Novel Channel Estimation Method Based on PN Sequence Reconstruction for Chinese DTTB System

Fang Yang, Jintao Wang, Jun Wang, Jian Song, and Zhixing Yang

Abstract — *Digital television/terrestrial multimedia broadcasting (DTMB), announced as the Chinese digital television terrestrial broadcasting (DTTB) standard, uses pseudo-noise (PN) sequence as guard interval (GI) and the training sequence for both multi- and single-carrier block transmissions. The conventional channel estimation methods can be performed through either the subtraction of the PN sequence or the cancellation of residual inter-symbol interference (ISI). In this paper, a novel iterative method is introduced to estimate the overlapping part of PN sequence, meanwhile the channel impulse response (CIR) estimation can be obtained and updated via the reconstruction of the PN sequence in an iterative manner. Simulations show that the proposed algorithm can effectively reduce the ISI and improve the symbol error rata (SER) of DTMB system over slow-varying broadcasting multipath channel even with long delay spread. In practice, the proposed algorithm can be applied to the DTMB receiver-chip directly with low-complexity cost*[1].

Index Terms — Channel estimation, DTMB, time-domain synchronous OFDM (TDS-OFDM), PN sequence reconstruction.

I. INTRODUCTION

Orthogonal frequency–division multiplexing (OFDM) is an attractive transmission technique to combat the inter-symbol interference (ISI) over the frequency-selective fading channels with high bandwidth efficiency and has been extensively adopted into wide-band 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 synchronous OFDM (TDS-OFDM) uses pseudo-noise (PN) sequences padding as the guard interval (GI) as well as the training sequence, which makes the synchronization faster than that of CP-OFDM systems [3]. The separation of training sequence and transmitted data makes the time/frequency synchronization independent on the specific modulation type and increases the modularity of the system design. Digital television/terrestrial multimedia broadcasting (DTMB), announced as the Chinese digital television terrestrial broadcasting (DTTB) standard, adopts TDS-OFDM as the baseline modulation technology [4].

TDS-OFDM systems, where OFDM symbols padded with PN sequences are not orthogonal any more, hence, cannot directly utilize channel estimation algorithms derived in CP-OFDM systems. Ideally, 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. However, since subtracting the PN sequences and the accurate estimation of CIR are inter-dependent in TDS-OFDM system, it is quite difficult to achieve the ideal performance in practice, therefore CIR estimation is a very challenging topic and some iterative algorithms were proposed. In [5] and [6], 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 complexity grows linearly with the square of the PN length. Furthermore, 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 [7]. Another way to solve this problem is to process the padded PN sequence and the data part separately [8]. Based on discrete Fourier transformation (DFT) of the received and local PN sequences, the channel estimation is performed in frequency domain. However, the algorithms above are very complicated and implementation cost is generally high. Literature [9] addresses this issue for a specific TDS-OFDM system padded by fixed PN sequence, where the "virtual" frame is introduced to avoid the inter-block interference (IBI) from the previous signal frame.

Several reconstruction methods are proposed for other communication systems, such as the residual ISI cancellation (RISIC) algorithms in the CP-OFDM system [10] and cyclic prefix reconstruction method for coded single-carrier system with frequency-domain equalization (SC-FDE) [11]. In this paper, we present a novel iterative channel estimation algorithm based on PN sequence reconstruction with low complexity. Different from the method in [8] that subtracts the

[1] This work was supported in part by "Multistandard integrated network convergence for global mobile and broadcast technologies" (MING-T, FP6 STREP Contract Nr.045461). Part of this work has been presented at the workshop on Digital Television and Mobile Multimedia Broadcasting of ICC 2008, Beijing, China, May 2008.

All the authors are with the Department of Electronic Engineering as well as the Tsinghua National Laboratory of Information Science and Technology (TNList), Tsinghua University, Beijing 100084, P.R. China (e-mail: yangfang01@mails.tsinghua.edu.cn).

Contributed Paper
Manuscript received September 22, 2008

PN sequence from received signal and reduces the residual ISI iteratively, the proposed algorithm reconstructs the overlapping part and uses the reconstruction signal to estimate CIR more accurately. This idea is derived 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. The proposed PN sequence reconstruction method is investigated in Section III. In section IV, the proposed iterative channel estimation algorithm is discussed in detail. Simulation results are presented in section V to verify the effectiveness of the proposed algorithm, and conclusions are drawn in Section VI.

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, an m-sequence, and a postamble. Here the m-sequence can be generated with a Fibonacci-type linear feedback shift register (LFSR) and both preamble and postamble are the cyclic extensions 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 P within each FB, $P=1$ for single-carrier mode and $P=3780$ for multi-carrier mode.

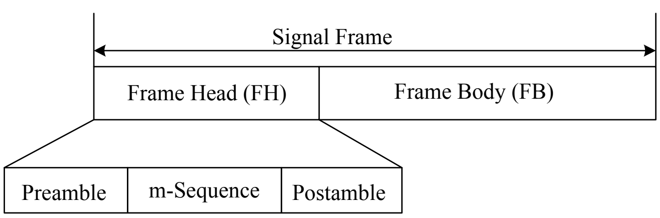

Fig. 1. Frame structure of the DTMB signal.

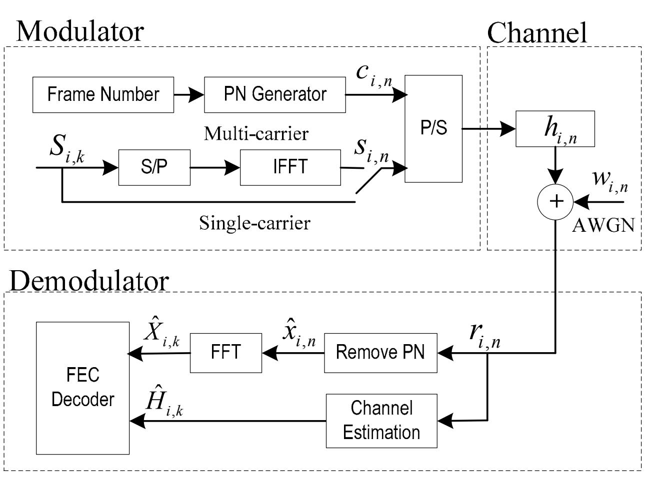

Fig. 2. Block diagram of the DTMB signal.

For multi-carrier mode, the ith transmitted symbol $\{S_{i,k}\}_{k=0}^{N-1}$ after inverse fast Fourier transform (IFFT) can be expressed as

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

where N denotes the FB length. For single-carrier mode, the ith transmitted symbol is defined and transmitted in time domain. 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 FH $\{c_{i,n}\}_{n=0}^{M-1}$ of length M is inserted between FBs and the power of FH is two times as much as that of FB. Then, the FHs and FBs form the DTMB signal frames and the transmitted frame is shown in a Fig. 3 (a).

The propagation channel is modeled as a quasi-static Lth-order FIR filter, where CIR $\{h_{i,n}\}_{n=0}^{L-1}$ does not vary within ith signal frame period, i.e., the coherence time of channel is much longer than the duration of one signal frame. With enough length of GI is assumed to mitigate the multipath effect ($M \ge L$), the received signal will be the linear correlation between the transmitted signal and the CIR as shown in Fig. 3(b) and can be represented as

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

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

where $\otimes$ denotes linear convolution operator, while $\{x_{i,n}\}_{n=0}^{N+L-2}$ and $\{y_{i,n}\}_{n=0}^{M+L-2}$ represent the linear convolution of FB and FH with the CIR, respectively. As shown in Fig. 3(c), the shadow areas represent the tails overlapping in time domain after propagation over a multipath environment of the FH or FB. Specifically, in a received signal frame, the FB is corrupted by the tail of the FH sequence in the same signal frame, while the FH suffers the interference from the tail of the FB in the previous signal frame. From (2) and (3), the expression of the ith received signal frame $\{r_{i,n}\}_{n=0}^{M+N+L-2}$ is given by

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

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

At the receiver side, if $\{y_{i,n}\}_{n=0}^{M+L-2}$ can be extracted from the received signal, the channel estimation $\{\hat{h}_{i,n}\}_{n=0}^{L-1}$ will be obtained simply and the rest signal will be the received FB which be free from FH interference. Denoting $\{X_{i,k}\}_{k=0}^{N-1}$ as the N-point DFT of $\{x_{i,n}\}_{n=0}^{N+L-2}$ after cyclic reconstruction, while $\{\hat{H}_{i,k}\}_{k=0}^{N-1}$ represents the estimated channel frequency response (CFR) which is the DFT of $\{\hat{h}_{i,n}\}_{n=0}^{L-1}$. Then, $\{X_{i,k}\}_{k=0}^{N-1}$ and $\{\hat{H}_{i,k}\}_{k=0}^{N-1}$ are applied to the forward error correction (FEC) decoder.

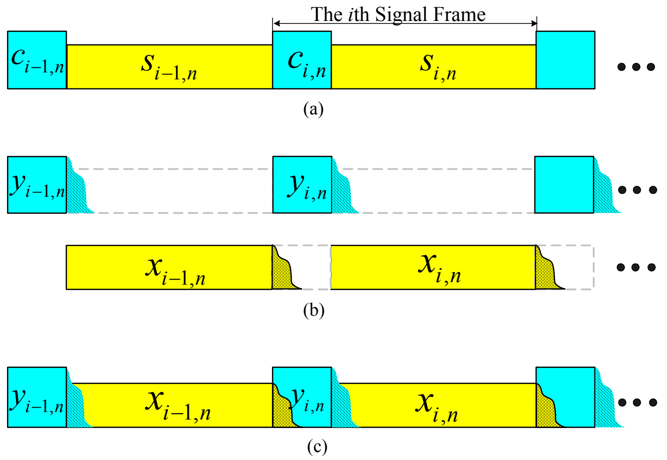

Fig. 3. The transmitted and received signals over multipath channel. (a) The transmitted signal frames; (b) Decomposition of the received signal frames; (c) The received signal frames.

III. PN Sequence Reconstruction Method

Without considering the noise, we can express the estimation of $\{y_{i,n}\}_{n=0}^{M+L-2}$ as

$$\hat{y}_{i,n}=\begin{cases} r_{i,n}-x_{i-1,n+N}, & 0\le n<L-1\\ r_{i,n}, & L-1\le n<M\\ r_{i,n}-x_{i,n-M}, & M\le n<M+L-1. \end{cases} \quad (5)$$

Since $\{x_{i-1,n}\}_{n=N}^{N+L-2}$ can be removed based on the estimation of FB and CIR of previous signal frame easily, then $\{y_{i,n}\}_{n=0}^{M-1}$ is obtained. Because of the relationship between the non-overlapping part $\{y_{i,n}\}_{n=0}^{M-1}$ and the overlapping part $\{y_{i,n}\}_{n=M}^{M+L-2}$, we can estimate $\{y_{i,n}\}_{n=M}^{M+L-2}$ and obtain more accurate CIR from $\{y_{i,n}\}_{n=0}^{M-1}$ and $\{\hat{h}_{i,n}^{(J)}\}_{n=0}^{L-1}$, where J denotes the iteration index, which is the starting point of the iterative reconstruction algorithm for PN sequence and channel estimation in the following section.

Hereafter, boldface letters represent matrices or vectors; $(\cdot)^T$, $(\cdot)^H$, and $(\cdot)^*$ denote the transpose, Hermitian, and complex conjugate operator, respectively; $\|\cdot\|$ denotes the spectrum norm of a matrix; $\mathbf{I}_K$ denotes the $K\times K$ identity matrix; $\mathrm{diag}(\mathbf{x})$ represents the diagonal matrix with the vector $\mathbf{x}$ on its diagonal.

To facilitate our explanation, $\mathbf{h}_i^{(J)}=[\hat{h}_{i,0}^{(J)},\cdots,\hat{h}_{i,L-1}^{(J)},\mathbf{0}_{1\times(K-L)}]^T$ denotes a $K\times 1$ vector which contains the estimated CIR of the Jth iteration cycle being zero-padded to length of K ($K\ge M+L-1$), $\mathbf{C}_i=\mathrm{diag}(C_{i,0},C_{i,1},\cdots,C_{i,K-1})$ represents the matrix with K-point FFT of $\{c_{i,n}\}_{n=0}^{M-1}$ on the diagonal. The PN reconstruction and CIR update algorithm proceeds as follows.

(1) Signal estimation. The estimation of received PN sequence of Jth iteration cycle based on estimated CIR $\mathbf{h}_i^{(J)}$ and local PN sequence $\mathbf{C}_i$ is given by

$$\mathbf{y}_i^{(J)}=\mathbf{F}^H\mathbf{C}\mathbf{F}\mathbf{h}_i^{(J)} \quad (6)$$

where $\mathbf{F}$ denotes K-point FFT matrix whose p, qth element is equal to $(1/\sqrt{K})\exp(-j2\pi pq/K)$, while $\mathbf{F}^H$ is the Hermitian of $\mathbf{F}$ and represents the IFFT matrix.

(2) PN sequence reconstruction. As shown in Fig. 4, the reconstruction of PN is the combination of the received non-overlapping part and the estimated overlapping part by forming

$$\mathbf{u}_i^{(J)}=\mathbf{P}\mathbf{y}_i+(\mathbf{I}_K-\mathbf{P})\mathbf{y}_i^{(J)} \quad (7)$$

where $\mathbf{y}_i$ is the vector of $\{y_{i,n}\}_{n=0}^{M+L-2}$ including $\mathbf{0}_{1\times(K-M-L+1)}$ padding to length of K , while $\mathbf{P}=\mathrm{diag}(\mathbf{I}_M,\mathbf{0}_{(K-M)\times(K-M)})$ represents a diagonal matrix to select the first M elements. Likewise, $(\mathbf{I}_K-\mathbf{P})$ is used to choose the last $(K-M)$ elements.

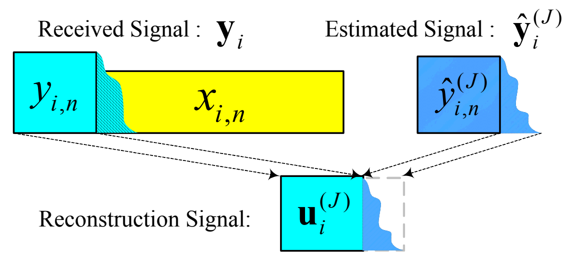

Fig. 4. Reconstruction of the PN sequence.

(3) CIR update. The CIR estimation can be obtained through the IFFT/FFT operation with least squared (LS) criterion. Afterwards, the result is passed through a rectangular window which set the last $(K-M)$ elements to zeros and the estimated CIR can be expressed as

$$\mathbf{h}_i^{(J+1)}=\mathbf{P}\mathbf{F}^H\mathbf{C}^{-1}\mathbf{F}\mathbf{u}_i^{(J)}\,. \quad (8)$$

From (6) to (8), we have

$$\begin{aligned}\mathbf{y}_i^{(J+1)}&=\mathbf{F}^H\mathbf{C}\mathbf{F}\mathbf{h}_i^{(J+1)}\\&=\mathbf{T}\left(\mathbf{P}\mathbf{y}_i+(\mathbf{I}_K-\mathbf{P})\mathbf{y}_i^{(J)}\right)\end{aligned} \quad (9)$$

where $\mathbf{T}=\mathbf{F}^H\mathbf{C}\mathbf{F}\mathbf{P}\mathbf{F}^H\mathbf{C}^{-1}\mathbf{F}$. Another important fact is that if $\mathbf{y}_i$ passed through the whole operation discussed above, the result is still $\mathbf{y}_i$, i.e.

$$\mathbf{y}_i=\mathbf{T}\mathbf{y}_i\,. \quad (10)$$

Therefore, the estimation error of $\mathbf{y}_i$ under the (J+1)th iteration cycle is given by

$$\begin{aligned}\mathbf{y}_i^{(J+1)}-\mathbf{y}_i&=\mathbf{T}\left(\mathbf{P}\mathbf{y}_i+(\mathbf{I}_K-\mathbf{P})\mathbf{y}_i^{(J)}\right)-\mathbf{T}\mathbf{y}_i\\&=\mathbf{T}(\mathbf{I}_K-\mathbf{P})(\mathbf{y}_i^{(J)}-\mathbf{y}_i).\end{aligned} \quad (11)$$

For the purpose of keeping the algorithm converge, the upper bound of $\|\mathbf{T}(\mathbf{I}_K-\mathbf{P})\|$ is less than 1 must be satisfied. However, it is difficult to obtain the tight upper bound of $\|\mathbf{T}(\mathbf{I}_K-\mathbf{P})\|$ and a loose upper bound is given by

$$\begin{aligned}&\|\mathbf{T}(\mathbf{I}_K-\mathbf{P})\|\\&\le\|\mathbf{F}^H\|\cdot\|\mathbf{C}\|\cdot\|\mathbf{F}\|\cdot\|\mathbf{P}\|\cdot\|\mathbf{F}^H\|\cdot\|\mathbf{C}^{-1}\|\cdot\|\mathbf{F}\|\cdot\|\mathbf{I}_K-\mathbf{P}\|\\&=\max_k|C_{i,k}|/\min_k|C_{i,k}|.\end{aligned} \quad (12)$$

It is clear that training sequence staying in flat in frequency domain is very important and an accurate initial CIR can speed up the convergence. In practice, the received PN sequence is corrupted by the residual ISI from previous signal frame and the impact of AWGN, some interference cancellation methods will be disused in detail in the following section.

IV. 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 ith received signal frame are summarized as the following steps.

(1) The reconstruction is performed frame by frame, when dealing with the ith signal frame, the estimations of the channel impulse responses $\{\hat{h}_{i-2,n}\}_{n=0}^{L-1}$, $\{\hat{h}_{i-1,n}\}_{n=0}^{L-1}$ for the previous two signal frames and $\{\hat{x}_{i-1,n}\}_{n=0}^{N+L-2}$ are already known. Based on $\{\hat{h}_{i-2,n}\}_{n=0}^{L-1}$ and $\{\hat{h}_{i-1,n}\}_{n=0}^{L-1}$, via a linear interpolation for both real and imaginary parts, $\{\hat{h}_{i,n}^{(0)}\}_{n=0}^{L-1}$ can be estimated [12]. In principle, high-order interpolation method can be utilized.

(2) Since the transmitted FHs associated with the signal frames are known after the synchronization, the linear convolution outputs $\{y_{i-1,n}\}_{n=0}^{M+L-2}$ and $\{\hat{y}_{i,n}^{(0)}\}_{n=0}^{M+L-2}$ can be calculated as

$$\hat{y}_{i-1,n} = \mathrm{IFFT}_K\left[\mathrm{FFT}_K(\hat{h}_{i-1,n}) \odot \mathrm{FFT}_K(c_{i-1,n})\right] \quad (13)$$

$$\hat{y}_{i,n}^{(0)} = \mathrm{IFFT}_K\left[\mathrm{FFT}_K(\hat{h}_{i,n}^{(0)}) \odot \mathrm{FFT}_K(c_{i,n})\right] \quad (14)$$

where $\mathrm{IFFT}_K[\cdot]$ / $\mathrm{FFT}_K[\cdot]$ denote the K-point IFFT/FFT operations respectively, while $\odot$ denotes the element-wise multiplication operation.

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

$$\hat{x}_{i-1,n} = \begin{cases} r_{i-1,n+M} - \hat{y}_{i-1,n+M}, & 0 \le n < L-1 \\ r_{i-1,n+M}, & L-1 \le n < N \\ r_{i-1,n+M} - \hat{y}_{i,n-N}^{(0)}, & N \le n < N+L-1. \end{cases} \quad (15)$$

(4) After interference from FHs being successfully removed, $\{\hat{x}_{i-1,n}\}_{n=0}^{N+L-2}$ is then equivalent to ZP-OFDM signal. From calculation complexity point of view, the ZP-OFDM overlap-add (ZP-OFDM-OLA) equalization method [13] is used here. Decision feedback method is adopted to eliminate the noise components of $\{\hat{x}_{i-1,n}\}_{n=0}^{N+L-2}$ with given $\{z_{i-1,n}\}_{n=0}^{N+L-2}$. The noise elimination method is partial decision which is a simple method to suppress the noise and will be introduced in Section IV-A. Set the iteration index J to zero.

(5) Calculate the estimation signal of $\{y_{i,n}\}_{n=0}^{M+L-2}$ under the Jth iteration cycle and can be expressed as

$$\hat{y}_{i,n}^{(J)} = \mathrm{IFFT}_K\left[\mathrm{FFT}_K(\hat{h}_{i,n}^{(J)}) \odot \mathrm{FFT}_K(c_{i,n})\right] \quad (16)$$

(6) Reconstruct the PN sequence with received signal $\{r_{i,n}\}_{n=0}^{M+N+L-2}$, $\{\hat{y}_{i,n}^{(J)}\}_{n=0}^{M+L-2}$ in Step 5, and $\{z_{i-1,n}\}_{n=0}^{M+L-2}$ in Step 4. The non-overlapping part is obtained by the received signal and the overlapping part is achieved by the estimated signal. Then, the reconstruction of the received PN sequence $\{u_{i,n}^{(J)}\}_{n=0}^{M+L-2}$ is given by:

$$u_{i,n}^{(J)} = \begin{cases} r_{i,n} - z_{i-1,n+N}, & 0 \le n < L-1 \\ r_{i,n}, & L-1 \le n < M \\ \hat{y}_{i,n}^{(J)}, & M \le n < M+L-1. \end{cases} \quad (17)$$

(7) From $\{u_{i,n}^{(J)}\}_{n=0}^{M+L-2}$, a more accurate CIR $\{\hat{h}_{i,n}^{(J+1)}\}_{n=0}^{L-1}$ is obtained. The corresponding channel estimation method is discussed in Section IV-B. If J is equal to the predefined maximum number of iterations J_0, $\{\hat{h}_{i,n}^{(J_0)}\}_{n=0}^{L-1}$ becomes the final estimation of $\{h_{i,n}\}_{n=0}^{L-1}$ (i.e. $\{\hat{h}_{i,n}\}_{n=0}^{L-1}$), 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 $\{\hat{x}_{i-1,n}\}_{n=0}^{N+L-2}$

In Step 4, after removing the impact of the head and tail parts from the FHs in (i-1)th and ith signal frame respectively, $\{\hat{x}_{i-1,n}\}_{n=0}^{N+L-2}$ 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 $\{\tilde{h}_{i-1,n}\}_{n=0}^{L-1}$ for $\{\hat{x}_{i-1,n}\}_{n=0}^{N+L-2}$. For the sake of simplicity, the average of $\{\hat{h}_{i-1,n}\}_{n=0}^{L-1}$ and $\{\hat{h}_{i,n}^{(0)}\}_{n=0}^{L-1}$ is used here

$$\tilde{h}_{i-1,n} = (\hat{h}_{i-1,n} + \hat{h}_{i,n}^{(0)})/2, \quad 0 \le n < L. \quad (18)$$

(2) $\{\hat{x}_{i-1,n}\}_{n=0}^{N+L-2}$ can be equalized by the one-tap frequency-domain equalizer through FFT/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:

$$\hat{S}_{i-1,k} = \frac{\hat{X}_{i-1,k}}{\tilde{H}_{i-1,k}}, \quad 0 \le k < N \quad (19)$$

where $\{\hat{X}_{i-1,k}\}_{k=0}^{N-1}$ and $\{\tilde{H}_{i-1,k}\}_{k=0}^{N-1}$ denote the N-point DFT of $\{\hat{x}_{i-1,n}\}_{n=0}^{N+L-2}$ after cyclic reconstruction and $\{\tilde{h}_{i,n}\}_{n=0}^{L-1}$, respectively.

LMMSE equalizer:

$$\hat{S}_{i-1,k} = \frac{\tilde{H}_{i-1,k}^{*} \cdot X_{i-1,k}}{\tilde{H}_{i-1,k}^{*} \cdot \tilde{H}_{i-1,k} + \sigma_w^2/\sigma_S^2}, \quad 0 \le k < N \quad (20)$$

(3) For multi-carrier mode, making a partial decision on $\{\hat{S}_{i-1,k}\}_{k=0}^{N-1}$. The partial decision is an effective method to eliminate the noise components and reduce the probability of error decision. Due to the impact the AWGN, the received signal falls within a neighborhood of the transmitted symbol. Therefore, make a decision on $\{\hat{S}_{i-1,k}\}_{k=0}^{N-1}$ can suppress the noise effect. However, the signals falling at the middle region of constellation mapping points should be processed carefully to avoid error decision. The partial decision is used to divide $\{\hat{S}_{i-1,k}\}_{k=0}^{N-1}$ into two regions: decision and non-decision regions as shown in Fig.5 where the shadow areas represent the decision regions and the other areas are the non-decision regions. For BPSK modulation with points of (-1, 0) and (1, 0) as an example, the partial decision is calculated as

$$\tilde{S}_{i,k} = \begin{cases} 1, & \hat{S}_{i,k} > \varepsilon \\ -1, & \hat{S}_{i,k} < -\varepsilon \\ \hat{S}_{i,k}, & -\varepsilon < \hat{S}_{i,k} < \varepsilon \end{cases} \tag{21}$$

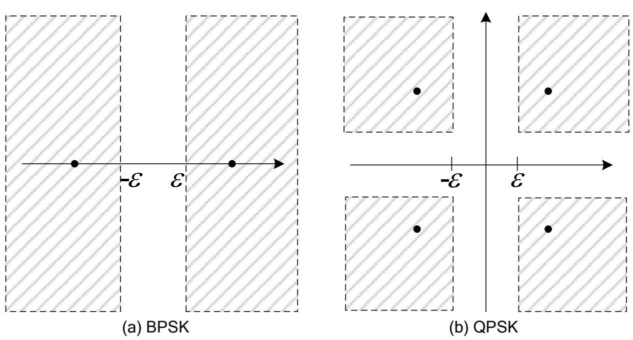

Fig. 5. Partial decision region for BPSK/QPSK.

where ε and $-\varepsilon$ are the decision thresholds to be chosen appropriately. For QPSK/m-QAM modulation, the partial decision method can be obtained straightforward as an extension of BPSK modulation. After making decisions on $\{\hat{S}_{i-1,k}\}_{k=0}^{N-1}$, the decision result are then converted back to time-domain symbols $\{\tilde{s}_{i-1,n}\}_{n=0}^{N-1}$. For single-carrier mode, transform $\{\hat{S}_{i-1,k}\}_{k=0}^{N-1}$ into the time domain. Afterwards, make a partial decision as well as multi-carrier mode.

(4) The final output $\{z_{i-1,n}\}_{n=N}^{N+L-2}$ is the linear convolution between the last M elements of $\{\tilde{s}_{i-1,n}\}_{n=0}^{N-1}$ denoting as $\{\tilde{s}_{i-1,n}\}_{n=N-M}^{N-1}$ and $\{\tilde{h}_{i,n}\}_{n=0}^{L-1}$.

B. Channel Estimation Method

In Step 7, $\{u_{i,n}^{(J)}\}_{n=0}^{M+L-2}$ is the combination of received and estimated signal as the estimation of the linear convolution of $\{c_{i,n}\}_{n=0}^{M-1}$ and $\{h_{i,n}\}_{n=0}^{L-1}$. The CIR estimation can be expressed as

$$\bar{h}_{i,n}^{(J+1)} = \mathrm{IFFT}_K\left[\frac{\mathrm{FFT}_K(u_{i,n}^{(J)})}{C_{i,k}}\right], \quad 0 \le n < N. \tag{22}$$

To reduce the impact of noise, a threshold is used to select the significant taps of CIR. Since the length of the CIR is known to be smaller than M, the CIR values $\{\bar{h}_{i,n}^{(J+1)}\}_{n=M}^{K-1}$ are noise-only existing part which can be used for AWGN level estimation. The adaptive threshold is given by [14]

$$\lambda = \frac{2}{K-M}\sum_{n=M}^{K-1}\left|\bar{h}_{i,n}^{(J)}\right|^2 . \tag{23}$$

Then the square of absolute values of $\{\bar{h}_{i,n}^{(J+1)}\}_{n=0}^{L-1}$ below the threshold are set to zero as well as $\{\bar{h}_{i,n}^{(J+1)}\}_{M=0}^{K-1}$. Finally, a new CIR is obtained as

$$\hat{h}_{i,n}^{(J+1)} = \begin{cases} \bar{h}_{i,n}^{(J+1)}, & \left|\bar{h}_{i,n}^{(J+1)}\right|^2 > \lambda,\ 0 \le n < M \\ 0, & otherwise \end{cases} \tag{24}$$

and will be used for the next iteration.

C. Computation Complexity Comparison

Taking DTMB system as an example with parameters of $N = 3780$, $M = 420$ and $K = 2048$, the computational complexity comparison in term of IFFT/FFT resources between the proposed scheme and the conventional method [8] is shown in Table I. It can be seen from this table, the proposed channel estimation scheme has a much lower complexity as the conventional one.

TABLE I
COMPUTATION COMPLEXITY COMPARISON BETWEEN CONVENTIONAL AND PROPOSED METHOD

Operation	Method in [8]	Proposed method
I/FFT 2048	4(*J*+1)	3*J*+4
I/FFT 3780	2	2
I/FFT 8192	3(*J*+1)	0

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 method in [8] 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.

V. SIMULATION RESULTS

In this section, simulations are performed to demonstrate the performance of the proposed iterative channel estimation algorithm. The main simulation parameters based on Chinese DTMB standard [15] are shown in Table II. Two typical broadcasting 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 Rayleigh fading and the overall path losses are normalized.

TABLE II
MAIN SIMULATION PARAMETERS

Symbol Rate f_s	7.56 MSPS
Signal Constellation	QPSK/ 16QAM
FH Length M	420
FB Length N	3780
FFT Size K	2048
Maximal Doppler Spread	10 Hz
Operating Mode	Multi-carrier

TABLE III
POWER DELAY PROFILE

Tap	Delay (μs)	Power (dB)
1	-1.8	-18
2	0	0
3	0.15	-20
4	1.8	-20
5	5.7	-10
6	30	0

The symbol error rate (SER) as a function of signal-to-noise ratio (SNR) in dB of a QPSK/16QAM modulated DTMB system performance under DVB-T F1 channel is presented in Fig. 6. The SNR performance based on the 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 [8] 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 observed from the figure that the proposed algorithm offers improvement of the SNR performance after the first iteration. Two iterations can bring much better performance, about 2 dB SNR improvement at SER of 5×10^{-2} for 16QAM modulation mode. Furthermore, the proposed iterative method with two iterations improves the performance of SNR about 1.5 dB than that of the method in [8] with the same iteration number, and is only about 1 dB SNR degradation compared to the lower bound at SER of 5×10^{-1}.

Fig. 7 shows the SNR performance of DTMB system over CDT8 channel with the same modulation scheme above. For a SER of 10^{-2}, the SNR degradation of the performance based on the proposed reconstruction method with two iterations is about 1 dB compared with the ideal estimation with QPSK modulation. Moreover, the SNR improvement is about 1 dB and 3 dB than the proposed method with 1 iteration and the method in [8], respectively, at SER of 10^{-2}.

VI. CONCLUSIONS

In this paper, an iterative channel estimation algorithm is proposed for the TDS-OFDM systems based on PN sequence reconstruction through the combination of the overlapping part and the non-overlapping part of the received FH 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 simulation, it is shown that the SNR degradation of the proposed channel estimation method is comparable to the 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 be applied to the DTMB receiver-chip design directly.

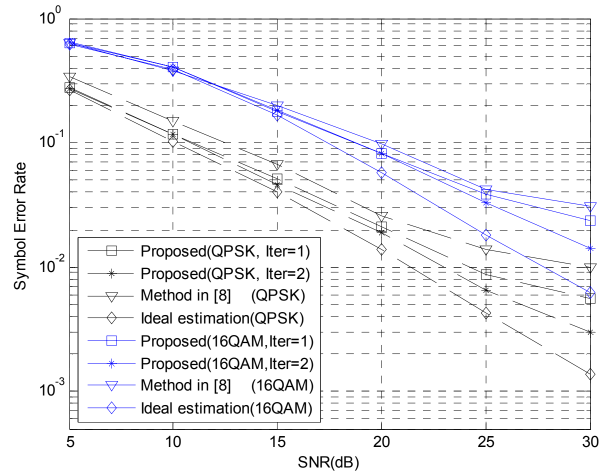

Fig. 6. SER performance of DTMB system over DVB-T F1 channel.

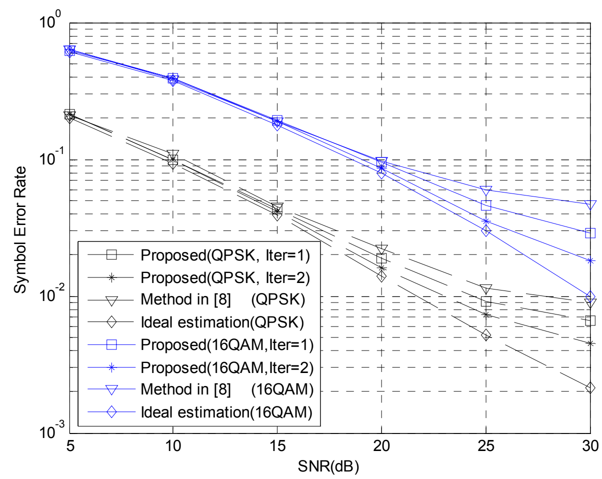

Fig. 7. SER performance of DTMB system over CDT8 channel.

REFERENCES

[1] "Digital Video Broadcasting (DVB): Frame Structure, Channel Coding and Modulation for Digital Terrestrial Television," *ETSI, Tech. Rep.* EN300 744 v1.1.2, Aug. 1997.

[2] I. Koffman and V. Roman, "Broadband wireless access solutions based on OFDM access in IEEE 802.16," *IEEE Commun. Mag.*, vol. 40, no. 4, pp. 96-103, Apr. 2002.

[3] Z. Yang, J. Wang, C. Pan, et al. "A combined code acquisition and symbol timing recovery method for TDS-OFDM," *IEEE Trans. Broadcast.*, vol. 49, no 3, pp. 304-308, Sept. 2003.

[4] J. Song, Z. Yang, L. Yang, K. Gong, C. Pan, J. Wang, and Y. Wu, "Technique review on Chinese digital terrestrial television broadcasting standard and measurements on some working modes," *IEEE Trans. Broadcast.*, vol. 53, no. 1, pp. 1-7, Mar. 2007.

[5] B. Song, L. Gui, Y. Guan, and W. Zhang, "On channel estimation and equalization in TDS-OFDM based terrestrial HDTV Broadcasting System," *IEEE Trans. Consumer Electronics*, vol. 51, no. 3, pp. 790-797, Aug. 2005.

[6] Z. Yang, J. Wang, C. Pan, L. Yang, and Z. Han, “Channel estimation of DMB-T,” in *Proc. 2002 IEEE Int. Conf. Communications, Circuits and Systems and West Sino Expositions* , vol. 2, pp. 1069-1072, 2002.
[7] G. Liu and J. Zhang, “ITD-DFE Based Channel Estimation and Equalization in TDS-OFDM Receivers,” *IEEE Trans. Consumer Electronics*, vol. 53, no. 2, pp. 304-309, May. 2007.
[8] J. Wang, Z. Yang, C. Y. Pan, J. Song, and L. Yang, “Iterative Padding Subtraction of the PN Sequence for the TDS-OFDM over Broadcast Channels,” *IEEE Trans. Consumer Electronics*, vol. 51, no. 4, pp. 1148-1152, Nov. 2005.
[9] S. Tang, F. Yang, K. Peng, C. Pan, and Z. Yang, “Iterative channel estimation for block transmission with know symbol padding – a new look at TDS-OFDM,” in *Proc. 2007 IEEE Global Telecommunications Conf. (GLOBECOM’07),* Washington, DC, pp. 4269-4273, Nov. 2008.
[10] D. Kim and G. L. Stüber, “Residual ISI cancellation for OFDM with application to HDTV broadcasting,” *IEEE J. Select. Areas Commun.*, vol. 16, no. 8, pp. 1590–1599, Oct. 1998.
[11] T. Hwang and Y. Li, “Iterative cyclic prefix reconstruction for coded single-carrier systems with frequency-domain equalization (SC-FDE),” in *Proc. IEEE Vehicular Technology Conf. (VTC’03)*, vol. 3, pp. 1841–1845, Apr. 2003.
[12] C. Yeh, Y. Lin, and Y. Wu, “OFDM system channel estimation using time-domain training sequence for mobile reception of digital terrestrial broadcasting,” *IEEE Trans. Broadcast.*, vol.46, no 3, pp. 215–220, Sept. 2000.
[13] B. Muquet, Z. Wang, G. B. Giannakis, M. de Courville, and P. Duhamel, "Cyclic Prefixing or Zero Padding for Wireless Multicarrier Transmissions?" *IEEE Trans. Commun.*, vol.50, no. 12, pp. 2136-2148, Dec. 2002.
[14] Y. Kang, K. Kim, and H. Park, “Efficient DFT-based channel estimation for OFDM systems on multipath channels,” *IET Commun.*, vol.1, no.2, pp. 197-202, Apr. 2007.
[15] “Framing Structure, Channel Coding and Modulation for Digital Television Terrestrial Broadcasting System (in Chinese)”, Chinese National Standard GB 20600-2006, Aug. 2006.

Fang Yang received the B. Eng. degree from the Department of Electronic Engineering in Tsinghua University, Beijing China, in 2005. Currently, he is pursuing the Ph.D. degree at the DTV Technology R&D Center, Tsinghua University. His research interests lie in the field of channel estimation and interference cancellation for digital wireless communication system, space-time coding and diversity techniques, as well as the training sequence design.

Jintao Wang received his B.Eng and Phd.D degrees in Electrical Engineering both from Tsinghua University, Beijing, China in 2001 and 2006, respectively. Since 2006, he has been an assistant professor of Tsinghua's DTV Technology R&D center. He is the standard committee member for the Chinese national digital terrestrial television broadcasting standard. His current research interest is in the area of the broadband wireless transmission, especially the channel estimation and space-time coding techniques.

Jun Wang was born in Henan, P. R. China, on October 5, 1975. He received the B. Eng. and Ph.D degree from the Department of Electronic Engineering in Tsinghua University, Beijing, China, in 1999 and 2003 respectively. He is an assistant professor and member of Digital TV R&D center of Tsinghua University since 2000. His main research interests focus on broadband wireless transmission techniques, especially synchronization and channel estimation. He is actively involved in the Chinese national standard on the Digital Terrestrial Television Broadcasting technical activities, and is selected by the Standardization Administration of China as the Standard committee member for drafting.

Jian Song received his B. Eng and Ph.D degrees in Electrical Engineering both from Tsinghua University, Beijing China in 1990 and 1995, respectively and worked for the same university upon his graduation. He has worked at The Chinese University of Hong Kong and University of Waterloo, Canada in 1996 and 1997, respectively. He has been with Hughes Network Systems in USA for 7 years before joining the faculty team in Tsinghua in 2005 as a professor. He is now the director of Tsinghua’s DTV Technology R&D center. His primary research interest is in physical layer and has been working in quite different areas of fiber-optic, satellite and wireless communications, as well as the powerline communications. His current research interest is in the area of digital TV broadcasting. Dr. Song has published more than 50 journal and conference papers and holds one US patent.

Zhixing Yang received is a full professor at the Department of Electronics Engineering of Tsinghua University, China. He is the executed director of the State Key Laboratory on Microwave & Digital Communications, China, and the executed director of the development group of the digital television terrestrial broadcasting state standard for China. He received several awards and held several patents. His research interests are in high-speed data transmission over broadband digital television terrestrial broadcasting, wireless links, wireless communication theory and communication systems design.