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Channel Estimation Based on Full ISI Interference Cancellation in TDS-OFDM System

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Abstract: By adopting the same block-type pseudo-random noise (PN) sequences as the guard interval (GI) as well as the training sequence in time domain synchronous orthogonal frequency division multiplexing (TDS-OFDM) system, the received PN signal will be corrupted by the adjacent OFDM symbol sequences interference. A novel channel estimation algorithm is introduced through the received PN data and the tail response. Based on the equalization of linear convolution and circular convolution and residual interference cancellation theorem, Overlap save (OLS) algorithm is utilized to reconstruct the circular convolution from the received PN signal and the initial channel impulse response (CIR) is obtained. With the same PN sequences inserted in TDS-OFDM, traditional frequency-domain equalization can also be employed. The OFDM symbol interference on the received PN signal can thus be estimated and mitigated to correct the received PN signal and update the channel estimation. Simulation results show that the proposed scheme provides large improvement over the conventional method and indicates robustness in channel estimation, even if in the presence of large channel delay. Meanwhile, the proposed algorithm has comparable complexity as the conventional method.

Keywords: TDS-OFDM, Overlap save, Channel estimation, Block-Type PN sequences.

1. Introduction

High-speed data transmission, capacity gains and spectral efficiency are required in the future wireless communication. Although more and more advanced technologies have been proposed, selection of a suitable air interface technology to meet the requirements of wireless communication service in the hot spot and a variety of constraints, such as the limited spectrum and equipment cost, is a thorny problem. Fortunately, orthogonal frequency division multiplexing (OFDM) [1]. which is an attractive transmission technique for high-speed data transmission and robustness against the frequency selection fading caused by the multipath delay spread, is considered as a potential wireless communication access technology. OFDM, as one of the key technologies for next-generation wireless communications, has been adopted in LTE downlink [2].

In traditional OFDM systems, the cyclic prefix (CP) consisting of the tail OFDM symbols, is used to combat the inter-symbol interference (ISI) caused by the multipath delay spread of the channel and the CP length must equal or exceed the length of channel impulse response [3]. Under the long delayed multipath channel, longer CP is needed with the cost of lowering spectral efficiency and power efficiency, specially, if channel estimation (CE) is completed by training sequences in the fast varying channel, the spectral efficiency would be further degraded. Time-domain synchronous orthogonal frequency division multiplexing (TDS-OFDM) has been proposed and applied to DMB-T system [4]-[8].

In TDS-OFDM system, instead of CP, block-type pilots are used as guard intervals (GI) to prevent ISI between the adjacent OFDM symbol sequences and also for channel estimation and synchronization purpose. Compared to conventional OFDM system appending cyclic prefix and training sequences, it can significantly improve the spectral efficiency. However, for the lack of CP, the cyclic property for the received time-domain OFDM signal does not hold any more, so both of the traditional channel estimation algorithm and one-tap frequency domain equalization

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method for CP-OFDM are no longer suitable for the TDS-OFDM system [7].

Several approaches have been proposed to cope with ISI in the insufficient cyclic prefixed OFDM system. Residual ISI cancellation (RISIC) algorithm [8] can eliminate the ISI by tail interference cancellation and cyclic reconstruction. Through the RISIC, most tail cancellation techniques can offer good performance under a moderate length of CIR. However, not only the perfect channel state information (CSI) is assumed, but also the performance of the RISIC algorithm mainly depends on the effect of the first iteration, in which the tail data of the linear convolution of the transmitted signal with the channel impulse response is directly ignored, thus the algorithm is only applied to moderate multipath delay channel. Cyclic prefix reconstruction (CPR) scheme [9] is studied and offer good performance even if the channel delay span is 50% of the OFDM symbol duration. But besides known CSI, the next received signal is directly approximated as the tail data of convolution in the initial cyclic restoration, so the accuracy is still limited. By the equalization between the linear convolution and circular convolution, the processing of tail interference cancellation and data reconstruction can be completed [10]. However, especial frame model is considered in that algorithm, in which sufficient cyclic prefix is only appended to the even-numbered frame. Without CP, the ISI caused by the even-numbered frame on the received odd-numbered frame, would become serious in the presence of large multipath delay, the performance of odd-numbered frame will thus deteriorate.

The perfect CSI is assumed in the traditional interference cancellation methods [8]-[11]. If the length of CP is less than the multipath delay, the channel estimation suffers from the ISI in the traditional CP-OFDM system.

In the TDS-OFDM system, CE has been studied in some literatures. Adaptive filtering algorithm [12] can be used for CE, but many more pilots are required to ensure the algorithm converge in the complicated multipath delay environment. The CIR could be acquired by the time-domain correlation of PN sequences [13]. Since it ignores the tail data, the accuracy of CE is still limited. A frequency-domain channel estimation method is proposed by partial interference cancellation and data reconstruction [14]. It is obvious that the performance of channel estimation will be affected for the incomplete interference cancellation.

In this paper, through the equalization of linear convolution and circular convolution [15], a channel estimation scheme is proposed by the iterative interference cancellation and data reconstruction in the TDS-OFDM system with identical PN sequences inserted. Considering that the received time-domain PN signal will be corrupted for the lack of cyclic prefix, two steps will be taken to complete channel estimation. Firstly, after eliminating the ISI from the forward OFDM data, the received PN sequences and its tail spread are converted into the circular convolution between the transmitted PN sequences and CIR by OLS algorithm and then initial CIR is extracted. With the identical GI inserted, the circular convolution relationship still remains and the traditional one-tap frequency-domain equalizer could also be used to equalize the data composed of OFDM symbol sequences and GI in the second step. The OFDM interference can be reconstructed and eliminated to improve the received PN signal, much more accuracy CIR can thus be obtained. Compared to conventional channel estimation meathod [14], the proposed scheme not only can eliminate the ISI on the received time-domain PN sequences more completely to improve the channel estimation, but also has the comparable complexity.

2. SYSTEM MODEL

2.1. SIGNAL MODEL OF TDS-OFDM

OFDM is a multi-carrier modulation technology, which divides the high-speed data stream into a number of low-speed data streams transmitted mutually on orthogonal subcarriers. With extending duration of the OFDM symbol, the diffuse of the multipath fading channel is reduced and frequency selective fading channel is transformed into several flat fading channels. Then inter-symbol interference is reduced greatly. At the same time, the Fast Fourier Transform based on carrier frequency orthogonal scheme is used in the OFDM system to ensure the orthogonality and improve spectral efficiency. OFDM is particular suitable for multi-users, high data rate communication systems.

As a key technology in the DMB-T in China, the TDS-OFDM technology maintains the advantage of OFDM. Different from the typical CP-OFDM system, pseudo-random PN sequence, instead of CP, is filled periodically as guard interval between each OFDM symbol in the TDS-OFDM. According to Fig.1, the basic transmission frame union consists of frame body composed of OFDM symbol block and frame head composed of the time-domain block PN guard interval. The frame body contains 3780 mapped OFDM symbols. The frame head with the length of 420 comprises a Pre-guard, a PN sequence with the length of 255 and a Post-guard. The first $NUM1$ samples of PN sequence compose the Post-guard and the last $NUM2$ samples of PN sequence compose the Pre-guard.

According to the frame structure of TDS-OFDM system, the Pre-guard, as cyclic prefix of PN sequence in frame head, is still added to avoid the ISI from the OFDM signal. In the case of HDTV broadcasting and satellite OFDM system, the very long delay spreads incite the possibility that the duration of the channel delay exceeds the length of the guard time interval, such as cyclic prefix. With a long CP to solve the problem, it implies a substantial reduction of bandwidth usage, which is very undesirable in the communication system. Therefore, in order to further analyze the effect of the inter-symbol interference on the channel estimation in the larger multipath delay environment, as the detailed analysis below, a simplified data frame structure about TDS-OFDM is exploited.

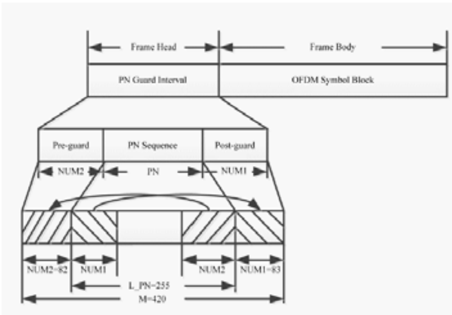

Figure 1 The frame structure of TDS-OFDM system.

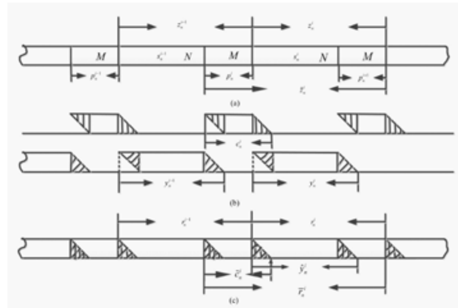

Figure 2 The transmitted and received signals over multipath channel. (a) The transmitted signal frames; (b) Decomposition of the received signal frames; (c) The received signal frames.

2.2. SIMPLIFIED MODEL OF TDS-OFDM

A simplified TDS-OFDM system is shown in Fig.2 (a). Without extra guard interval in the frame head, the PN sequences are inserted among the OFDM data. After encoding and PSK modulation, the data stream is divided into blocks of length N and modulated by using an N-point IFFT. For the ith transmitted OFDM symbol $S_k^i = \left[S_0^i, S_1^i, S_2^i, ..., S_{N-1}^i\right]^T$, the time domain sequence as the output of the IFFT is:

$$s_n^i = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S_k^i \exp\left(\frac{j2\pi kn}{N}\right), 0 \le n < N \qquad (1)$$

According to Fig.2 (a), the block-type pilot $p_n = [p_0, p_1, ..., p_{M-1}]^T$ is appended to the header of each OFDM symbol sequences of length N to form TDS-OFDM signal. The ith transmitted data frame, $\{z_n^i\}_{n=0}^{N_2-1}$, consists of the ith OFDM symbol sequences and tail PN sequences, namely,

$$z_n^i = \begin{cases} s_n^i, 0 \le n < N-1 \\ p_n^{i+1}, N \le n < N_2 \end{cases} \qquad (2)$$

where $\{s_n^i\}_{n=0}^{N-1}$ is the transmitted OFDM symbol sequences and p_n^{i+1} is the $(i+1)\,th$ PN sequences. $N_2 = N + M$ Similarly, the ith super transmitted OFDM block is defined as,

$$\tilde{z}_n^i = \begin{cases} z_n^i, 0 \le n < N_2 - 1 \\ p_n^i, -M \le n < 0 \end{cases} \qquad (3)$$

where $\tilde{z}_n^i$ includes not only the PN sequences appended in the present data frame $\{z_n^i\}_{n=0}^{N_2-1}$, but also the PN sequences appended in the previous data frame $\{z_n^{i-1}\}_{n=0}^{N_2-1}$. Specially, if the identical PN sequences are inserted, $\{p_n^i\}_{n=0}^{M-1} = \{p_n^{i+1}\}_{n=0}^{M-1}$, the PN sequences p_n^i is equivalent to the cyclic prefix of z_n^i.

3. CHANNEL ESTIMATION FOR TDS-OFDM

3.1. Theorem Analysis of the Proposed Scheme

According to the equalization of the linear convolution and the circular convolution [20], the theorem Overlap save (OLS), can be concluded as follows:

Assuming $r = [r_0, r_1, ..., r_{N+M-1}]$ is the linear convolution of $x = [x_0, x_1, ..., x_{N-1}]$ and $h = [h_0, h_1, ..., h_M]$, so,

$$\begin{aligned} r &= \bar{x} \otimes_{N+M} \bar{h} \\ \bar{x} &= \left[x_0, x_1, ..., x_{N-1}, \underbrace{0, ..., 0}_{M} \right] \\ \bar{h} &= \left[h_0, h_1, ..., h_M, \underbrace{0, ..., 0}_{N-1} \right] \end{aligned} \qquad (4)$$

where $\otimes_{N+M}$ is the $N + M$-point circular convolution. From the OLS algorithm, the linear convolution of two sequences can be used to form the circular convolution. Thus, one-tap frequency-domain equalizer is also used to complete the detection to x.

$$x = \Phi_{N\times(M+N)} \times IFFT\left[\frac{FFT\,[r, N+M]}{FFT\,[h, N+M]}\right] \qquad (5)$$

where $\Phi = [I_N, O_{N\times M}]$, I/FFT(X,P) represent P-point IFFT/FFT to X.

The propagation channel is modeled as a quasi-static $L{+}1th$ order FIR filter, where the CIR $\{h_n^i\}_{n=0}^{L}$ remain constant over the super OFDM block $\tilde{z}_n^i$ duration, which is shorter than the coherence time of channel. With the length of GI satisfied $M > L$ in Fig.2, in addition to the noise, the ith received signal, $\{\bar{r}_n^i\}_{n=0}^{N+2M-1}$, comprises three overlapping parts: $\{c_n^i\}_{n=0}^{M+L-1}$, $\{y_n^i\}_{n=0}^{N+L-1}$ and $\{c_n^{i+1}\}_{n=0}^{M+L-1}$, which are given by:

$$\mathrm{c}_n^j = p_n^j * h_n^i = \sum_{l=0}^{L} h_l^i p_{n-l}^j, 0 \le n < M+L, j = i, i+1 \quad (6)$$

$$y_n^i = s_n^i * h_n^i = \sum_{l=0}^{L} h_l^i s_{n-l}^i, 0 \le n < N+L \quad (7)$$

respectively, where $*$ is the linear convolution. By applying (6) (7), the ith received time-domain signal $\{\bar{r}_n^i\}_{n=0}^{N+2M-1}$ with noise $\{w_n^i\}_{n=0}^{N+2M-1}$ is thus expressed as:

$$\bar{r}_n^i = v_n^i + w_n^i, 0 \le n < N+2M \quad (8)$$

that is,

$$v_n^i = \begin{cases} c_n^i + y_{n+N}^{i-1}, 0 \le n < L \\ c_n^i, L \le n < M \\ c_n^i + y_{n-M}^i, M \le n < M+L \\ y_{n-M}^i, M+L \le n < M+N \\ y_{n-M}^i + c_{n-M-N}^{i+1}, M+N \le n < M+N+L \\ c_{n-M-N}^{i+1}, M+N+L \le n < 2M+N \end{cases} \quad (9)$$

where w_n^i represents the AWGN of mean zero and variance σ^2.

According to (9), in the absence of guard interval, both the received time-domain PN sequences and OFDM signal are subject to inter-symbol interference from the adjacent OFDM data or training sequences. Some mathematical analysis to the impact of ISI is derived from the time domain and frequency domain. For convenience but without loss of generality, the ith received OFDM signal $\hat{y}_n^i = \left[\hat{y}_0^i, \hat{y}_1^i, \hat{y}_2^i, ..., \hat{y}_{N+L-1}^i\right]^T$, which is the data response of transmitted OFDM data $\left\{s_n^i\right\}_{n=0}^{N-1}$ and sacrificed by the ISI from the adjacent PN sequences p_n^i and p_n^{i+1}. Without considering the noise, $\hat{y}_n^i$ can be calculated from (9):

$$\hat{y}^i = y_1^i + y_2^i + y_3^i \quad (10)$$

where $y_1^i = h_n^i * s_n^i = A^i \times s^i$, $y_2^i = B^i \times p^i$ and $y_3^i = C^i \times p^{i+1}$. A^i is an $(N+L) \times N$ matrix with the elements given by $A^i\,(k,n) = h_{k-n}^i$, $0 \le k \le N+L-1, 0 \le n \le N-1$ in (12), where h_{k-n}^i is the channel impulse response during the OFDM block interval at the $(k-n)$ channel tap; Both B^i and C^i are $(N+L) \times M$ matrixes and the elements in them are $B^i\,(k,n) = h_{k-n+M}^i, 0 \le k \le N+L-1, 0 \le n \le M-1$ and $C^i\,(k,n) = h_{k-N-n}^i, 0 \le k \le N+L-1, 0 \le n \le M-1$, respectively.

Both B^i and C^i are sparse matrices with structures shown in (13) and (14). With assuming the quasi-static channel over the OFDM block duration, the element of B^i, C^i is time-invariant. y_1^i is the desired time-domain term without ISI, y_2^i is the time-domain ISI component due to the ith PN sequences p_n^i and y_3^i is also the time-domain ISI component due to the $(i+1)\,th$ PN sequences p_n^{i+1}. Demodulating the $\hat{y}_n^i$ by taking the $(N+L)$-point FFT, the frequency-domain signal can be obtained:

$$\hat{Y}^i = Y_1^i + Y_2^i + Y_3^i \quad (11)$$

where

$$\begin{aligned} \hat{Y}^i &= \left[\hat{Y}_0^i, \hat{Y}_1^i, ..., \hat{Y}_{N+L-1}^i\right]^T \\ &= F_{N+L}\hat{y}^i \end{aligned}$$

$$\begin{aligned} Y_1^i &= \left[Y_0^i, Y_1^i, ..., Y_{N+L-1}^i\right]^T \\ &= F_{N+L} y_1^i \\ &= \bar{H}^i \times \bar{X}^i \end{aligned}$$

$$\begin{aligned} Y_2^i &= F_{N+L} y_2^i \\ &= F_{N+L} B^i F_{N+L}^H F_{N+L} y_2^i \\ &= F_{N+L} B^i F_{N+L}^H \times P^i \end{aligned}$$

$$\begin{aligned} Y_3^i &= F_{N+L} y_3^i \\ &= F_{N+L} C^i F_{N+L}^H F_{N+L} y_3^i \\ &= F_{N+L} C^i F_{N+L}^H \times P^{i+1} \end{aligned}$$

where F_{N+L}, F_{N+L}^H represent the $(N+L)$-point FFT and IFFT matrix respectively. Y_1^i is the desired frequency-domain component, which is the ideal frequency-domain term of the OFDM signal over multipath channel. $\bar{H}^i$ is the frequency-domain response of the estimated channel state response $\left\{\hat{h}_n^i\right\}_{n=0}^{L}$. P^i and P^{i+1} is the $(N+L)$-point FFT of PN sequences p_n^i and p_n^{i+1}, respectively.

$$A^i = \begin{bmatrix} h_0^i & 0 & 0 & ... & ... & ... & ... & ... & ... & 0 \\ h_1^i & h_0^i & 0 & 0 & ... & ... & ... & ... & 0 & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ h_L^i & h_{L-1}^i & ... & h_1^i & h_0^i & 0 & ... & ... & 0 & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & 0 & 0 \\ \dots & \dots & \dots & \dots & \dots & h_L^i & \dots & \dots & h_1^i & h_0^i \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \dots & \dots & \dots & \dots & \dots & \dots & \dots & h_L^i \end{bmatrix} \quad (12)$$

$$B^i = \begin{bmatrix} 0 & \dots & 0 & h_L^i & h_{L-1}^i & \dots & h_1^i \\ 0 & \dots & \dots & 0 & h_L^i & \dots & h_2^i \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ \vdots & \vdots & \vdots & \vdots & \vdots & 0 & h_L^i \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 \end{bmatrix} \quad (13)$$

$$C^i = \begin{bmatrix} 0 & 0 & \dots & \dots & \dots & \dots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \dots & \dots & \dots & \dots & \dots \\ h_0^i & 0 & 0 & \dots & \dots & \dots & \dots \\ h_1^i & h_0^i & 0 & \dots & \dots & \dots & 0 \\ \vdots & \vdots & \vdots & 0 & \vdots & \vdots & 0 \\ h_{L-1}^i & \dots & h_1^i & h_0^i & 0 & \vdots & 0 \end{bmatrix} \quad (14)$$

For the lack of guard interval in the transmitted OFDM signal, the received time-domain signal $\hat{y}_n^i$ is corrupted by the ISI. In such case, the additional ISI on the received OFDM signal destroys the orthogonality of subcarriers, resulting in inter-carrier interference (ICI). Without eliminating the ISI, the traditional one-tap frequency-domain equalizer, such as ZF or MMSE, could not be directly applied to the received data. Similarly, the received time-domain PN sequences are also corrupted by the ISI from the OFDM signal. Hence, some solution must be taken to eliminate the ISI to correct the received time-domain PN data. If the "purified" response $\{c_n^i\}_{n=0}^{M+L-1}$ to PN sequence p_n^i can be extracted from $\bar{r}_n^i$ at the receiver, the CIR, $\{h_n^i\}_{n=0}^{L}$, can be obtained simply by OLS algorithm in (4) (5).

3.2. Channel Estimation

Besides for guard interval, the padded PN sequences are also used to finish channel estimation. With the identical PN sequences used in fig.2 (a), TDS-OFDM still fulfills the circular convolution relationship. The received time-domain data $\{r_n^i\}_{n=0}^{N_2-1}$ satisfies:

$$r_n^i = z_n^i \otimes h_n^i + w_n^i, 0 \le n < N_2 \tag{15}$$

where $\otimes$ denotes the circular convolution. Accordingly, the frequency-domain representation by FFT is,

$$R_k^i = H_k^i \times Z_k^i + W_k^i, 0 \le k < N_2 \tag{16}$$

where $R_k^i, Z_k^i, H_k^i, W_k^i$ represent the frequency-domain response of $r_k^i, z_k^i, h_n^i, w_k^i$. If the CIR h_n^i is estimated, simple one-tap frequency-domain equalizer could be used to obtain the transmitted data frame z_n^i.

From (9), the ith received time-domain PN sequences $\{\tilde{c}_n^i\}_{n=0}^{M+L-1}$ can be calculated as,

$$\tilde{c}_n^i = \begin{cases} c_n^i + \underbrace{y_{n+N}^{i-1}}_{ISI-1} + w_n^i, 0 \le n < L \\ c_n^i + w_n^i, L \le n < M \\ c_n^i + \underbrace{y_{n-M}^{i-1}}_{ISI-2} + w_n^i, M \le n < M+L \end{cases} \tag{17}$$

where $\{y_n^{i-1}\}_{n=N}^{N+L-1}$, namely ISI-1, is the interference of $(i-1)\,th$ OFDM symbol sequences on the received initial PN sequences $\{\tilde{c}_n^i\}_{n=0}^{L-1}$ and $\{y_n^i\}_{n=0}^{L-1}$, namely ISI-2, is the interference of OFDM symbol sequences on the received tail PN sequences $\{\tilde{c}_n^i\}_{n=M}^{M+L-1}$. In order to eliminate the inter-symbol interference and reconstruct the circular convolution by OLS algorithm, two steps can be taken. The first step is to remove the ISI-1 in (17) through the estimated $(i-1)\,th$ OFDM symbol $\hat{s}_n^{i-1}$ and CIR $\hat{h}_n^{i-1}$ in the previous signal frame. The partially corrected linear convolution result between the PN sequences p_n^i and h_n^i is approximately calculated as,

$$\bar{c}_n^i = \begin{cases} c_n^i, 0 \le n < M \\ c_n^i + \underbrace{y_{n-M}^{i-1}}_{ISI-2}, M \le n < M+L \end{cases} \tag{18}$$

The circular convolution between the PN sequences p_n^i and h_n^i can thus be reconstructed from the corrected $\bar{c}_n^i$ and the initial CIR $\hat{h}_n^i$ can be extracted by OLS algorithm. The second step is eliminated the ISI-2 in (18). Owing to the identical PN sequences and quasi-static channel, the transmitted data frame $\{z_n^i\}_{n=0}^{n=N_2-1}$ could be estimated through the initially estimated $\hat{h}_n^i$ from (16). The transmitted OFDM symbol sequences $\{\hat{s}_n^i\}_{n=0}^{N-1}$ is also calculated as,

$$\hat{s}_n^i = \hat{z}_n^i, 0 \le n < N \tag{19}$$

Combining the $\hat{h}_n^i$ and $\hat{s}_n^i$ in (7), the ith OFDM symbol sequences response, namely interference ISI-2 on the received PN sequences, can be estimated and mitigated. We could update the received PN sequences $\bar{c}_n^i$ in (18) and obtain more accuracy CIR by OLS algorithm in an iterative method.

The proposed channel estimation algorithm for the ith PN sequences p_n^i can be described as follows:

	Algorithm: Iterative channel estimation by OLS
1:	Start with the separation of $\tilde{c}_n^i$ and r_n^i from the received $\bar{r}_n^i$. p_n^i is converted into $\bar{p}_n^i$ with L zero symbols appended to the tail.
2:	Eliminate the ISI-1 on the $\tilde{c}_n^i$ labeled in (17) to obtain $\bar{c}_n^i$.
3:	Begin iterative, iterative number I
4:	Estimate CIR by OLS algorithm: $\hat{h}_n^i = IFFT\left(\frac{FFT(\bar{c}_n^i)}{FFT(\bar{p}_n^i)}\right)$
5:	Update the CIR: $\hat{h}_n^I = (1-\alpha)\,\hat{h}_n^i + \alpha \hat{h}_n^{I-1}$
6:	Estimate the transmitted data frame $\hat{z}_n^i$. Reconstruct and eliminate the ISI-2 labeled in (18) on the $\bar{c}_n^i$.
7:	Continue till convergence.

After J iterations, the channel state information $\hat{h}^i = \left[\hat{h}_0^i, \hat{h}_1^i, ..., \hat{h}_L^i\right]^T$ can be obtained and its frequency-domain response $\hat{H}^i = \left[\hat{H}_0^i, \hat{H}_1^i, ..., \hat{H}_{N_2-1}^i\right]^T$ is calculated by N_2-point FFT,

$$\hat{H}^i = F_{N_2} \times \hat{h}^i \tag{20}$$

With the assumed data model above, the additional PN sequences are the same, the ith PN sequence p_n^i can thus be considered as the cyclic prefix of the OFDM data frame z_n^i. A simple frequency-domain equalizer, such as ZF, could be used to equalize the received signal r_n^i from (16).

$$Z^i = Q \times R^i - Q \times W^i \tag{21}$$

where R^i and W^i represent the N_2-point FFT to $\left\{r_n^i\right\}_{n=0}^{N_2-1}$ and $\left\{w_n^i\right\}_{n=0}^{N_2-1}$ respectively. Q is $N_2 \times N_2$ diagonal matrix, namely equalization coefficient matrix, whose diagonal element $Q\,(k,k) = 1/\hat{H}_k^i$, $0 \le k \le N_2 - 1$. According to FFT/IFFT, the transmitted OFDM data frame z_n^i can be calculated as,

$$z^i = F_{N_2}^H Z^i = F_{N_2}^H \left(Q \times R^i\right) - F_{N_2}^H \left(Q \times W^i\right) \tag{22}$$

Hence, the ith transmitted time-domain OFDM data can be extracted from z_n^i,

$$\begin{aligned} s^i &= \Phi_{N\times N_2} z^i \\ &= \Phi_{N\times N_2}\left(F_{N_2}^H\left(Q\times R^i\right)\right) \\ &\quad -\Phi_{N\times N_2}\left(F_{N_2}^H\left(Q\times W^i\right)\right) \end{aligned} \tag{23}$$

where $\Phi=[I_N, 0_{N\times M}]$. From (23), the transmitted OFDM symbol,

$$\begin{aligned} S^i &= F_N s^i \\ &= F_N\left[\Phi_{N\times N_2}\left(F_{N_2}^H\left(Q\times R^i\right)\right)\right] \\ &\quad -F_N\left[\Phi_{N\times N_2}\left(F_{N_2}^H\left(Q\times W^i\right)\right)\right] \end{aligned} \tag{24}$$

Here defining,

$$\hat{S}^i = F_N\left[\Phi_{N\times N_2}\left(F_{N_2}^H\left(Q\times R^i\right)\right)\right] \tag{25}$$

where $\hat{S}^i$ are the detected ith OFDM symbol, Hence, It also can be described as,

$$\hat{S}^i = S^i + F_N\left[\Phi_{N\times N_2}\left(F_{N_2}^H\left(Q\times W^i\right)\right)\right] \tag{26}$$

where F_N, $F_{N_2}^H$ represent the N-point FFT and N_2-point IFFT respectively. S^i are the modulation constellation points. Further, (26) can be simplified,

$$\hat{S}^i = S^i + G\times W^i \tag{27}$$

where

$$G{=}F_N\Phi_{N\times N_2}F_{N_2}^H Q \tag{28}$$

$$\begin{bmatrix} \hat{S}_0^i \\ \hat{S}_1^i \\ \vdots \\ \hat{S}_{N-1}^i \end{bmatrix} = \begin{bmatrix} S_0^i \\ S_1^i \\ \vdots \\ S_{N-1}^i \end{bmatrix} + \begin{bmatrix} g_{0,0} & g_{0,1} & \cdots\cdots & g_{0,N_2-1} \\ g_{1,0} & g_{1,1} & \cdots\cdots & g_{1,N_2-1} \\ \vdots & \vdots & \vdots \;\; \vdots & \vdots \\ g_{N-1,0} & g_{N,1} & \cdots\cdots & g_{N-1,N_2-1} \end{bmatrix} \begin{bmatrix} W_0^i \\ W_1^i \\ \vdots \\ W_{N_2-1}^i \end{bmatrix} \tag{29}$$

In (29),

$$\hat{S}_k^i = S_k^i + \sum_{n=0}^{N_2-1} g_{k,n}W_n^i, 0\le k\le N-1 \tag{30}$$

where W_n^i not only satisfies the Gaussian distribution with mean zeros and variance σ^2, but also meets Independent and identically distributed. The distribution function of estimated $\hat{S}_k^i$ can be obtained, namely

$$\hat{S}_k^i \sim N\left(S_k^i, \sigma^2\sum_{n=0}^{N_2-1}\left|g_{k,n}\right|^2\right) \tag{31}$$

For the coded TDS-OFDM system, the estimated OFDM symbol $\hat{S}_k^i$ applied to the decoder for soft-decision.

3.3. Computational Complexity Evaluation

With the parameters for TDS-OFDM system listed in Table 1, the computational complexity comparison between the proposed channel estimation and the conventional method [19] is shown in Table 2. The required number, including addition, multiplication and FFT, is considered as the criterion to evaluate the complexity for the received super OFDM block. From Table II, we can see that the proposed channel estimation has the considerable complexity as the previous scheme, except a small amount of multiplication and addition operation needed in initial channel estimation.

Table 1 Parameter of the TDS-OFDM System

Parameter	Definition
Signal Constellation	QPSK
Channel Coding	Turbo,[13, 15],Ratio=$1/2$
No. of TDS-OFDM Subcarriers N	1024
Symbols in PN Sequence M	400
No. of Iteration J	3

Table 2 Computational Complexity Comparison

Operation	Proposed	Conventional	Proposed $(L=100)$
Addition	$(L+1)\,L/2$	0	5050
Multiplication	$(L+1)\,L/2$	0	5050
800-point FFT	$2J$+1	$2J$+1	$2J$+1
1424-point FFT	$5J$	$5J$	$5J$

Table 3 The COST207 Typical Urban (TU) Channel

Tap	Delay(μs)	Power(dB)
1	0	-3
2	0.2	0
3	0.5	-5
4	1.6	-6
5	2.3	-8
6	5	-10

Table 4 Two-Delayed Path Channel

Tap	Delay(μs)	Power(dB)
1	0	-0.01
2	$1, 2.5, 7.5$	-0.01

4. SIMULATION RESULTS

With all the analysis and discussions above, we can evaluate the performance of the proposed algorithm over two typical channels for the TDS-OFDM system, using major simulation parameters in Table 1. The channel models include COST207 Typical Urban channel described in Table 3 and two delayed paths channel shown in Table 4. The weighting coefficient α is set to 0.125. The mean square error (MSE) is defined as,

$$MSE = \frac{1}{M}\sum_{i=1}^{M}\left|\bar{H}_i - H_i\right|^2 \qquad (32)$$

where $H_i$is the perfect channel frequency response (CFR) and $\bar{H}_i$is the estimated CFR after J iterations.

Fig.3 shows MSE performance comparison of the proposed channel estimator and the conventional one [19] versus the average signal to noise ratio (SNR) in the COST207 TU with the baseband symbol rate of 20MHz and the normalized Doppler shift f_d of 0.01824. Through the initially estimated CIR under the ISI proportion 75%, namely the rate of the channel delay span to the PN sequences duration, the inter-symbol interference on the received pilot sequences can be mitigated in an iterative method. It is clear that the proposed scheme offers large improvement in the MSE performance over the conventional one by the ideal feedback OFDM data in the high SNR. When the MSE maintains 3e-4, the proposed scheme is about 6.0dB better than the conventional algorithm. The reason for the improvement is that the OFDM interference on the received PN sequences is eliminated more completely in the proposed scheme.

Fig.4 illustrates the MSE performance of the proposed channel estimation algorithm versus the SNR for the two delayed paths channel with different normalized Doppler frequency f_d labeled 0.00456 and 0.0228, respectively. The baseband sampling rate is 40KHz and ISI proportion is set to 10%, 25% and 75%, respectively. The proposed scheme achieves robust performance under different channel situations. The advantage of the proposed scheme becomes more obvious with the increase of ISI, promising excellent robustness of our scheme to channel estimation and combat with the multipath delay spread.

The demodulating performance to the OFDM symbol, namely BER-Hard, as a function of SNR defined as the ratio of the average signal power to the noise power, are present in Fig.5 and Fig.6 based on the estimated CIR in the different Doppler shift and ISI proportion environment. The demodulation performance to the proposed method and the lower bound based on the perfect CIR, is depicted in the solid. The dotted line represents the performance without ISI cancellation. According to the results, the proposed method shows a good performance gain based on the channel estimation and data detection. In the low ISI proportion, R=10%, the performance of the proposed way is close to the lower bound in the different normal Doppler frequency set 0.00456 and 0.0228. It is obvious that the

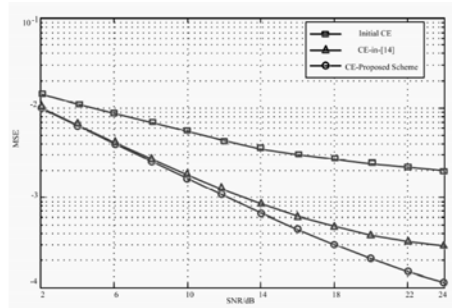

Figure 3 Comparison of the channel estimation.

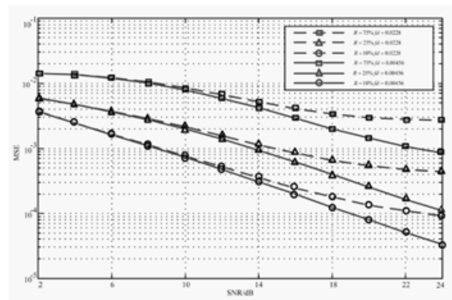

Figure 4 MSE vs. SNR over two-delayed path channel.

means in the paper still work well in a large ISI proportion 75%, but the way, without ISI cancellation, hardly work depicted in Fig.5 and Fig.6. Fig.7 contains similar results for the decoding performance with the normal Doppler frequency 0.0228 in the different ISI proportion. For the ISI proportion of 75%, the BER performance of the proposed method can reduce to 1e-5 in the SNR of 16dB, which is a high gain compared to the way without ISI cancellation.

The following observations can be depicted consistently from the simulation results. Firstly, the proposed channel estimation scheme is superior to the conventional method. The proposed algorithm can effectively eliminate the ISI to avoid the error floor [19] caused by the incompletely ISI cancellation. Secondly, the proposed method promises robustness in the presence of large channel delay and the increase of Doppler frequency.

5. CONCLUSIONS

This letter presents a novel channel estimation scheme in TDS-OFDM system with the same PN sequences inserted. This proposed method can offer significant performance gain over the traditional way and indicates robustness for channel estimation in the time-varying fading channel. Although a small amount of computation is introduced, the total complexity is comparable.

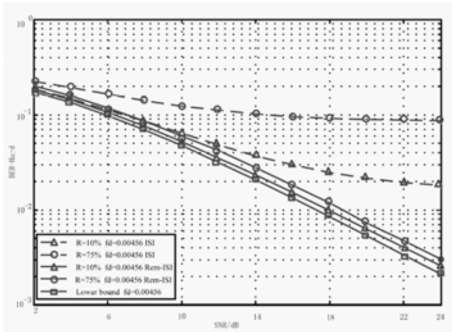

Figure 5 BER-Hard performance of TDS-OFDM with Normal fd=0.00456 over two-delayed path channel.

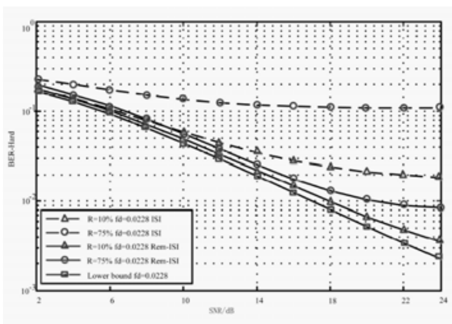

Figure 6 BER-Hard performance of TDS-OFDM with Normal fd=0.0228 over two-delayed path channel.

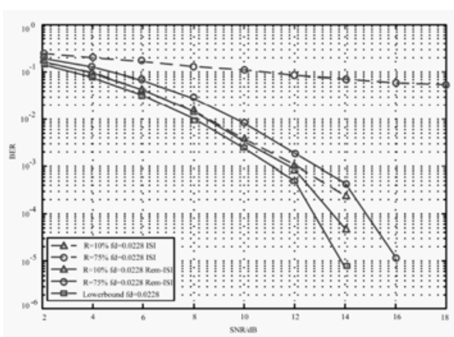

Figure 7 BER performance of TDS-OFDM with Normal fd=0.0228 over two-delayed path channel.

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References

[1] A. R. S. Bahai and B. R. Saltzberg, Multi-Carrier Digital Communications-Theory and Applications of OFDM. Norwell, MA: Kluwer (1999).

[2] 3GPP TS 362 1 1 V840Evolved Universal Terrestrial Radio Access(E-UTRA): Physical Channels and Modulation(Release8,2008-09).

[3] T. Hwang, C. Yang, G. Wu, S. Li and Y. Li. OFDM and Its Wireless Applications: A Survey, IEEE Transactions on Vehicular Technology, 58, 1673-1694 (2009).

[4] Z.X. Yang, L. Tong and L. Yang, Outage probability comparison of CP-OFDM and TDS-OFDM for broadcast channels, IEEE Global Telecommunications Conference, November 17-21, Taiwan ,Taipei.594-598 (2002).

[5] J. Song, Z. Yang and L. Yang. Technical review on Chinese digital terrestrial television broadcasting standard and measurements on some working modes, IEEE Trans. on Broadcasting, 53, 1-7 (2007).

[6] YANG Lin,YANG Zhixing. Terrestrial Digital MultimediaTelevision Broadcasting System: China,00123597.4, (2001).

[7] W. Zhang, Y. Guan and W. Liang et al., An introduction of the Chinese DTTB standard and analysis of the PN595 working modes, IEEE Trans. on Broadcasting, 53, 8-13 (2007).

[8] D. Kim and G.L. Stuber, Residual ISI cancellation for OFDM with application to HDTV broadcasting. IEEE JSAC, 16, 1590-1599 (1998).

[9] C. J. Park and G. Im. Efficient Cyclic Prefix Reconstruction for Coded OFDM Systems, IEEE Communications letters, 8, 274-276 (2004).

[10] W. Bai, Y.Xiao, G. Wu and S. Li. Improved single- carrier frequency domain equalization systems, the 7th International Conference on Information, Communications and Signal Processing, (2009) December 1-5; Macau, SAR China.

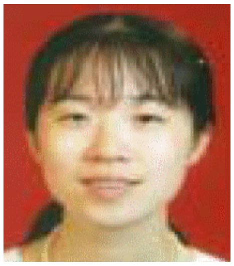
[11] T.H and Y. Li. Iterative cyclic prefix reconstruction for coded single-carrier system with frequency-domain equalization (SC- FDE), the 57th Semiannual Conference of Vehicular Technology, April 22-25, Jeju, Korea, 1841-1845 (2003).

[12] B. Liu, L. Gui, W. Zhang, Bowei Song, On Channel Estimation Method Using Time Domain Sequences in OFDM Systems, IEEE Transactions on Broadcasting, 54, 786-791 (2008).

[13] Z. Zheng, Z. Sun, Robust channel estimation scheme for the TDS-OFDM based digital television terrestrial broadcasting system, IEEE Transactions on Consumer electronics, 54, 1576-1582 (2008).

[14] S. Tang, F. Yang and K. Peng, Iterative Channel Estimation for Block Transmission with Known Symbol Padding - A New Look at TDS-OFDM, IEEE Global Telecommunications. Conference, November 26-30, Washington D.C, US, 4269-4273 (2007).

[15] S. K. Mitra, Digital Signal Processing-A Computer-Based Approach, 3rd, McGraw-Hill, New York (2006).



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