_c}{T_c}} p(t) & \text{for } N_d \leq t \leq N_c - 1 \end{cases}, \quad (2)$$

where E_c and T_c represent the chip energy and the chip length, respectively, and $\lfloor x \rfloor$ represents the largest integer smaller than or equal to x and $p(t)$ represents the pilot chip sequence of N_g chips with $|p(t)| = 1$.

The transmitted signal is received by N_r receive antennas at the receiver. Assuming that the channel has L independent propagation paths with T_c -spaced distinct time delays $\{\tau_l; l=0 \sim L-1\}$, the discrete-time impulse response $h_m(t)$ of the multipath channel experienced by the m th antenna is expressed as

$$h_m(t) = \sum_{l=0}^{L-1} h_{m,l} \delta(t - \tau_l), \quad (3)$$

where $h_{m,l}$ is the l th path gain with $\sum_{l=0}^{L-1} E[|h_{m,l}|^2] = 1$. The received signal $r_m(t)$, on the m th antenna, $m=0 \sim N_r - 1$, can be expressed as

$$r_m(t) = \sum_{l=0}^{L-1} h_{m,l} s(t - \tau_l) + \eta_m(t), \quad (4)$$

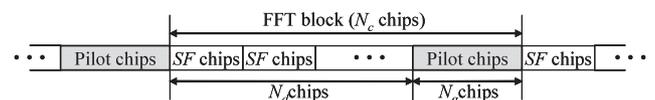


Fig. 2 Chip block structure.

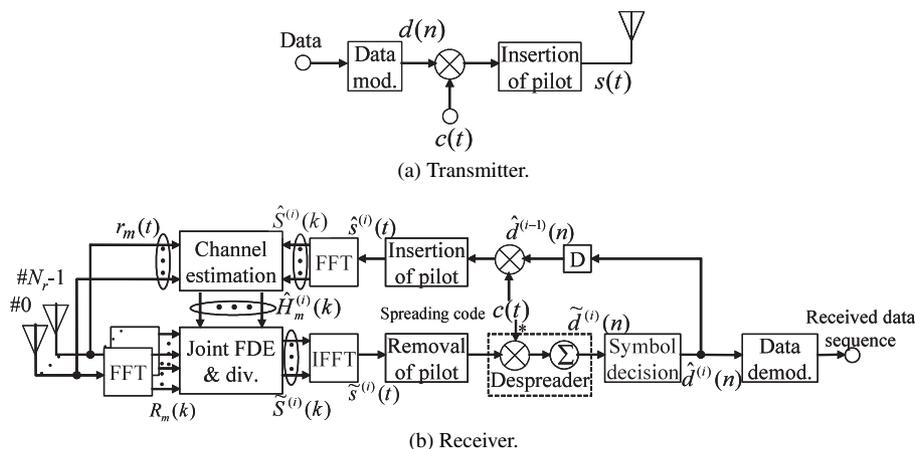


Fig. 1 Transmitter/receiver structure.

where $\eta_m(t)$ represents a zero-mean additive white Gaussian noise (AWGN) process having variance equal to $2N_0/T_c$ with N_0 representing the single sided power spectrum density. Here, we have assumed block fading for the sake of brevity, where path gains remain constant over one chip block; however in the computer simulation, we assume continuous fading.

At the receiver, the received signal $r_m(t)$ is decomposed into N_c frequency components $\{R_m(k); k = 0 \sim N_c - 1\}$ by applying N_c -point FFT. $R_m(k)$ is given by

$$\begin{aligned} R_m(k) &= \sum_{t=0}^{N_c-1} r_m(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ &= H_m(k) \{P(k) + D(k)\} + \Pi_m(k), \end{aligned} \quad (5)$$

where $P(k)$ and $D(k)$ are the k th frequency components of the transmitted pilot and data chip sequences, respectively, and $H_m(k)$ and $\Pi_m(k)$ are the channel gain and noise component due to AWGN at the k th frequency, respectively. They are given by

$$\begin{cases} P(k) = \sum_{t=N_d}^{N_c-1} p(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ D(k) = \sum_{t=0}^{N_d-1} d\left(\left\lfloor \frac{t}{SF} \right\rfloor\right) c(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ H_m(k) = \sqrt{\frac{2E_c}{T_c}} \sum_{l=0}^{L-1} h_{m,l} \exp\left(-j2\pi k \frac{\tau_l}{N_c}\right) \\ \Pi_m(k) = \sum_{t=0}^{N_c-1} \eta_m(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \end{cases} \quad (6)$$

Since the pilot chip sequence $p(t)$ and the spreading chip sequence $c(t)$ are assumed to be random, $E[|P(k)|^2] = N_p$ and $E[|D(k)|^2] = N_d$.

The iterative channel estimation will be described in Sect. 3. The channel estimate for $H_m(k)$ obtained after the i th iteration is denoted by $\hat{H}_m^{(i)}(k)$. Joint FDE and N_r -antenna diversity combining based on the minimum mean square error (MMSE) criterion is carried out using $\hat{H}_m^{(i)}(k)$ to obtain the k th frequency component $\tilde{S}^{(i)}(k)$ as

$$\tilde{S}^{(i)}(k) = \sum_{m=0}^{N_r-1} w_m^{(i)}(k) R_m(k), \quad (7)$$

where $w_m^{(i)}(k)$ is the MMSE weight. We assume that the residual interference is modeled by a zero-mean complex Gaussian process and the sum of residual interference and noise is treated as a new Gaussian noise. $w_m^{(i)}(k)$ is given by [9]

$$w_m^{(i)}(k) = \frac{\hat{H}_m^{(i)*}(k)}{N_c \sum_{m=0}^{N_r-1} |\hat{H}_m^{(i)}(k)|^2 + 2\sigma_i^2}, \quad (8)$$

where $2\sigma_i^2$ is the variance of the sum of noise and residual interference ($\hat{H}_m^{(i)}(k)$ and σ_i^2 are obtained in Sect. 3) and

* denotes the complex conjugate operation. Then, N_c -point inverse FFT (IFFT) is applied to obtain the time-domain signal $\tilde{s}^{(i)}(t)$, which is given by

$$\tilde{s}^{(i)}(t) = \frac{1}{N_c} \sum_{k=0}^{N_c-1} \tilde{S}^{(i)}(k) \exp\left(j2\pi t \frac{k}{N_c}\right). \quad (9)$$

Despreading is performed on $\tilde{s}^{(i)}(t)$ to obtain the decision variable

$$\tilde{d}^{(i)}(n) = \frac{1}{SF} \sum_{t=nSF}^{(n+1)SF-1} \tilde{s}^{(i)}(t) c^*(t), \quad (10)$$

based on which symbol decision is performed. The recovered data symbol is denoted by $\hat{d}^{(i)}(n)$.

In the proposed iterative channel estimation, a series of FDE operation, despreading, symbol decision, respreading and channel estimation is repeated a sufficient number of times. Finally, data-demodulation is carried out.

3. Iterative Channel Estimation

For computing the MMSE weight, $\hat{H}_m^{(i)}(k)$ and σ_i^2 are necessary as shown in Eq. (8). Figure 3 illustrates a detailed structure of the channel estimation block in Fig. 1. Only the pilot is used for the initial channel estimation ($i=0$) and both the pilot and recovered data chips are used as new pilot chips for the first iteration onwards ($i \geq 1$).

In the proposed scheme, the GI is used as the pilot so as not to reduce the transmission efficiency while achieving a good tracking ability against fading. Although the idea of channel estimation using the GI as the pilot has been suggested for the SC transmission with FDE in [7], [21], [22], the implementation of the channel estimation scheme based on the idea was not presented in [7], [21]. The channel estimation proposed in Ref. [22] uses a short pilot chip sequence (whose length is half the GI) and repeats it twice in the GI as shown in Fig. 4. This scheme can only be applied to a channel having a maximum time delay difference less than $N_g/2$ chips. On the other hand, our proposed channel estimation

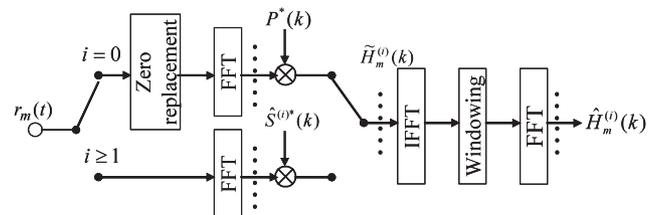


Fig. 3 Channel estimation block.

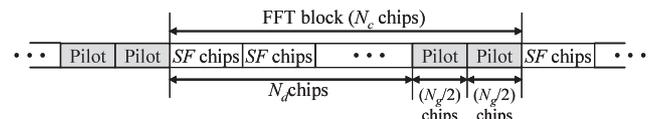


Fig. 4 Chip block structure of the channel estimation scheme proposed in Ref. [22].

scheme can be applied to a channel having a maximum time delay difference up to N_g chips. However, in a frequency-selective fading channel, the received pilot chip block suffers from a large interference from the data chip part in a block and hence, the channel estimation accuracy significantly degrades. Therefore, in the proposed scheme, the initial channel estimation is carried out after the data chip interference is suppressed by applying the rectangular windowing. Then, the channel estimation is carried out again using both the pilot and decision-feedback data sequence as a new pilot. This is repeated a sufficient number of times.

In this paper, we refer to [11] as the conventional channel estimation scheme to compare with the proposed iterative channel estimation. For the channel estimation for DSSS signal transmission with rake combining [17]–[19], the channel impulse response is estimated using the time-domain correlation method and then, the channel transfer function (or the channel gain at each frequency) is estimated by applying the Fourier transform. Another way to directly estimate the channel transfer function is to use the reverse modulation (or removing the pilot modulation) in the frequency-domain as used in the channel estimation for DSSS signal transmission using FDE [11] and also for OFDM signal transmission [12], [13]. Since the channel transfer function is the Fourier transform of the channel impulse response, the channel estimation proposed in [11] is equivalent to those in [17]–[19].

3.1 Initial Channel Estimation Using Pilot Only ($i = 0$)

We assume that the maximum time delay difference $\Delta\tau_{max}$ between the propagation paths is less than the GI, i.e., $\Delta\tau_{max} < N_g$. To reduce the interference from the data chip sequence and the noise, the received signal $r_m(t)$ is replaced with zeros over the time interval of $t = N_g \sim N_d - 1$, where $N_d = N_c - N_g$ (or rectangular windowing), as shown in Fig. 5. Then, N_c -point FFT is applied to decompose the zero-replaced received signal into N_c frequency components:

$$R_m^{(0)}(k) = \sum_{t=0}^{N_g-1} r_m(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) + \sum_{t=N_d}^{N_c-1} r_m(t) \exp\left(-j2\pi k \frac{t}{N_c}\right)$$

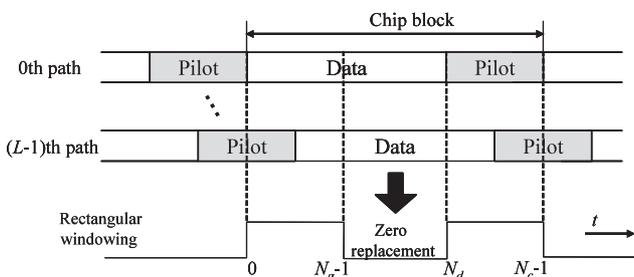


Fig. 5 Zero replacement for the initial channel estimation.

$$= H_m(k)P(k) + D_m^{(0)}(k) + \Pi_m^{(0)}(k), \quad (11)$$

where $D_m^{(0)}(k)$ and $\Pi_m^{(0)}(k)$ are the interference components from the data chip sequence and the noise component, respectively, and are given by

$$\left\{ \begin{aligned} D_m^{(0)}(k) &= \sqrt{\frac{2E_c}{T_c}} \sum_{l=0}^{L-1} \sum_{t=l}^{N_g-1} h_{m,l} d\left(\left\lfloor \frac{t-\tau_l}{SF} \right\rfloor\right) c(t-\tau_l) \\ &\cdot \exp\left(-j2\pi k \frac{t}{N_c}\right) + \sqrt{\frac{2E_c}{T_c}} \sum_{l=1}^{L-1} \sum_{t=N_d}^{N_d+l-1} h_{m,l} d\left(\left\lfloor \frac{t-\tau_l}{SF} \right\rfloor\right) \\ &\cdot c(t-\tau_l) \exp\left(-j2\pi k \frac{t}{N_c}\right) \\ \Pi_m^{(0)}(k) &= \sum_{t=0}^{N_g-1} \eta_m(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) + \sum_{t=N_d}^{N_c-1} \eta_m(t) \exp\left(-j2\pi k \frac{t}{N_c}\right) \end{aligned} \right. \quad (12)$$

We want to estimate $H_m(k)$. First, the instantaneous estimate of the channel gain $H_m(k)$ is obtained by removing the pilot modulation as

$$\begin{aligned} \tilde{H}_m^{(0)}(k) &= \frac{1}{N_p} R_m^{(0)}(k) P^*(k) = \frac{1}{N_p} H_m(k) |P(k)|^2 \\ &+ \frac{1}{N_p} \{D_m^{(0)}(k) + \Pi_m^{(0)}(k)\} P^*(k). \end{aligned} \quad (13)$$

However, $\tilde{H}_m^{(0)}(k)$ is perturbed by the interference and noise. To reduce the interference and noise and then improve the channel estimation, frequency-domain filtering [14] and delay time-domain windowing [11]–[13] can be applied. In this paper, the delay time-domain windowing method is used, as shown in Fig. 6. First, the channel impulse response estimate $\{\tilde{h}_m^{(0)}(\tau)\}$ is obtained by applying N_c -point IFFT to $\{\tilde{H}_m^{(0)}(k)\}$ as follows:

$$\begin{aligned} \tilde{h}_m^{(0)}(\tau) &= \frac{1}{N_c} \sum_{k=0}^{N_c-1} \tilde{H}_m^{(0)}(k) \exp\left(j2\pi k \frac{\tau}{N_c}\right) \\ &= \frac{1}{N_p N_c} \sum_{k=0}^{N_c-1} |P(k)|^2 h_m(\tau) \\ &+ \frac{1}{N_p N_c} \sum_{k=0}^{N_c-1} |P(k)|^2 \left[\sum_{\substack{\tau'=0 \\ \tau' \neq \tau}}^{N_c-1} h_m(\tau') \exp\left(j2\pi k \frac{\tau - \tau'}{N_c}\right) \right] \\ &+ \frac{1}{N_p N_c} \sum_{k=0}^{N_c-1} \left[\{D_m^{(0)}(k) + \Pi_m^{(0)}(k)\} P^*(k) \right] \exp\left(j2\pi k \frac{\tau}{N_c}\right), \end{aligned} \quad (14)$$

where the 2nd and 3rd terms are the interference and noise

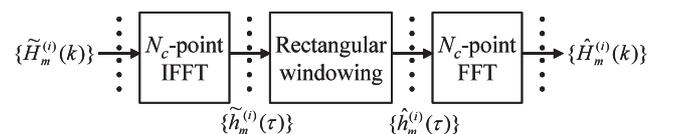


Fig. 6 Delay time-domain windowing.

components, respectively. The interference and noise components are uniformly distributed over the entire range of the delay time ($\tau = 0 \sim N_c - 1$). Assuming that the actual channel impulse response is present only within the GI, the impulse response beyond the GI can be replaced with zeros (or rectangular windowing):

$$\hat{h}_m^{(0)}(\tau) = \begin{cases} \tilde{h}_m^{(0)}(\tau), & \text{if } 0 \leq \tau \leq N_g - 1 \\ 0, & \text{otherwise} \end{cases}. \quad (15)$$

Then, N_c -point FFT is applied to $\{\hat{h}_m^{(0)}(\tau)\}$ to obtain the improved channel gain estimates $\{\hat{H}_m^{(0)}(k)\}$ as follows:

$$\hat{H}_m^{(0)}(k) = \sum_{\tau=0}^{N_c-1} \hat{h}_m^{(0)}(\tau) \exp\left(-j2\pi k \frac{\tau}{N_c}\right). \quad (16)$$

The variance $2\sigma_0^2$ of the sum of residual interference and noise is necessary for computing the MMSE weight of Eq. (8), which can be estimated as

$$2\sigma_0^2 = \frac{1}{N_r N_c} \sum_{m=0}^{N_r-1} \sum_{k=0}^{N_c-1} |R_m^{(0)}(k) - \hat{H}_m^{(0)}(k)P(k)|^2. \quad (17)$$

3.2 Channel Estimation Using Pilot and Data Chips ($i \geq 1$)

The recovered data symbol sequence $\{\hat{d}^{(i-1)}(n)\}$ obtained at the $(i-1)$ th iteration is re-spread and the pilot chip sequence is inserted to generate the transmitted signal replica $\hat{s}^{(i)}(t)$:

$$\hat{s}^{(i)}(t) = \begin{cases} \hat{d}^{(i-1)}\left(\left\lfloor \frac{t}{SF} \right\rfloor\right) \cdot c(t), & 0 \leq t \leq N_d - 1 \\ p(t), & N_d \leq t \leq N_c - 1 \end{cases}, \quad (18)$$

which is decomposed into N_c frequency components $\{\hat{S}^{(i)}(k)\}$ by applying N_c -point FFT. The channel gain estimate $\tilde{H}_m^{(i)}(k)$ is obtained by removing the pilot and data modulation from $R_m(k)$ in Eq. (5) as

$$\tilde{H}_m^{(i)}(k) = \frac{1}{N_c} R_m(k) \hat{S}^{(i)*}(k), \quad (19)$$

where $\hat{S}^{(i)}(k)$ is given by

$$\hat{S}^{(i)}(k) = P(k) + \hat{D}^{(i-1)}(k), \quad (20)$$

where $\hat{D}^{(i-1)}(k)$ is the k th frequency component of the recovered data symbol sequence $\{\hat{d}^{(i-1)}(n)\}$ and is given by

$$\hat{D}^{(i-1)}(k) = \sum_{t=0}^{N_d-1} \left\{ \hat{d}^{(i-1)}\left(\left\lfloor \frac{t}{SF} \right\rfloor\right) \cdot c(t) \right\} \exp\left(-j2\pi k \frac{t}{N_c}\right). \quad (21)$$

The substitution of Eqs. (5) and (20) into Eq. (19) gives

$$\begin{aligned} \tilde{H}_m^{(i)}(k) &= \frac{1}{N_c} H_m(k) \{P(k) + D(k)\} \{P(k) + \hat{D}^{(i-1)}(k)\}^* \\ &\quad + \frac{1}{N_c} \Pi_m(k) \hat{S}^{(i)*}(k). \end{aligned} \quad (22)$$

If the symbol decision in the $(i-1)$ th iteration is correct, the first term in Eq. (22) becomes $H_m(k) |P(k) + D(k)|^2 / N_c$.

Comparing this with Eq. (13) shows that the channel estimation accuracy can be much improved.

To further improve the estimation accuracy, the delay time-domain windowing is applied, similar to the $i=0$ case. Applying N_c -point IFFT to $\{\tilde{H}_m^{(i)}(k)\}$, the channel impulse response estimate $\{\tilde{h}_m^{(i)}(\tau)\}$ is obtained. The impulse response estimate beyond the GI is replaced with zeros and the improved channel gain estimates $\{\hat{H}_m^{(i)}(k)\}$ are obtained by applying FFT. $2\sigma_i^2$ in Eq. (8) for the i th iteration is obtained using

$$2\sigma_i^2 = \frac{1}{N_r N_c} \sum_{m=0}^{N_r-1} \sum_{k=0}^{N_c-1} |R_m(k) - \hat{H}_m^{(i)}(k) \hat{S}^{(i)}(k)|^2. \quad (23)$$

3.3 Complexity Consideration

The computational complexity of the proposed channel estimation is compared with the conventional one [11], based on the number of FFT and IFFT operations. The complexity of the despreading operation is neglected since it is much less than those of FFT/IFFT operations.

In the conventional pilot-assisted channel estimation scheme, the instantaneous channel gain estimate is obtained by reverse modulation of each frequency component of the received pilot block after performing FFT. Then, the IFFT operation and the delay time-domain windowing are applied to obtain the noise-suppressed instantaneous channel impulse response. Finally, the improved channel transfer function (or the channel gain at each frequency) is obtained by applying FFT. Therefore, the total number of FFT/IFFT operations is 3 per pilot block for the conventional channel estimation. If a decision feedback channel estimation is used as in [11], the number of FFT/IFFT operations per block is 3.

Next, we consider the complexity of our proposed scheme. In the initial estimation, the received signal is replaced with zeros over the time interval $t = N_g \sim N_d - 1$ as shown in Sect. 3.1 and then decomposed into frequency components by applying FFT. The estimate of instantaneous channel transfer function is obtained by reverse modulation. Then, IFFT is applied to obtain the instantaneous channel impulse response. After the noise is suppressed by the delay time-domain windowing, the noise-suppressed channel transfer function to be used for FDE is obtained by applying FFT. After performing FDE, IFFT is applied to bring the frequency-domain signal into a time-domain chip sequence to obtain the despreading and tentative symbol decision. Therefore, the total number of FFT/IFFT operations required for the initial channel estimation is 4. The tentatively recovered data symbol sequence is fed back and decomposed into frequency components by applying FFT and the channel transfer function is estimated again using both the pilot and data. Therefore, the total number of FFT/IFFT operations per iteration is 3. In the proposed iterative channel estimation, the above procedure is repeated. However, as shown in Sect. 4, only two iterations are sufficient. As a result, the total number of FFT/IFFT operations for the pro-

posed channel estimation is 10 per block.

The proposed scheme requires 7 more FFT/IFFT operations than the conventional one. Therefore, the proposed scheme provides better tracking ability against fast fading at the cost of increased complexity.

3.4 Comparison with the Channel Estimation Scheme Proposed in [22]

How the channel estimation process of our scheme is different from that of [22] is described below.

The channel estimation proposed in [22] uses a short pilot chip sequence (whose length is half the GI) and repeats it twice in the GI as shown in Fig. 4. At a receiver, the received pilot chip sequence is decomposed into $N_g/2$ orthogonal frequency components by applying $N_g/2$ -point FFT and the pilot modulation is removed from each component. Channel gains at N_c frequencies need to be estimated for performing FDE. However, the channel estimation scheme in [22] provides channel gains at only $N_g/2$ frequencies. Therefore, some interpolation techniques are needed. In [22], the zero padding is used for interpolation. $N_g/2$ -point IFFT is first applied to the channel gain estimates of $N_g/2$ frequencies to get the impulse response over a time delay interval of $N_g/2$ chips. Using zero-padding and N_c -point FFT, the improved channel gain estimates of N_c frequencies are obtained. On the other hand, our proposed scheme uses an N_g -sample pilot, which is affected by the interference from the data chip sequence, and therefore, we first apply the zero replacement to the received chip block in order to suppress the noise and the interference.

In [22], since the first pilot chip sequence of $N_g/2$ chips plays a role of the GI for the second pilot chip sequence of $N_g/2$ chips, no interference is produced from the data chip sequence if the maximum time delay difference between the propagation paths is less than $N_g/2$ chips. However, our proposed channel estimation scheme can be used for a channel having a maximum time delay difference up to N_g chips.

4. Computer Simulation Results

Table 1 shows the computer simulation conditions. We assume block length (=FFT window) of $N_c=1024$ chips, pilot chip length (or GI length) of $N_g=64$ chips, and quadrature-phase shift keying (QPSK) data modulation. A chip-spaced $L=16$ -path frequency-selective Rayleigh fading channel having an exponential power delay profile with a decay factor of α is assumed. Each path gain $h_{m,l}$ is generated based on the Jake's model [2] assuming 64 plane waves coming from all directions with the same amplitude. In the computer simulation, a sequence of 10 million bits is transmitted to compute the average BER. In the next generation mobile communication systems, much higher speed data transmissions than the present systems are required. To achieve high speed data transmissions for the given chip rate (or bandwidth), the spreading factor SF must be small. In high speed downlink packet access (HSDPA), $SF=16$ is

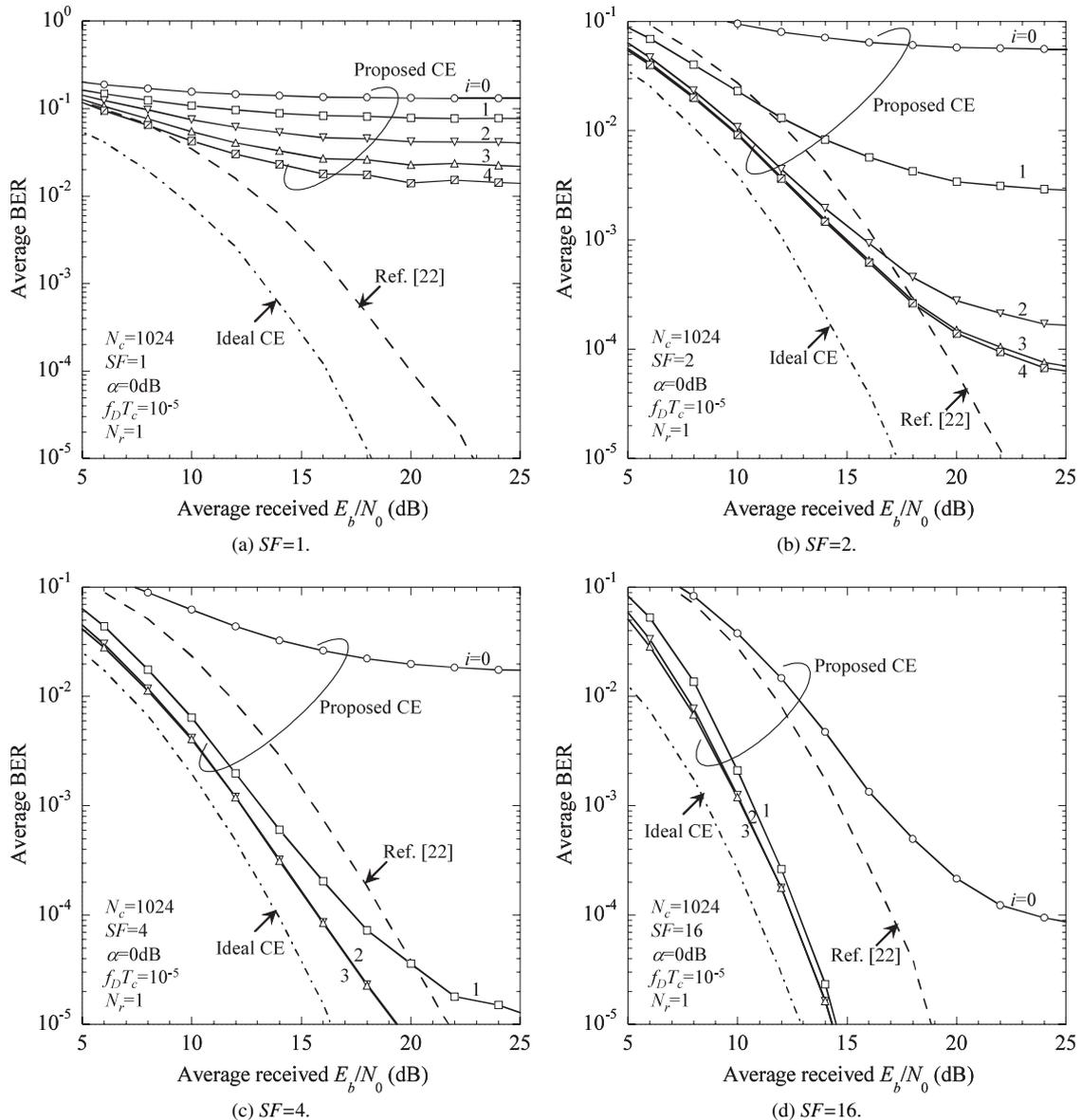
Table 1 Simulation conditions.

Transmitter	Data modulation	QPSK
	Number of FFT points	$N_c=1024$
	Pilot chip length	$N_g=64$
	Spreading factor	$SF=1\sim 16$
Channel	Channel model	Frequency-selective Rayleigh fading
	Power delay profile	$L=16$ -path exponential power delay profile, decay factor $\alpha=0, 8$ (dB)
	Normalized Doppler frequency	$f_D T_c=1\times 10^{-6}\sim 1\times 10^{-4}$
Receiver	Number of received antennas	$N_r=1, 2, 4$
	FDE	MMSE

used [23]. Therefore, $SF=1\sim 16$ is considered. Ideal sampling timing is also assumed at a receiver.

First, we examine the effect of iterative channel estimation (CE) for no antenna diversity ($N_r=1$) case. Figure 7 shows the average BER performance as a function of the average received signal energy per bit-to-the AWGN power spectrum density ratio E_b/N_0 ($=0.5SF(N_c/N_d)(E_c/N_0)$) with the number of iterations as a parameter when $f_D T_c=10^{-5}$, decay factor $\alpha=0$ dB, $N_r=1$, and $\tau_l = l$ (i.e., the maximum time delay difference between the propagation paths is less than $N_g/2$). For comparison, ideal CE and the channel estimation scheme of [22] are also plotted (since the maximum time delay difference of the channel is 15 chips which is shorter than half the GI and therefore, the channel estimation of [22] can be applied). In our proposed scheme, the channel estimation accuracy is degraded due to the interference from the data chips and the performance severely degrades for no spreading case ($SF=1$), i.e., SC transmission. However, since the interference can be suppressed by the despreading process in the DSSS transmission case, our proposed channel estimation scheme provides better BER performances than the channel estimation scheme proposed in [22] when $SF \geq 4$. Even with two iterations ($i=2$), the BER performance is significantly improved and the E_b/N_0 degradation for BER= 10^{-3} , from ideal CE, is only about 1.3 (1.6) dB (including a pilot insertion loss of 0.28 dB) when $SF=4$ (16). The reason for the superiority of our channel estimation scheme to the one proposed in [22] is as follows. At the initial channel estimation ($i=0$), the channel estimation accuracy is poor due to the interference from the data chip sequence. However, since the channel estimation is repeated using both the pilot and the data chip replica obtained from the decision-feedback as a new pilot, the channel estimation accuracy improves. For the perfect decision feedback, the signal-to-noise power ratio (SNR) of the channel estimate is given from Eq. (22) as

$$\frac{S}{N} = \frac{|H(k)|^2 E\{[D(k) + P(k)]^4\}}{E\{[\Pi(k)]^2\} E\{[D(k) + P(k)]^2\} (N_g/N_c)}$$


Fig. 7 Effect of iterative CE.

$$= \frac{E_c}{N_0} \cdot \frac{N_c}{N_g} |H(k)|^2. \quad (24)$$

On the other hand, the average SNR for the channel estimation scheme proposed in [22] is given as

$$\frac{S}{N} = \frac{|H(k)|^2 E[|P(k)|^4]}{E[|\Pi(k)|^2] E[|P(k)|^2]} = \frac{E_c}{N_0} |H(k)|^2. \quad (25)$$

Comparison of Eqs. (24) and (25) shows that the proposed channel estimation in this paper provides N_c/N_g times larger SNR and hence our channel estimation accuracy is better. The channel estimation scheme of [22] can only be applied to a channel having a maximum time delay difference less than $N_g/2$ chips. However, our proposed channel estimation can be applied to a channel having 2-times longer maximum time delay difference. Furthermore, the proposed scheme can provide better BER performance at the cost of increased

computational complexity.

We can see that almost the same E_b/N_0 degradation from ideal CE is seen for both $SF=4$ and 16. Additional BER performance improvement obtained with $i=3$ is very small. Therefore, in what follows, the BER performance is evaluated with $i=2$ assuming $SF=4$.

We examine the impact of the delay factor α . Figure 8 shows the average BER performance with the proposed CE as a function of the average received E_b/N_0 with α as a parameter when $f_D T_c = 10^{-5}$. The proposed CE provides better BER performance than the channel estimation scheme of [22]. The BER performance degrades as α becomes larger (or the channel frequency-selectivity becomes weaker) because of less frequency diversity gain, obtained by FDE. However, the E_b/N_0 degradation from ideal CE for $\text{BER}=10^{-3}$ is almost insensitive to α and only about 1.3 dB.

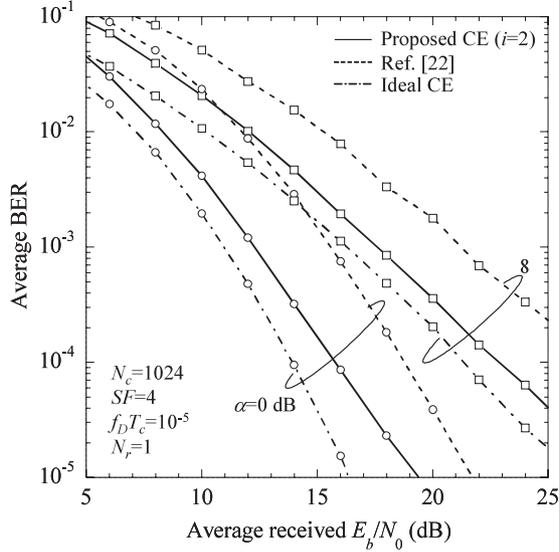
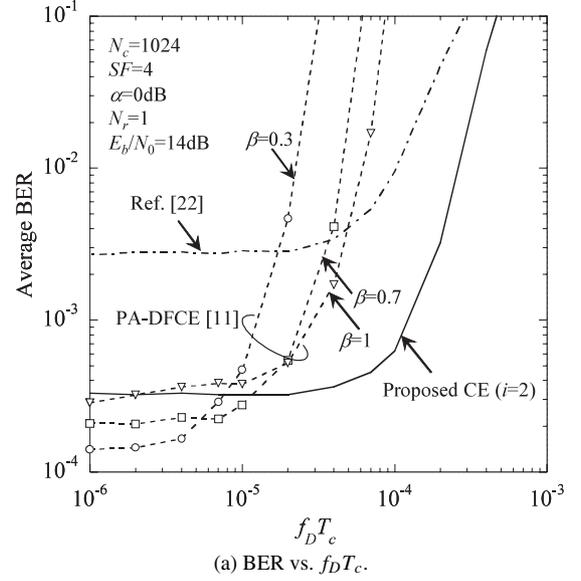


Fig. 8 Impact of decay factor α .

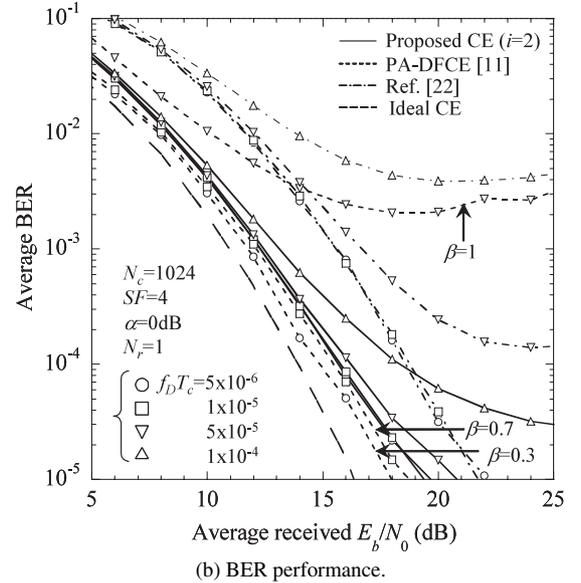
To properly perform FFT, the channel gains must be constant over the FFT window (i.e., a block of N_c chips). However, the channel gains vary within one block for a fast fading. This distorts the signal frequency components obtained by FFT. Here, we examine the impact of the fading rate. Figure 9(a) shows the average BER performance using the proposed CE, as a function of $f_D T_c$. For comparison, the BER performances with PA-DFCE [11] and channel estimation scheme of [22] are also plotted. For PA-DFCE, we assumed that a pilot block of 1024 chips is inserted after every N data chip blocks of 1024 chips each (see Fig. 10). For the same pilot-to-data chip ratio as our proposed CE (i.e., the same data rate for the given chip rate), we have $N=255$ from Eq. (1). In PA-DFCE, the first order filter with the forgetting factor β ($0 \leq \beta \leq 1$) using decision feedback of the previous blocks is employed [11]. It can be seen that the proposed CE provides a slightly worse BER performance than PA-DFCE when $f_D T_c \leq 10^{-5}$. However, as the fading becomes faster (or $f_D T_c$ becomes larger), the tracking ability of PA-DFCE tends to be lost, thereby degrading the achievable BER performance. It can also be seen that the proposed CE provides better BER performance than the channel estimation scheme of [22].

Figure 9(b) shows the average BER performance using the proposed CE as a function of the average received E_b/N_0 with $f_D T_c$ as a parameter. β is optimized for each $f_D T_c$. It can be seen from Fig. 9(b) that as the fading becomes faster, the achievable BER performance of PA-DFCE degrades. However, with the proposed CE, almost no performance degradation is seen even when $f_D T_c = 5 \times 10^{-5}$ ($f_D T_c = 5 \times 10^{-5}$ corresponds to a moving speed of 1.08×10^3 km/h for a carrier frequency of 5 GHz and a chip rate of 100 Mcps).

To clearly show that our proposed scheme is superior to the channel estimation scheme of [22], Fig. 11 plots the BER dependency on the time delay when $L=16$. The l th path



(a) BER vs. $f_D T_c$.



(b) BER performance.

Fig. 9 Impact of fading rate.

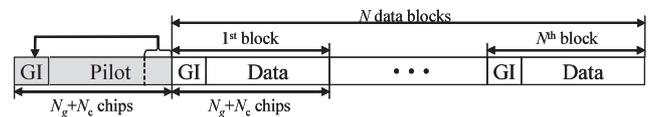


Fig. 10 Block structure for PA-DFCE [11].

time delay τ_l is set to $\tau_l = l, 2l, 3l$ and $4l$ chips ($l=0\sim 15$). Accordingly, the maximum time delay difference $\Delta\tau_{max}$ becomes $\Delta\tau_{max}=15, 30, 45$ and 60 chips, respectively. It can be seen that our proposed channel estimation scheme performs better than the channel estimation scheme of [22]. The BER performance of the proposed scheme is almost insensitive to τ_l as far as the maximum time delay difference $\Delta\tau_{max}$ is shorter than the CP ($N_g=64$). However, the BER performance of the channel estimation scheme of [22] degrades when $\tau_l = 3l$ and $4l$ chips. This is because channel estima-

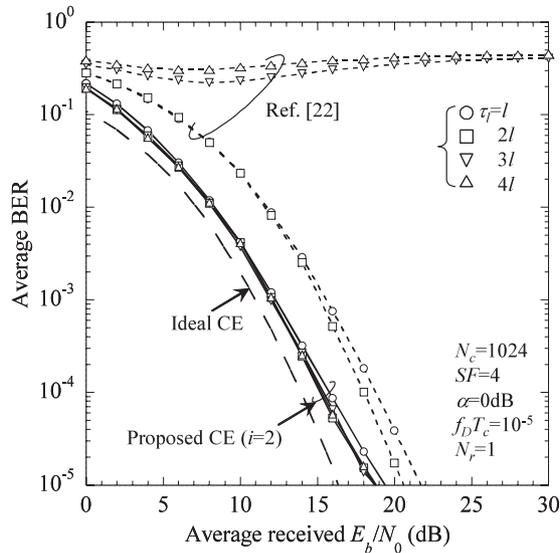


Fig. 11 BER dependency on the time delay.

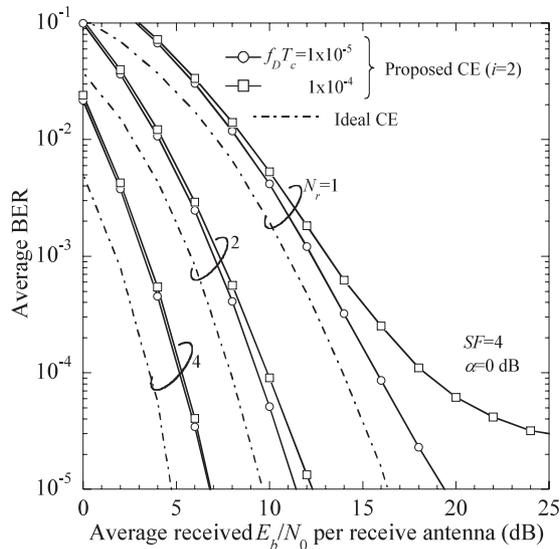


Fig. 12 Effect of antenna diversity.

tion is possible only if the maximum time delay difference $\Delta\tau_{max}$ is shorter than $N_g/2$ chips.

Finally, we examine the BER performance with antenna diversity. It is seen from Fig. 12 that the use of receive antenna diversity is always beneficial. As the fading becomes faster (e.g., $f_D T_c = 10^{-4}$), the achievable BER performance without antenna diversity ($N_r = 1$) degrades and a BER floor appears. However, with antenna diversity reception ($N_r = 2$ and 4), almost no performance degradation is seen even in a very fast fading environment with $f_D T_c = 10^{-4}$.

5. Conclusion

In this paper, an iterative channel estimation scheme suitable for DSSS signal transmission with FDE was proposed and the BER performance was evaluated by computer sim-

ulation. The proposed channel estimation scheme uses a pilot, which is time-multiplexed within the FFT block and therefore, a very good tracking ability against fast fading is achieved. Moreover, since the pilot also acts as the cyclic-prefix, high transmission efficiency is achieved. The simulation results obtained in the paper can be summarized as follows:

- Since the pilot is time-multiplexed within the FFT block, channel estimation accuracy without iteration is very poor due to a large interference from the data chips. However, even two iterations provide sufficient estimation accuracy and the performance loss from the ideal channel estimation in E_b/N_0 is as small as 1.3 dB when $SF=4$.
- Even for a very fast fading (e.g., $f_D T_c = 10^{-4}$), a good tracking ability against fading is obtained and the E_b/N_0 degradation for $BER=10^{-4}$ is as small as 2.0 dB when two-antenna receive diversity is used.

The objective of this paper is to propose a new pilot-assisted channel estimation scheme and to show how close is the BER performance of our proposed scheme to the ideal channel estimation performance. Therefore, in this paper, only the uncoded BER performance was presented. The evaluation of the coded BER performance using the proposed channel estimation scheme is left as a future study.

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