Novel Decision-Directed Channel Estimation Method for TDS-OFDM System

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***Abstract*— Time-domain synchronous orthogonal frequency division multiplexing system (TDS-OFDM) employs known pseudo-noise (PN) sequences as guard interval padding between adjacent OFDM symbols. Accurate estimation of channel impulse response (CIR) is difficult but very important to recover transmitted OFDM symbols by subtracting padded PN sequences. Conventional channel estimation methods only use PN sequences as the training sequences to obtain CIR. Actually, decision-directed channel estimation (DDCE) method which takes advantage of the fact that M-QAM modulated subcarriers in OFDM symbol also contain much valuable hidden-information due to regular constellation mapping and can be used to improve the accuracy of channel estimation. In this paper, the accuracy of channel estimation based on both PN sequence and OFDM symbol is investigated. Moreover, a novel DDCE method based on an accurate initial CIR obtained from two consecutive training sequence and partial decision results of OFDM subcarriers for TDS-OFDM system is proposed. Simulation results show that the proposed algorithm works well over the slow-varying frequency-selective fading channel.**

I. Introduction

Orthogonal frequency-division multiplexing (OFDM) technique has been widely adopted into wide-band and high-speed wireless communication systems, such as digital audio/video broadcasting [1], [2] and wireless local area network [3], etc.

To avoid delay spread impact from adjacent OFDM symbols, all above systems use cyclic prefix (CP) as guard interval (GI). As an alternative to CP-OFDM, time domain synchronous OFDM (TDS-OFDM) [4] is proposed by replacing CP with known pseudo-noise (PN) sequence as GI. Digital television/terrestrial multimedia broadcasting (DTMB) system, the Chinese digital television terrestrial broadcasting standard [5], has adopted TDS-OFDM to achieve both fast channel acquisition/estimation and high spectrum efficiency.

Due to delay spread in wireless channel, the padded PN sequence and the OFDM symbol will interfere against each other at the receive side. Ideally, after subtracting the impact of PN sequences from the received TDS-OFDM signal, the remaining signal can be taken as the zero-padding (ZP) OFDM [6] signal. However, since subtracting the PN sequences and the accurate estimation of channel impulse response (CIR) are interdependent, CIR estimation is very challenging and complex iterative algorithms were proposed. In [7], a channel estimation method via iterative subtraction of PN sequence is presented. Literature [8] addresses this issue for a particular TDS-OFDM system with fixed PN sequence padding, where the ISI-free "virtual" frame is introduced to avoid the interference of PN sequences on OFDM symbols. Literature [9] shows a cyclic reconstruction and tail cancellation method to estimate the CIR efficiently, where the noise impact on OFDM symbol is effectively suppressed with the help of partial decision on the received OFDM symbols. However, those channel estimation methods based on PN sequence usually work in an iterative manner to improve the accuracy of channel estimation and the implementation cost is generally high.

Decision-directed channel estimation (DDCE) method which exploits the data symbol after decision as reference is well known and has been widely investigated in many literatures. In [10], a reliable region decision DDCE method which separates the high bit error rate (BER) region from the low BER region based on the magnitude of the channel frequency response (CFR) is presented. The reliable region is completely equalized, while data in the unreliable region is not. Data in the unreliable region can be equalized via linear interpolation, which may experience some problem in severe frequency-selective channel. A DFT-based DDCE method in which temporary received data symbols are demodulated via hard decision is introduced in [11]. However, this method cannot reduce the probability of error decision and may cause error propagation. DDCE method joint with iterative decoding procedure, i.e. soft decision feedback, are proposed to improve the system performance [12], [13]. However, the complexity of these methods is generally high and sometimes impractical for wireless communication with long time interleaving.

In this paper, we present a DDCE algorithm for TDS-OFDM system based on three observations. First, the energy of OFDM symbols is much higher than that of PN sequences, which leads to more accurate CIR estimation. Second, QPSK/16QAM modulated OFDM sub-carriers are more flat than that of PN sequences, thus the CIR estimation can be simple and efficient, especially for constant magnitude constellation mapping (e.g. QPSK). Third, a partial decision can be used to demodulate the received OFDM symbols before time de-interleaving and forward error code decoding. The partial decision can help recover the OFDM symbols contaminated by noises and estimation errors while keeping low probability of error decision. As a result the demodulated OFDM symbol can be treated as a training sequence to help estimate the CIR. Analysis and simulations show that our proposed algorithm can achieve better accuracy for CIR estimation than those based on PN sequences only.

The rest of this paper is organized as follows. Section II gives a brief introduction to TDS-OFDM system model and the accuracy of channel estimation method based on PN sequence and OFDM symbol is investigated. The novel DDCE algorithm is presented and discussed in Section III. Simulation results are shown in Section IV to demonstrate the proposed algorithm, and Section V concludes this paper.

II. System Model and Channel Estimation

Fig. 1 is the typical baseband TDS-OFDM system assuming perfect synchronization at receiver. The ith transmitted symbol $\{S_{i,k}\}_{k=0}^{N-1}$ is modulated by N-point IFFT and the time-domain signal can be expressed as:

$$s_{i,n} = \frac{1}{N}\sum_{k=0}^{N-1} S_{i,k}\exp(\frac{j2\pi kn}{N}), \quad 0 \le n < N. \tag{1}$$

The elements of the input symbol sequence $\{S_{i,k}\}_{k=0}^{N-1}$ are assumed to be zero-mean random variables with variance σ_S^2 according to constellation mapping. PN sequence $\{c_{i,n}\}_{n=0}^{M-1}$ of length M is padded between OFDM symbols and its power is λ times as much as that of OFDM symbols, thus the variance of PN sequence is $\lambda\sigma_S^2$ (In DTMB system, λ is selected to 1 or 2). The PN sequences and OFDM symbols form the TDS-OFDM signal frames as shown in Fig. 2. To solve the problem of CIR initialization, a slot structure based on TDS-OFDM system is proposed as shown in Fig. 3. The slot consists of a slot head (SH) and F signal frames. With the additional transmission overhead of SH, the spectrum efficiency penalty is about 0.5% with $F = 20$ for PN420 mode of DTMB system.

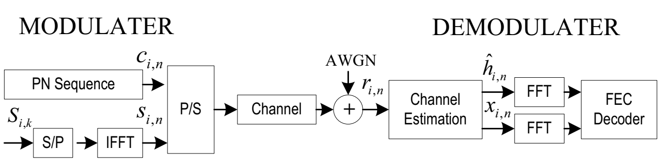

Fig. 1. Baseband model of the TDS-OFDM system.

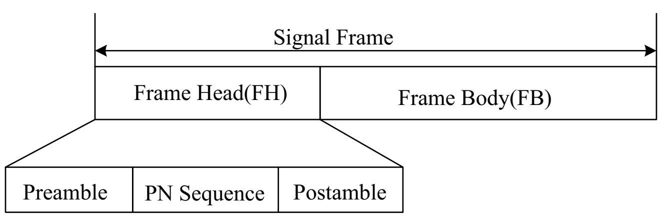

Fig. 2. Frame structure of the TDS-OFDM system.

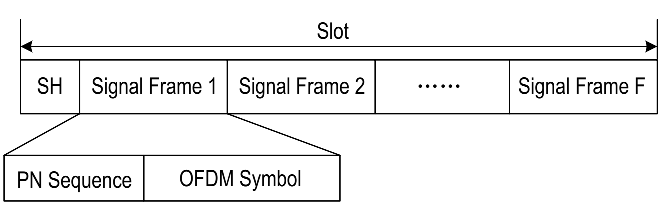

Fig. 3. Slot structure of the proposed TDS-OFDM system.

The transmission channel is modeled as quasi-static Lth order FIR filter, where the CIR $\{h_{i,n}\}_{n=0}^{L-1}$ is static at least within each OFDM symbol period. The enough length of PN sequence ($M \ge L$) is assumed to eliminate the inter-block interference (IBI) between adjacent OFDM symbols. The linear convolution output between the OFDM symbols $\{s_{i,n}\}_{n=0}^{N-1}$ and the CIR is given by:

$$x_{i,n} = s_{i,n} * h_{i,n} + w_{i,n}, \quad 0 \le n < N+L-1 \tag{2}$$

where $*$ denotes linear convolution and $w_{i,n}$ represents the additive white Gaussian noise (AWGN) with variance σ_w^2. The linear convolution output of the PN sequences $\{c_{i,n}\}_{n=0}^{M-1}$ and the CIR is given by:

$$y_{i,n} = c_{i,n} * h_{i,n} + w_{i,n}, \quad 0 \le n < M+L-1. \tag{3}$$

Thus, the ith received signal frame can be expressed as

$$r_{i,n} = \begin{cases} x_{i-1,n+N} + y_{i,n} + w_{i,n}, & 0 \le n < L-1 \\ y_{i,n} + w_{i,n}, & L-1 \le n < M \\ x_{i,n-M} + y_{i,n} + w_{i,n}, & M \le n < M+L-1 \\ x_{i,n-M} + w_{i,n}, & M+L-1 \le n < M+N. \end{cases} \tag{4}$$

At the receiver side, the CIR can be estimated with the help of Discrete-Fourier Transform (DFT).

A. Channel Estimation Based on PN Sequence

With a cyclic reconstruction:

$$\tilde{y}_{i,n} = \begin{cases} y_{i,n} + y_{i,n+M}, & 0 \le n < L \\ y_{i,n}, & L \le n < M \end{cases} \tag{5}$$

where $\{\tilde{y}_{i,n}\}_{n=0}^{M-1}$ is the circular convolution of $\{c_{i,n}\}_{n=0}^{M-1}$ with the CIR and can be rewritten as

$$\tilde{y}_{i,n} = c_{i,n} \otimes h_{i,n} + w_{i,n}, \quad 0 \le n < M \tag{6}$$

where $\otimes$ is the circular convolution operator. Denote $\{C_{i,k}\}_{k=0}^{M-1}$, $\{H_{i,k}\}_{k=0}^{M-1}$,$\{\tilde{Y}_{i,k}\}_{k=0}^{M-1}$ and $\{W_{i,k}\}_{k=0}^{M-1}$ are the M-point DFT of $\{c_{i,n}\}_{n=0}^{M-1}$, $\{h_{i,n}\}_{n=0}^{L-1}$, $\{\tilde{y}_{i,n}\}_{n=0}^{M-1}$ and $\{w_{i,n}\}_{n=0}^{M-1}$, respectively. Then the frequency-domain representation becomes

$$\tilde{Y}_{i,k} = C_{i,k}H_{i,k} + W_{i,k}. \tag{7}$$

To minimize the impact of noise, least-square error (LS) or minimum mean squared error (MMSE) criterion of channel estimation is adopted. Since the complexity of MMSE-based method is high, the LS criterion is taken usually. The LS estimation of H_i is

$$\hat{H}_{i,k} = \tilde{Y}_{i,k}/C_{i,k}. \tag{8}$$

The expectation on squared error between $\hat{H}_i$ and H_i can be represented as

$$E\left(\left\|\hat{H}_i - H_i\right\|^2\right) = E\left(\sum_{k=0}^{M-1}\left\|\frac{W_{i,k}}{C_{i,k}}\right\|^2\right) = \sum_{k=0}^{M-1}\frac{E\left(\left|W_{i,k}\right|^2\right)}{\left|C_{i,k}\right|^2} \tag{9}$$

where $E(\cdot)$ denote the expectation of a random variable. Since

$C_{i,k}$ is the DFT of $c_{i,n}$, $\sum_{k=0}^{M-1} |C_{i,k}|^2 = M \sum_{k=0}^{M-1} |c_{i,k}|^2 = \lambda M^2 \sigma_s^2$, and for the white noise $E(W_{i,k} W_{i,k}^H) = M\sigma_n^2$. Literature [14] shows that if and only if the magnitude of $C_{i,k}$ is constant, the squared error between $\hat{H}_i$ and H_i can be minimized with the constraint of fixed power of PN sequence. The expectation of the squared error is given by

$$
\begin{aligned}
E\left(\left\|\hat{H}_i - H_i\right\|^2\right) &= M\sigma_n^2 \sum_{k=0}^{M-1} \frac{1}{|C_{i,k}|^2} \\
&= M\sigma_n^2 \frac{\alpha M}{\lambda M \sigma_s^2} = \alpha M \frac{\sigma_n^2}{\lambda \sigma_s^2}
\end{aligned}
\tag{10}
$$

where $\alpha = \sum_{k=0}^{M-1} \left(\lambda\sigma_s^2 / |C_{i,k}|^2\right) \geq 1$ is a constant coefficient for known sequence, $\alpha = 1$ if and only if the magnitude of $C_{i,k}$ is constant. After an inverse discrete Fourier transform (IDFT) operation, the CIR estimation is obtained. Since the length of CIR is known to be no longer than L, all the last $(M - L)$ elements of IDFT result can be set to zeros, and the final estimation of CIR $\{h_{i,n}\}_{n=0}^{L-1}$ is obtained. Thus, the expectation of squared error between estimated CIR from PN sequence and real CIR is

$$
E\left(\left\|\hat{h}_{\text{PN},i} - h_i\right\|^2\right) = \frac{L}{M}\frac{1}{M}\alpha_{\text{PN}} M \frac{\sigma_n^2}{\lambda\sigma_s^2} = \frac{\alpha_{\text{PN}} L \sigma_n^2}{\lambda M \sigma_s^2}. \tag{11}
$$

where α_{PN} is the constant coefficient of PN sequence. As a typical PN sequence padding mode in DTMB, PN420 consists of m-sequence of 255 and its cyclic extension. In PN420 mode, for the purpose to estimate longer delay echoes, the whole PN sequence can be used to estimate CIR and the coefficient $\alpha \approx 4$ for the PN420 with DFT of 420-point. In practice, MMSE criterion and its simplified version are adopted to reduce the noise for the PN sequence with large coefficient.

B. Channel Estimation Based on OFDM Symbol

Using PN sequence as training sequence for CIR estimation is discussed above. However, the estimation may not be accurate enough due to various reasons, such as mutual-interferences between PN sequences and OFDM symbols, noise, and channel delay spread. From equation (11), we observe that the accuracy of estimated CIR is inverse proportional to the coefficient α, and proportional to sequence length M and SNR at certain propagation environment with fixed channel length.

However, using PN for CIR estimation does not take advantage of constellation information provided by OFDM symbols. In TDS-OFDM system, OFDM symbol length N is much longer than that of PN sequence. For example, in PN420 padding mode, the length of OFDM symbol is 8 times longer than that of PN sequence and the α of OFDM symbol for QPSK, 16QAM, and 64QAM constellation modulation scheme is 1, 2.78, and 6.15, respectively.

Since information carried by OFDM symbols are unknown and interfered by noise and PN sequences, the decision on the received OFDM symbols will surely not always be reliable. However, the decision error can be modeled as additional noise. The erasure decision, a kind of partial decision, is a simple and effective method to suppress the noise components and reduce the probability of error decision. Taking BPSK modulation with points of (-1, 0) and (1, 0) as an example, the partial decision on $\{\hat{S}_{i,k}\}_{k=0}^{N-1}$ is given by

$$
\tilde{S}_{i,k} = \begin{cases} 1, & \eta_1\sigma_s < \hat{S}_{i,k} < \eta_2\sigma_s \\ -1, & -\eta_2\sigma_s < \hat{S}_{i,k} < -\eta_1\sigma_s \\ \hat{S}_{i,k}, & otherwise \end{cases} \tag{12}
$$

where η_1 and η_2 are the decision thresholds to be chosen appropriately. For BPSK modulation, the probability of error decision can be represented as

$$
\begin{aligned}
P_{\text{BPSK}} &= 2\int_{\eta_1\sigma_s}^{\eta_2\sigma_s} \frac{1}{\sqrt{2\pi}\sigma_n} \exp\left(\frac{-(x+\sigma_s)^2}{2\sigma_n^2}\right) dx \\
&= \text{erfc}\left(\frac{(1+\eta_1)\sigma_s}{\sqrt{2}\sigma_n}\right) - \text{erfc}\left(\frac{(1+\eta_2)\sigma_s}{\sqrt{2}\sigma_n}\right)
\end{aligned}
\tag{13}
$$

where $\text{erfc}(\cdot)$ is the complementary error function. For QPSK/16QAM modulation, the probability of error decision can be obtained straightforward as an extension of BPSK modulation. Denoting ε^2 as the variance of decision error and can be treated as additional noise. Finally, the expectation of squared error between estimated CIR based on OFDM symbols and real CIR is

$$
E\left(\left\|\hat{h}_{\text{OFDM},i} - h_i\right\|^2\right) = \frac{\alpha_{\text{OFDM}} L\left(\sigma_n^2 + E(\varepsilon^2)\right)}{N\sigma_s^2}. \tag{14}
$$

where α_{OFDM} is the constant coefficient of OFDM symbol.

III. Decision-Directed Channel Estimation

In this section, a novel DDCE method is proposed and described in detail. It is assumed that the perfect timing and frequency synchronization are available at the receiver. The proposed algorithm is summarized as follows:

(1) CIR initialization and PN sequences subtraction. The channel estimation method is performed on each signal frame. If the channel estimator starts working at the first time, the CIR initialization estimation method is performed. Since the SH is the same as the first PN sequence, therefore SH is the cyclic prefix of the first PN sequence and an accurate initial CIR can be obtained easily and efficiently. Then, the decision-directed channel estimator is used to track the channel variation and obtain more accurate CIR. Otherwise, the CIR estimation of current frame $\{\hat{h}_{i,n}\}_{n=0}^{L-1}$ and next signal frame $\{\hat{h}_{i+1,n}\}_{n=0}^{L-1}$ are obtained via a linear or high-order interpolation by CIR of previous signal frames [15], then the average of $\{\hat{h}_{i,n}\}_{n=0}^{L-1}$ and $\{\hat{h}_{i+1,n}\}_{n=0}^{L-1}$ is calculated as the CIR estimation for current OFDM symbol , i.e.

$$
\tilde{h}_{i,n} = (\hat{h}_{i,n} + \hat{h}_{i+1,n})/2. \tag{15}
$$

The estimation of the impact of the received PN sequence is obtained by K-point ($K \geq M + L - 1$) FFT/IFFT operation respectively:

$$\hat{y}_{i,n} = \text{IFFT}_K\left[\text{FFT}_K[\hat{h}_{i,n}] \odot C_{i,k}\right] \quad (16)$$

$$\hat{y}_{i+1,n} = \text{IFFT}_K\left[\text{FFT}_K[\hat{h}_{i+1,n}] \odot C_{i+1,k}\right] \quad (17)$$

where $\odot$ denotes the element-wise multiplication operation, and $\text{FFT}_K[\cdot]$ / $\text{IFFT}_K[\cdot]$ denote the K-point fast Fourier transform (FFT) function and inverse FFT function, respectively.

(2) Tail cancellation and cyclic reconstruction for OFDM symbol as

$$z_{i,n}=\begin{cases} r_{i,n+M}-\hat{y}_{i,n+M}+r_{i+1,n}-\hat{y}_{i+1,n}, & 0 \leq n < L \\ r_{i,n+M}, & L \leq n < N \end{cases} \quad (18)$$

where $\{z_{i,n}\}_{n=0}^{N-1}$ is the circular convolution between the ith transmission OFDM symbol and the CIR. We conduct $\{Z_{i,k}\}_{k=0}^{N-1}$ is the N-point FFT of $\{z_{i,n}\}_{n=0}^{N-1}$.

(3) Equalization and partial decision. Equalize $\{Z_{i,k}\}_{k=0}^{N-1}$ in discrete frequency-domain, and the result $\{\hat{S}_{i,k}\}_{k=0}^{N-1}$ is:

$$\hat{S}_{i,k} = \frac{Z_{i,k}}{\text{FFT}_N[\tilde{h}_{i,n}]} \quad (19)$$

Then, partial decision is used to divide the equalized symbol into two groups: decision and non-decision group (i.e., Ω_1 , Ω_2) based upon the OFDM symbol constellation. The decision regions are illustrated in Fig. 4, where the shadow areas represent the decision regions Ω_1 and the other areas are the non-decision regions Ω_2 . Hard decision is made for equalized symbol in frequency domain to the nearest constellation point in region Ω_1 and set the equalized symbol to infinity in region Ω_2, which means that the non-decision symbols are not used in the next step for CIR estimation to suppress the effect of error decision.

(4) CIR estimation based on recovered OFDM symbol. After make partial decision on $\{\hat{S}_{i,k}\}_{k=0}^{N-1}$, the decision result $\tilde{S}_{i,k}$ ($k \in \Omega_1$) and $Z_{i,k}$ ($k \in \Omega_1$) are used to obtain the partial channel frequency response (CFR) estimate:

$$\bar{H}_{i,k} = \frac{Z_{i,k}}{\tilde{S}_{i,k}}, \qquad k \in \Omega_1 \quad (20)$$

Thus, the CIR estimate is given by

$$\bar{h}_{i,n} = \text{IFFT}_N[\bar{H}_{i,k}], \qquad 0 \leq n < N. \quad (21)$$

Since the channel length is known to be smaller than N, $\{\bar{h}_{i,n}\}_{n=0}^{N-1}$ is passed through a rectangular window to set $\{\bar{h}_{i,n}\}_{n=L}^{N-1}$ to zeros. The results are truncated by a threshold to reduce the effect of noise. The insignificant taps with magnitude below the threshold are set to zeros.

(5) CIR power normalization. Since some subcarriers are not used to calculate the CFR, the power of CIR should be normalized, especially for high-order amplitude modulation scheme. Finally, the new CIR is obtained:

$$\hat{\bar{h}}_{i,n} = \beta_i \cdot \bar{h}_{i,n}, \qquad 0 \leq n < L \quad (22)$$

where β_i is the normalization factor.

After processes discussed above, move on to next signal frame. Furthermore, the proposed algorithm can be processed in an iterative manner to achieve better performance.

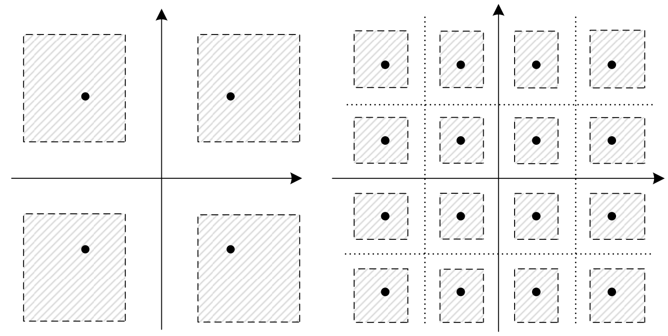

Fig. 4. Partial decision region for QPSK/16QAM modulation.

IV. Simulation Results

In this section, computer simulation is performed to evaluate the performance of proposed DDCE algorithm. The TDS-OFDM system mentioned in Section II is adopted and the major system parameters are shown in Table I. Two typical broadcast multipath channel models with their power delay profiles are given in Table II. Channel I consists of a long delay and large-amplitude echoes for single frequency network (SFN) environment. Channel II is the Brazilian channel model A used for Brazilian field test. It is assumed that each path subjects to an independent Rayleigh fading and the overall path losses has been normalized.

The symbol error rate (SER) as a function of SNR in dB of a QPSK-modulated TDS-OFDM system under above two channel conditions is presented in Fig. 5. The SNR performance under static and Doppler spread conditions based on the proposed DDCE algorithms are presented and compared with the lower bound in which the SER is obtained based on ideal channel estimation. It can be seen from figure that the

TABLE I

Main Simulation Parameters

Symbol Rate f_s	7.56 MSPS
Signal Constellation	QPSK / 16QAM
PN Sequence Length M	420
OFDM Symbol Length N	3780
FFT Size K	2048
Signal Frame Number F	20
Maximal Doppler Spread f_d	0/20 Hz

TABLE II

Power Delay Profiles of the Multipath Channels

	Channel I		Channel II	
Tap	Delay (μs)	Power (dB)	Delay (μs)	Power (dB)
1	-1.8	-18	0	0
2	0	0	0.15	-13.8
3	0.15	-20	2.22	-16.2
4	1.8	-20	3.05	-14.9
5	5.7	-10	5.86	-13.6
6	30	0	5.93	-16.4

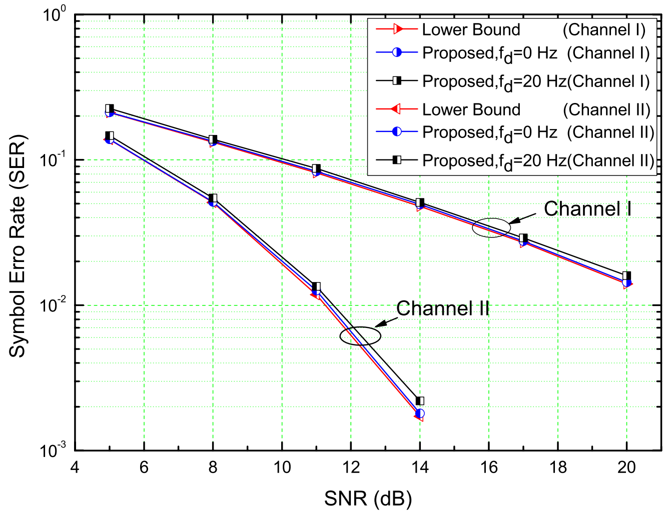

Fig. 5. SER performance of the proposed QPSK TDS-OFDM system.

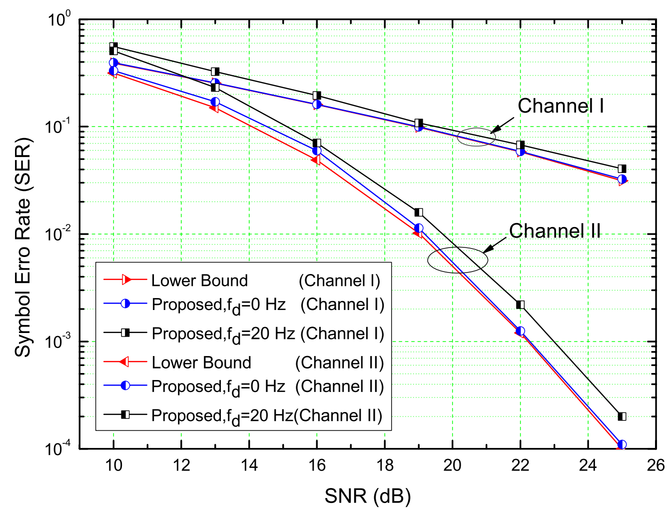

Fig. 6. SER performance of the proposed 16QAM TDS-OFDM system.

proposed algorithm offers almost perfect performance without iteration. Compared with that can be achieved by the lower bound, there is only about 0.3 dB degradation at SER of 10^{-1} for slow-varying fading channel condition with 20Hz Doppler spread.

Fig. 6 shows the SNR performance of the proposed TDS-OFDM system with 16QAM modulation over the same channels. The SNR degradation based on the proposed algorithm is almost identical to that of the lower bounder under the same static channel condition, and about 0.8 dB degradation for 20 Hz Doppler spread condition at SER of 10^{-1}.

V. Conclusion

This contribution presents a DDCE algorithm for TDS-OFDM system, where the initial accurate estimation of CIR is assumed and the partial decision results of OFDM symbols are used to track the channel variations. The updating of CIR estimation and the removal of the residual interference can be performed without iteration to reduce the implementation cost. Analysis shows the proposed method can achieve more accurate channel estimation than those based on the PN sequences only. The simulations demonstrate that the SNR performance based on the proposed channel estimator almost reaches the lower bound under slow-varying frequency-selective fading channels for two typical cases.

The proposed channel estimation algorithm is expected to provide an alternative estimation method for TDS-OFDM system and this method, in principle, can also be applied to CP-OFDM system with minor modification.

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