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Channel Estimation for the Chinese DTTB System Based on a Novel Iterative PN Sequence Reconstruction

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***Abstract*—Digital television/terrestrial multimedia broadcasting (DTMB), announced as the Chinese digital television terrestrial broadcasting (DTTB) standard in August, 2006, recommends three PN padding modes as guard intervals (GI) and training sequences for both multi- and single-carrier block transmissions. The conventional channel estimation methods can be performed through either the subtraction of the pseudo-noise (PN) sequence or the cancellation of residual inter-symbol interference, which are quite complex. In this paper, a simple method is introduced to separate the PN sequence from the OFDM symbol and the channel impulse response (CIR) estimation can be obtained and updated via the reconstruction of the PN sequence in an iterative manner. Analysis shows that the proposed algorithm achieves low-complexity channel estimation and can be applied to the DTMB receiver directly. Simulation shows that the proposed algorithm works well under slow-fading broadcast channel even with long delay spread.**

I. INTRODUCTION

ORTHOGONAL frequency–division multiplexing (OFDM) is an attractive transmission technique and has been widely adopted into broadband wireless communication systems, such as digital video broadcasting (DVB) [1], wireless local area network [2], etc.

Conventionally, adjacent OFDM symbols are separated by the cyclic-prefix (CP) which converts the linear convolution to the circular convolution between transmitted signal and channel impulse response (CIR) to simplify the data recovery at receiver. Time domain synchronization OFDM (TDS-OFDM) uses pseudo-noise (PN) sequences padding as the guard interval (GI) as well as training sequence, which makes the synchronization faster than CP-OFDM systems [3]. Since no continuous and scattered pilots are needed, TDS-OFDM can reduce the overall transmission overhead and provide higher spectrum efficiency than that of CP-OFDM. Digital television/terrestrial multimedia broadcasting (DTMB) adopts TDS-OFDM as the baseline modulation technology [4].

However, with the PN sequences padding, the orthogonality between the OFDM symbols does not hold any more. As a result, conventional channel estimation algorithms derived in the CP-OFDM system can not be directly applied to the TDS-OFDM system. In [5], the CIR is estimated in time domain based on the cross correlation between the received signal and the local PN sequences. It may be problematic in the presence of long channel delay-spread and make the implementation difficult as the complexity grows linearly with the square of the PN length. Another way to solve this problem is to process the padded PN sequence and the data part separately [6]. Based on discrete Fourier transformation (DFT) of the received and local PN sequences, the channel estimation is performed in frequency domain. After subtracting the padding PN sequence from the TDS-OFDM signal, the remaining signal can be taken as the Zero Padding (ZP) OFDM and all the techniques for the ZP-OFDM system can be applied. Literature [7] introduces a method of iterative threshold detection is introduced which combined with decision feedback equalization (ITD-DFE) to improve the accuracy of channel estimation in the time domain. However, the above algorithms are very complicated and implementation cost is generally high.

In this paper, we present an iterative channel estimation algorithm with low complexity, which is similar to the residual inter-symbol interference cancellation (RISIC) algorithms in the CP-OFDM system [8] and cyclic prefix reconstruction method for coded single-carrier system with frequency-domain equalization (SC-FDE) [9]. Different from the method in [6] where the overlapping part of PN sequences and the OFDM symbols are used for channel estimation, our proposed algorithm reconstructs the overlapping part and uses the reconstruction for channel estimation. This idea is originated from the observation that the non-overlapping and overlapping parts in the received PN sequence are related. Moreover, the power of PN sequence is twice as much as that of OFDM symbol.

The rest of this paper is organized as follows. Section II describes the frame structure of the DTMB system. In section III, the proposed iterative channel estimation algorithm is discussed. Simulation results are presented in section IV to demonstrate effectiveness of the proposed technique, and conclusions are made in Section V.

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II. SYSTEM MODEL OF DTMB

In the physical layer transmission protocol of the DTMB system, the PN sequence, i.e., frame header (FH), is padded between frame bodies (FB) as GIs for either multi- or single-carrier block transmission (as shown in Fig.1). The FH consists of a preamble, a PN sequence, and a postamble. Here the PN sequence is defined as an m-sequence and can be generated with a Fibonacci-type linear feedback shift register (LFSR) and both pre- and post-amble are the cyclic extension of the m-sequence.

A simplified baseband DTMB system is shown in Fig.2. The DTMB system has two options on the number of sub-carrier C within each FB, C=1 for single carrier mode and C=3780 for multi carrier mode. In this paper, only multi-carrier mode is discussed.

For ith transmitted symbol $\{S_{i,n}\}_{n=0}^{N-1}$, the IFFT output is:

$$s_{i,k}=\frac{1}{\sqrt{N}}\sum_{n=0}^{N-1}S_{i,n}\exp(-\frac{2\pi kn}{N}),\quad 0\le k<N \qquad (1)$$

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. The PN sequences $\{c_{i,k}\}_{k=0}^{M-1}$ of length M are inserted between FBs and the FHs and FBs form the DTMB signal frames. The ith transmitted frame $\{f_{i,k}\}_{k=0}^{M+N-1}$ as shown in Fig.3 (a) is defined as the ith FB with a length-M FH padding:

$$f_{i,k}=\begin{cases}c_{i,k}, & 0\le k<M\\ s_{i,k-M}, & M\le k<M+N\end{cases} \qquad (2)$$

The transmission channel is modeled as a quasi-static Lth order FIR filter, where CIR $\{h_{i,k}\}_{k=0}^{L-1}$ does not vary within ith OFDM symbol period. The received signal will be the linear correlation between signal frames and the enough length of guard interval is assumed to mitigate the multi-path effect ($M\ge L$). As shown in Fig.3 (b), the linear convolution output between the FB $\{s_{i,k}\}_{k=0}^{N-1}$ and the CIR is given by:

$$x_{i,k}=s_{i,k}*h_{i,k},\quad 0\le k<N+L-1 \qquad (3)$$

where * denotes linear convolution. And $\{y_{i,k}\}_{k=0}^{M+L-2}$ denote the linear convolution output of the FH $\{c_{i,k}\}_{k=0}^{M-1}$ and the CIR:

$$y_{i,k}=c_{i,k}*h_{i,k},\quad 0\le k<M+L-1 \qquad (4)$$

The linear convolution phenomenon is illustrated in Fig.3. The shadow areas represent the tails overlap in time domain after propagation over a multi-path environment of the FH or FB due to the linear convolution of the transmitted signal with the CIR. As shown in Fig.3(c), in a received signal frame, the FB is corrupted by the tail of the FH sequence in the same signal frame, and the FH is corrupted by the tail of the FB in the previous signal frame. The expression of the ith received signal frame $\{r_{i,k}\}_{k=0}^{M+N+L-2}$ is given by:

$$r_{i,k}=f_{i,k}*h_{i,k}+w_{i,k},\quad 0\le k\le M+N+L-1 \qquad (5)$$

where $w_{i,k}$ represents the additive white Gaussian noise (AWGN) with variance σ_w^2.

From (2) to (4), we have

$$r_{i,k}=\begin{cases}x_{i-1,k+N}+y_{i,k}+w_{i,k}, & 0\le k<L-1\\ y_{i,k}+w_{i,k}, & L\le k<M\\ x_{i,k-M}+y_{i,k}+w_{i,k}, & M\le k<M+L-1\\ x_{i,k-M}+w_{i,k}, & M+L-1\le k<M+N\\ x_{i,k-M}+y_{i+1,k-M-N}+w_{i,k}, & M+N\le k<M+N+L-1\end{cases} \qquad (6)$$

If $\{y_{i,k}\}_{k=0}^{M+L-2}$ can be extracted from the received signal, the channel estimation $\{\hat{h}_{i,k}\}_{k=0}^{L-1}$ will be obtained simply by one-tap frequency-domain equalization and the rest signal will be the FB free from signal interference over multi-path channel. The received FB $\{X_{i,k}\}_{k=0}^{N-1}$ frequency-domain and the estimated channel frequency response (CFR) $\{\hat{H}_{i,k}\}_{k=0}^{N-1}$ are applied to the forward error correction (FEC) decoder for the soft-decision, where $\{X_{i,k}\}_{k=0}^{N-1}$ and $\{\hat{H}_{i,k}\}_{k=0}^{N-1}$ represent the N-point discrete Fourier transform of $\{x_{i,k}\}_{k=0}^{N+L-2}$ and $\{\hat{h}_{i,k}\}_{k=0}^{L-1}$, respectively. Without considering the noise, we can express the estimation of $\{y_{i,k}\}_{k=0}^{M+L-2}$ as

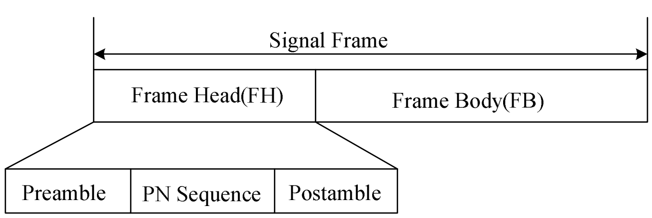

Fig. 1. Frame structure of the DTMB signal.

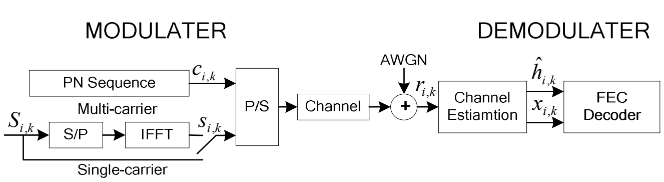

Fig. 2. Block diagram of the DTMB signal.

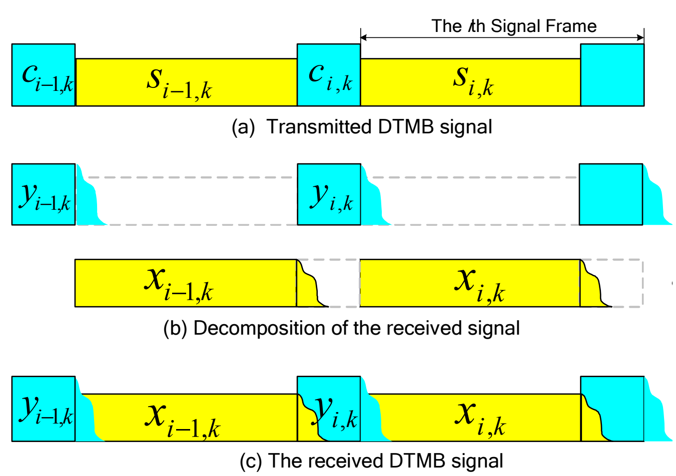

Fig. 3. Transmission of DTMB signal over multi-path channel

$$\hat{y}_{i,k} = \begin{cases} r_{i,k} - x_{i-1,k+N}, & 0 \le k < L-1 \\ r_{i,k}, & L \le k < M \\ r_{i,k} - x_{i,k-M}, & M \le k < M+L-1 \end{cases} \quad (7)$$

$\{x_{i-1,k}\}_{k=N}^{N+L-2}$ can be obtained by the estimation of FB and CIR of previous signal frame easily. Since the relationship between the non-overlapping part $\{y_{i,k}\}_{k=0}^{M-1}$ and the overlapping part $\{y_{i,k}\}_{k=M}^{M+L-2}$ and the power of $\{y_{i,k}\}_{k=0}^{M+L-2}$ is twice as much as that of $\{x_{i,k}\}_{k=0}^{N+L-2}$, we can estimate $\{y_{i,k}\}_{k=M}^{M+L-2}$ and obtain more precise CIR from $\{\hat{y}_{i,k}\}_{k=M}^{M+L-2}$ and $\{r_{i,k}\}_{k=0}^{M-1}$, which is the starting point of the iterative reconstruction algorithm for PN sequence and channel estimation in the following section.

III. Iterative Channel Estimation Method

In this section, the proposed iterative channel estimation algorithm will be investigated. Besides being used as GI, the padded PN sequence can also be used for synchronization and channel estimation. For simplicity, we assume that perfect timing and frequency synchronization can be achieved at the receiver. The proposed method reconstructs the PN sequence and performs channel estimation in an iterative manner. The details of the proposed channel estimation algorithm for the *i*th received signal frame are summarized as the following steps:

(1) The reconstruction is performed frame by frame, when we deal with the *i*th signal frame, the estimations of the channel impulse response $\{\hat{h}_{i\text{-}2,k}\}_{k=0}^{L-1}$, $\{\hat{h}_{i\text{-}1,k}\}_{k=0}^{L-1}$ and $\{\hat{x}_{i\text{-}1,k}\}_{k=0}^{N+L-1}$ for the previous two signal frames are already known. Assuming the channel impulse response (CIR) is quasi-static. Based on $\{\hat{h}_{i\text{-}2,k}\}_{k=0}^{L-1}$ and $\{\hat{h}_{i\text{-}1,k}\}_{k=0}^{L-1}$, via a linear interpolation for both real and imaginary parts, $\{\hat{h}_{i,k}^{iter=0}\}_{k=0}^{L-1}$ can be estimated [10]. In principle, high-order interpolation method can be utilized.

(2) Since $\{c_{i-1,k}\}_{k=0}^{M-1}$ and $\{c_{i,k}\}_{k=0}^{M-1}$ associated with the signal frames are known after the synchronization, the linear convolution outputs $\{y_{i-1,k}\}_{k=0}^{M+L-2}$ and $\{y_{i,k}\}_{k=0}^{M+L-2}$ can be calculated with N_1-point FFT respectively, where $N_1 \ge M+L-1$.

(3) Subtract $\{y_{i-1,k}\}_{k=0}^{M+L-2}$ and $\{y_{i,k}\}_{k=0}^{M+L-2}$ from the (*i*-1)th received signal frame $\{r_{i-1,k}\}_{k=0}^{M+L-2}$, and obtain an estimate of $\{\hat{x}_{i-1,k}\}_{k=0}^{N+L-2}$:

$$\hat{x}_{i-1,k} = \begin{cases} r_{i-1,k+M} - \hat{y}_{i-1,k+M}, & 0 \le k < L-1 \\ r_{i-1,k+M}, & L-1 \le k < N \\ r_{i-1,k+M} - \hat{y}_{i,k-N}, & N \le k < N+L-1 \end{cases} \quad (8)$$

(4) After interference being successfully removed, $\{\hat{x}_{i-1,k}\}_{k=0}^{N+L-2}$ is then equivalent to ZP-OFDM signal. From calculation complexity point of view, the ZP-OFDM-OLA equalization method [11] is used here. Decision feedback method is used on $\{\hat{x}_{i-1,k}\}_{k=0}^{N+L-2}$ to remove the noise components from $\{\hat{x}_{i-1,k}\}_{k=0}^{N+L-2}$ with given $\{z_{i-1,k}\}_{k=0}^{M+L-2}$. The corresponding method is described in Section III-A. Set the iteration index J to zero and go to step (5).

(5) Calculate the linear convolution of $\{\hat{h}_{i,k}^{iter=J}\}_{k=0}^{L-1}$ and $\{C_{i,k}\}_{k=0}^{M}$ with a N_1-point FFT, and the output is $\left\{\hat{y}_{i-1,k}^{iter=J}\right\}_{k=0}^{M+L-2}$.

(6) Reconstruct the PN sequence at the receiver with $\{r_{i-1,k}\}_{k=0}^{M+L-2}$, $\{\hat{y}_{i-1,k}^{iter=J}\}_{k=0}^{M+L-2}$ in step (5), and $\{z_{i-1,k}\}_{k=0}^{M+L-2}$ in step (4) as shown in Fig.4. The non-overlapping part is obtained by the received signal and the overlapping part is achieved by estimation signal. An estimation of the received PN sequence $\{u_{i,k}^{iter=J}\}_{k=0}^{M+L-2}$ can be obtained as:

$$u_{i,k}^{iter=J} = \begin{cases} r_{i,k} - z_{i-1,k+M}, & 0 \le k < L-1 \\ r_{i,k}, & L-1 \le k < M \\ \hat{y}_{i,k}^{iter=J}, & M \le k < M+L-1 \end{cases} \quad (9)$$

(7) From $\{u_{i,k}^{iter=J}\}_{k=0}^{M+L-2}$, a more accurate CIR $\{\hat{h}_{i,k}^{iter=J+1}\}_{k=0}^{L-1}$ is obtained. The corresponding channel estimation method is described in Section III-B. If J is equal to the predefined maximum number of iterations J_0, $\{\hat{h}_{i,k}^{iter=J_0}\}_{k=0}^{L-1}$ becomes the final estimation of $\{\hat{h}_{i,k}\}_{k=0}^{L-1}$ (i.e. $\{h_{i,k}\}_{k=0}^{L-1}$), move on to the (*i*+1)th signal frame and start again from step (1). Otherwise, set *J*=*J*+1 and go back to step (5).

A. Decision Feedback of $\{\hat{x}_{i-1,k}\}_{k=0}^{N+L-2}$

In step (4), after removing the impact of the head and tail parts from by the PN sequence in (*i*-1)th and *i*th signal frame respectively, $\{\hat{x}_{i-1,k}\}_{k=0}^{N+L-2}$ is still contaminated by the noise and residual ISI, the decision feedback method is used to reduce the noise components as follows:

(1) Calculate the channel estimation $\{\tilde{h}_{i,k}\}_{k=0}^{L-1}$ for $\{\hat{x}_{i-1,k}\}_{k=0}^{N+L-2}$. For simplicity, the average of $\{\hat{h}_{i\text{-}1,k}\}_{k=0}^{L-1}$ and $\{\hat{h}_{i,k}^{iter=0}\}_{k=0}^{L-1}$ is used here:

$$\tilde{h}_{i,k} = (\hat{h}_{i\text{-}1,k} + \hat{h}_{i,k}^{iter=0})/2, \quad 0 \le l < L \quad (10)$$

(2) $\{\hat{x}_{i-1,k}\}_{k=0}^{N+L-2}$ can be equalized by the one-tap frequency-domain equalizer through N-point FFT and IFFT operations with ZP-OFDM-OLA equalization method. Two commonly used equalizer, namely, the zero-forcing (ZF) and linear minimum mean-square error (LMMSE) can be constructed as follows.

ZF equalizer:

$$\hat{S}_{i-1,k} = \frac{X_{i-1,k}}{\tilde{H}_{i-1,k}}, \quad 0 \le k < N \quad (11)$$

where $X_{i-1,k}$ and $\tilde{H}_{i-1,k}$ denote the discrete Fourier transform (DFT) of $\{\hat{x}_{i-1,k}\}_{k=0}^{N+L-2}$ and $\{\tilde{h}_{i,k}\}_{k=0}^{L-1}$ respectively.

LMMSE equalizer:

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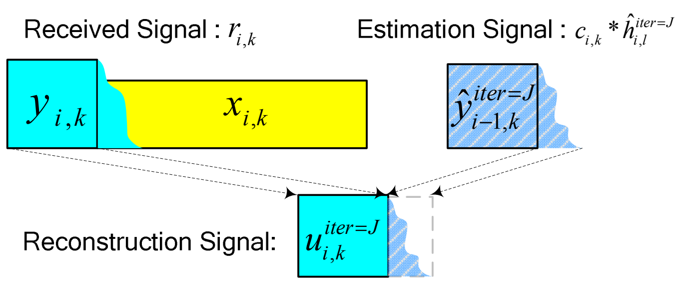

Fig. 4. Reconstruction of the received PN sequence

$$\hat{S}_{i-1,k} = \frac{\tilde{H}_{i-1,k}^{H} \odot X_{i-1,k}}{\tilde{H}_{i-1,k}^{H} \odot \tilde{H}_{i-1,k} + \sigma_w^2 / \sigma_S^2}, \quad 0 \le k < N \quad (12)$$

where the symbol $\odot$ and the superscript H denote the element-wise multiplication operation and the complex conjugate operation respectively.

(3) Then make a decision feedback on $\{\hat{S}_{i-1,k}\}_{k=0}^{N-1}$. After making decisions on $\{\hat{S}_{i-1,k}\}_{k=0}^{N-1}$, the decisions are then converted back to time-domain with the result of $\{\tilde{s}_{i-1,k}\}_{k=0}^{N-1}$.

(4) The final output $\{z_{i-1,k}\}_{k=0}^{M+L-2}$ is the linear convolution between the last M element of $\{\tilde{s}_{i-1,k}\}_{k=0}^{N-1}$ denote as $\{\tilde{s}_{i-1,k}\}_{k=N-M}^{N-1}$ and $\{\tilde{h}_{i,k}\}_{k=0}^{L-1}$, which can also be calculated by multiplication operation after both being converted to frequency domain with N_1 points FFT respectively.

B. Channel estimation methods

In step (7), $\{u_{i,k}^{iter=J}\}_{k=0}^{M+L-2}$ is the estimation of the linear convolution result of $\{c_{i,k}\}_{k=0}^{M-1}$ and $\{h_{i,k}\}_{k=0}^{L-1}$, we denote $\{C_{i,k}\}_{k=0}^{N_1-1}$ and $\{Y_{i,k}^{iter=J}\}_{k=0}^{N_1-1}$ are the N_1-point FFT of $\{c_{i,k}\}_{k=0}^{M-1}$ and $\{u_{i,k}^{iter=J}\}_{k=0}^{M+L-2}$ being zero-padded to the length of N_1 respectively. The CIR estimation on $\{\hat{h}_{i,k}^{iter=J+1}\}_{k=0}^{N_1-1}$ is given by:

$$\hat{h}_{i,k}^{iter=J+1} = IFFT\left(\frac{Y_{i,k}^{iter=J}}{C_{i,k}}\right), \quad 0 \le k < N_1 - 1 \quad (14)$$

Since the length of the CIR is known to be smaller than M, all the CIR values $\{\hat{h}_{i,k}^{iter=J+1}\}_{k=L}^{N_1-1}$ are set to zeros. To reduce the effect of noise, a threshold h_{th} is used to select the significant taps and the absolute values of $\{\hat{h}_{i,k}^{iter=J+1}\}_{k=0}^{L-1}$ below the threshold are set to zero [7]. Finally, a new CIR is obtained and the result $\{\hat{h}_{i,k}^{iter=J+1}\}_{k=0}^{L-1}$ will be used for the next iteration.

Taking DTMB system as an example with parameters of $N =$ 3780, M = 420 and $N_1 = 2048$, the computational complexity comparison in term of IFFT/FFT resources between the proposed scheme and the conventional method [6] is shown in Table I. As shown in this table, the proposed channel estimation scheme has a much lower complexity as the conventional one.

An N-point FFT/IFFT operation computation cost is supposed to be as $O(N \log_2 N)$. The complexity of proposed method is only 24% and 22 % of that in method [6] when J=0 and J=1, respectively, and the algorithm can be applied to practical DTMB receiver directly. Other than implementation cost saving, the satisfactory performance will be demonstrated in the following.

IV. Simulation Results

In this section, simulations are performed to demonstrate the performance iterative channel estimation algorithm. The main simulation parameters based on Chinese DTMB standard [12] are shown in Table II. Two typical broadcast channels model are chosen: Fixed Reception F1 version of DVB-T channel model [1] and China DTV Test 8th channel model (CDT8). The power delay profiles of CDT8 are shown in Table III and it consists of a long delay and large amplitude echo. The maximal Doppler shift is 10Hz so that the channel gain remains almost the same during one signal frame. It is assumed that each path is subject to an independent Raleigh fading and the overall path losses are normalized.

The symbol error rate (SER) as a function of signal-to-noise ratio (SNR) in dB of a 16QAM modulated DTMB system performance under DVB-T F1 channel is presented in Fig.5. The SNR performance based on the iterative estimated CIR is compared with the lower bound in which the SER is calculated by ideal channel estimation. The corresponding curve using the method in [6] with 2 iterations is also provided for reference. The numbers of iteration are set to 1, and 2, respectively for our method. It can be seen from figure that the proposed algorithm offers improvement of the SNR performance after the first iteration. Two iterations can bring much better performance, about 2dB SNR improvement at SER of 10^{-1}(normal operating point). From Fig. 5, the reconstruction iterative method with two iterations improves the performance of SNR at about the same level as that of the method in [6] with the same iteration number, and is only about 2.5dB degradation compared to the lower bound.

Fig.6 shows the SNR performance of DTMB system over CDT8 channel with the same modulation scheme. For a SER of 10^{-1}, the SNR degradation of the performance based on reconstruction method with two iterations is about 2 dB. The performance of the proposed algorithm is a little better than method in [6] as the same. Since the slight taps of the

TABLE I
Computation Complexity Comparison between the Conventional and Proposed Method

Operation	Method in [6]	Proposed method
I/FFT2048	4(J+1)	3J+4
I/FFT3780	2	2
I/FFT8192	3(J+1)	0

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TABLE II
MAIN SIMULATION PARAMETERS

Symbol Rate f_s	7.56 MSPS
Signal Constellation	16QAM
PN Sequence Length M	420
FFT Size N	3780
FFT Size N_1	2048
Maximal Doppler Shift	10Hz
Sub-carrier Spacing	2kHz

TABLE III
POWER DELAY PROFILES OF CDT8 CHANNEL

Tap	Delay(us)
1	1.85
2	0
3	1.95
4	3.57
5	7.54
6	31.9

estimation of CIR are set to zeros due to thresholding, the proposed algorithm has an error floor shown in both Fig.5 and Fig.6 at high SNR region.

V. CONCLUSION

In this paper, an iterative channel estimation algorithm is proposed based on PN sequence reconstruction. The overlapping part of the FH is estimated by the non-overlapping part in the received signal, CIR estimation updated and the residual ISI removed iteratively. Analysis shows that the proposed method has a much lower complexity compared with that of the conventional method. Through the simulations, it is shown that the SNR degradation of the proposed channel estimator is a slightly better than conventional one.

The proposed iterative algorithm is expected to give an alternative channel estimation method for the TDS-OFDM system with quite satisfactory performance at much lower complexity and this method, in principle, can also be applied to the single-carrier working mode of DTMB system with small modification.

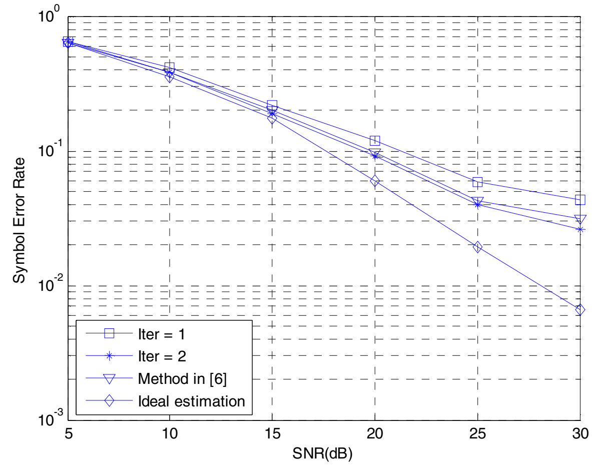

Fig. 5. SER performance of DTMB system over DVB-T F1 channel

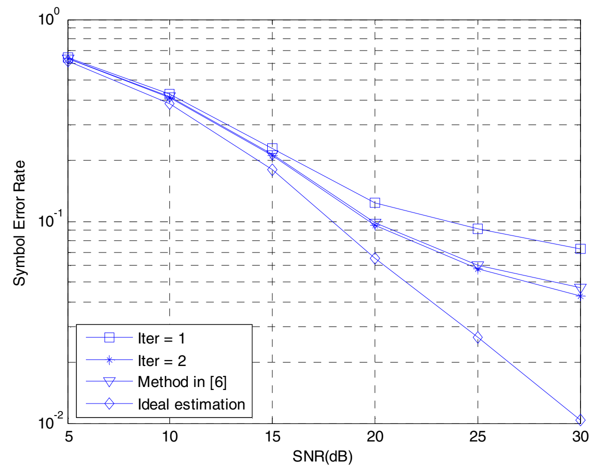

Fig. 6. SER performance of DTMB system over CDT8 channel

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