ITD-DFE Based Channel Estimation and Equalization in TDS-OFDM Receivers

Guanghui Liu, *Member,* IEEE, and Junling Zhang

Abstract —*As an alternative to the traditional cyclic prefixed OFDM (CP-OFDM) technology, time-domain synchronous OFDM (TDS-OFDM) provides higher spectrum efficiency due to no pilot in the transmitted signal. However, the pilot-Aided channel estimation techniques in CP-OFDM cannot be applied to TDS-OFDM. In this paper, channel impulse responses are estimated in time domain based on the linear correlation between the received samples and the locally generated PN sequence. The iterative threshold detection (ITD), as well as the combination with decision feedback equalization (DFE), is proposed to improve the precision of channel estimation. It is shown by simulation that the proposed algorithm, ITD combined with DFE (ITD-DFE), performs very well even in multipath environments with long-delay and large-amplitude echoes*[1].

Index Terms — TDS-OFDM, channel estimation, linear correlation, iterative threshold detection, decision feedback equalization.

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) is a very effective technique to mitigate inter-symbol interference (ISI) in handling time dispersion of multipath fading channels. Conventionally, OFDM symbols are separated by CP in order that receivers can demodulate data very simply. Recently, TDS-OFDM has been attracting more and more interests, in which pseudonoise (PN) sequences take the place of CP for serving as guard intervals and also as training symbols. Chinese standard of digital terrestrial television broadcasting (DTTB), has adopted TDS-OFDM as one of its modulation schemes [1].

In TDS-OFDM systems, the inserted PN sequences acts as training symbols for the purposes of the receiver's synchronization and channel estimation, which makes it unnecessary to add pilots, as DVB-T system [2], in the transmitted signal spectrum. For this reason, basically TDS-OFDM can obtain higher channel throughout than CP-OFDM. Nevertheless, many algorithms in CP-OFDM including synchronization and channel estimation are so different that they cannot be applied to TDS-OFDM directly. In this paper, we will concentrate on the channel estimation and equalization in TDS-OFDM systems.

There are several literatures devoted to channel estimation in TDS-OFDM systems. In [3-4], on the basis of the discrete Fourier transformation (DFT) of the received PN sequence, the channel estimation is performed in frequency domain. This method may be problematic in the presence of long time-delay echoes. In [5], the channel impulse response (CIR) is estimated in time domain based on the cyclical correlation between the received frame head and locally generated PN sequence, and the algorithm with three DFE iterations can estimate long-delay echoes. Literature [6] introduces a method of iterative padding subtraction to cancel ISI between PN and random data blocks, and the algorithm in [7] can deal with channel equalization over doubly selective channels.

In this paper, the receiver first computes the linear correlation between the received samples and the locally generated PN, and estimates CIR according to the correlation results in which all kinds of interference are cancelled iteratively. In the case of long-delay and large-amplitude echoes, the algorithm is combined with DFE to improve the CIR estimation precision. The proposed method is termed as iterative threshold detection combined with decision feedback equalization (ITD-DFE). The estimated CIR is used to cancel ISI between PN and OFDM data, and channel equalization is performed in frequency domain on the basis of the DFT of the zero-padding CIR.

The remaining paper is organized as follows. Section II briefly describes the frame structure as well as the baseband system model. The ITD-DFE algorithm is formulated in section III. In section IV, we will discuss the ISI cancellation and channel equalization, and simulation results are presented in section V to demonstrate the effectiveness of the proposed algorithm. Finally, our conclusions are drawn in section VI.

II. TDS-OFDM SYSTEM DESCRIPTION

In TDS-OFDM systems, the transmitted signal consists of two parts (shown as Fig. 1), the frame head and the frame body. The L_{PN} -length frame head has a pre-amble, PN sequence and a post-amble. The PN sequence with L_m samples is generated based on m-sequence, and the pre-/post-amble is the cyclical extensions of the PN and has the length of L_{pre} and L_{post} respectively. OFDM modulation scheme is used for the frame body, i.e. the N_c -length IDFT block.

The baseband model of the TDS-OFDM system is shown as Fig. 2. At the transmitter side, the frame body is generated on the basis of the IFFT of the mapping symbols $\{S_i(k)\}_{k=0}^{N_c-1}$ with $N_c = 3780$ denoting the size of IFFT. After the insertion of

[1] Guanghui Liu is with Samsung Electronics Co. Ltd., Suwon, Republic of Korea (e-mail: ghliu@ieee.org).

Junling Zhang is with Philips Research East Asia (PREA), Shanghai, China (e-mail: eric.jl.zhang@philips.com).

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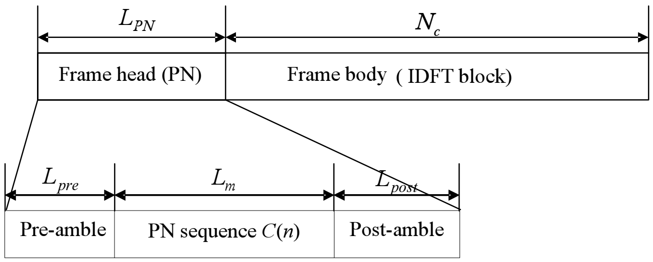

Fig. 1. The frame structure of TDS-OFDM system

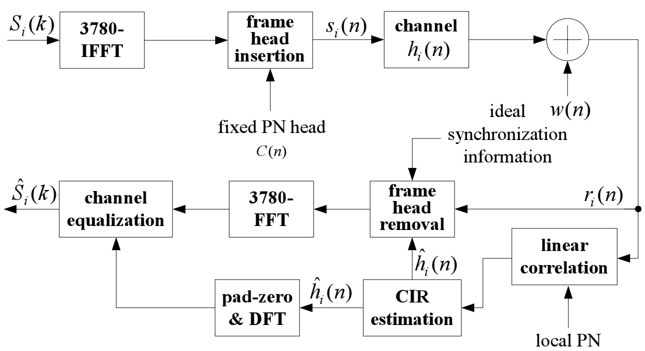

Fig. 2. The functional block diagram of TDS-OFDM system

frame head $\{p(n)\}_{n=0}^{L_{PN}-1}$, we get the baseband signal $s_i(n)$ of the i-th OFDM symbol with sampling rate f_s. BPSK modulation is used in the frame head and the real part is the same as the imaginary part for robust synchronization and channel estimation. At the receiver side, the frame head is removed from the received signal $r_i(n)$ in the upper arm, and in the lower arm the receiver computes the linear correlation between the local PN $C(n)$ and $r_i(n)$, subsequently estimates CIR according to the correlation result. OFDM demodulation is implemented by 3780-point FFT operation, and the frequency domain signal is equalized based on the channel frequency response derived from the estimated CIR $\hat{h}_i(n)$. For the sake of simplicity, we assume perfect frequency and timing synchronization, and channel forward error control (FEC) is not considered in this model.

III. ITD-DFE Based Channel Estimation

The most popular radio channel model is the so-called discrete multipath time-domain model, which is usually modeled as a linear filter and it is assumed to be a slow-varying channel whose response does not vary within an OFDM symbol period. The CIR can be expressed as

$$h_i(n) = \sum_{r=0}^{L-1} h_r \delta(n - \tau_r) \tag{1}$$

where parameters L, h_r and τ_r represent the number of paths, the complex-valued path gain and time delay of the r-th path, respectively. $\delta(n)$ denotes the Dirac delta function. The task of channel estimation is mainly to estimate the CIR parameters.

A. Linear Correlation Analysis

As shown in Fig. 2, the sampled received signal is

$$r_i(n) = s_i(n) \text{ e } h_i(n) + w(n) \tag{2}$$

where $w(n)$'s are additive Gaussian white noise (AWGN) with zero mean and variance σ^2, and e stands for linear convolution operation. Due to the linear property of mathematical operation, we can compute linear correlation between the received samples and the local PN sequence as

$$\begin{aligned} R_{rc}(n) &= [s_i(n) \text{ e } h_i(n) + w(n)] \otimes C(n) \\ &= s_i(n) \text{ e } h_i(n) \otimes C(n) + w(n) \otimes C(n) \\ &= [s_i(n) \otimes C(n)] \text{e } h_i(n) + w'(n) \end{aligned} \tag{3}$$

where $\otimes$ denotes linear correlation operation, and $w'(n)$'s are also the AWGN with zero mean and variance σ^2 for the power normalized PN sequence. From (3), it follows that we could regard the correlation result as the channel CIR's convolution with the linear correlation between the transmitted samples and the local PN. The linear correlation is rewritten as

$$R_{sc}(n) = s_i(n) \otimes C(n) = \sum_{k=0}^{L_m-1} C^*(k) s_i(n+k) \tag{4}$$

where * denotes the complex conjugate.

Each OFDM symbol can be decomposed into two non-overlapping parts in time domain

$$s_i(n) = PN(n) + d_i(n) \tag{5}$$

where

$$PN(n) = \begin{cases} p(n), & 0 \le n < L_{PN} \\ 0, & \text{else} \end{cases} \tag{6}$$

$$d_i(n) = \begin{cases} s_i(n), & L_{PN} \le n < L_{PN} + N_c \\ 0, & \text{else} \end{cases} \tag{7}$$

From (4) and (5), we have

$$\begin{aligned} R_{sc}(n) &= PN(n) \otimes C(n) + d_i(n) \otimes C(n) \\ &= R_{pc}(n) + R_{dc}(n) \end{aligned} \tag{8}$$

In (8), $R_{pc}(n)$'s and $R_{dc}(n)$'s denote the local PN's linear correlation with the frame head and the transmitted frame-body data respectively, and the latter is the interference term for the CIR estimation.

We can express $R_{pc}(n)$ as

$$\begin{aligned} R_{pc}(n) &= \sum_{k=0}^{L_m-1} C^*(k) PN(n+k) \\ &= \begin{cases} -2, & 0 \le n < L_{pre} + L_{post}, n \ne L_{pre} \\ 2L_m, & n = L_{pre} \\ noise, & \text{else} \end{cases} \\ n &= -L_m + 1, -L_m + 2, \cdots, L_{PN} - 1 \end{aligned} \tag{9}$$

Only within the range $0 \le n < L_{pre} + L_{post}$, $p(n+k) = C(n+k)$ holds for $k = 0, 1, \cdots, L_m - 1$, so the linear correlation is equivalent to the cyclical correlation. Note that the correlation amplitude is real and doubled because the real and imaginary parts of $p(n)$ are the same. It is obvious that the correlation results in (9) do not satisfy the properties of Delta function, which could introduce the inter-path interference for the CIR detection. We can decompose $R_{pc}(n)$ into two terms: the expectation term and the interference term $I(n)$.

$$R_{pc}(n+L_{pre}) = G\delta(n) + I(n), \quad n = -L_m - L_{pre}+1,\cdots,L_{PN}-L_{pre}-1 \tag{10}$$

where

$$G = R_{pc}(L_{pre}) \qquad I(n) = \begin{cases} 0, & n=0 \\ R_{pc}(n+L_{pre}), & \text{else} \end{cases} \tag{11}$$

The normalized factor G depends upon the average power of the frame head.

Figure 3 shows $R_{pc}(n+L_{pre})$ vs. n when $L_{pre}=83$, $L_{post}=82, L_m=255, L_{PN}=420$, and the average power of the frame head is normalized to two times that of the transmitted OFDM signal. From the correlation result, in addition to the main peak, we can see two parasitical peaks at ±255 resulting from the pre-/post-amble. In the central area ($-L_{pre} \le n < L_{post}$), owing to the cyclically extensive structure of the frame head, the linear correlation is equivalent to cyclical correlation. Outside the cyclical correlation area, there is correlation noise regarded as the ruined cyclical correlation due to the limited length of the pre-/post-amble.

From (3) and (8), we can get the receiver's correlation

$$\begin{aligned} R_{rc}(n) &= R_{pc}(n) \otimes h_i(n) + R_{dc}(n) \otimes h_i(n) + w'(n) \\ &= \sum_{k=0}^{M-1} h_k R_{pc}(n-\tau_k) + R_{dc}(n) \otimes h_i(n) + w'(n) \\ &= \sum_{k=0}^{M-1} G h_k \delta(n-\tau_k) + \underbrace{\sum_{k=0}^{M-1} h_k I(n-\tau_k)}_{\text{Type I}} + \underbrace{R_{dc}(n) \otimes h_i(n)}_{\text{Type II}} + \underbrace{w'(n)}_{\text{Type III}} \end{aligned} \tag{12}$$

In (12) we change the time scale, which is implemented by means of symbol timing synchronization. From (12), it follows that the first term contains the expected correlation peaks to estimate the CIR gains h_k's, and the corresponding path delay τ_k can be obtained by the position of the peaks, other terms are the interference terms classified as the following three types:

- Type I: the inter-path interference resulting from the non-ideal correlation properties of the frame head.
- Type II: the random-data time delay spread.
- Type III: channel AWGN.

B. Channel Estimation Based on Iterative Threshold Detection (ITD)

It is shown by the above discussion that if three types of interference in (12) are ignored, we can get the CIR estimate:

$$\hat{h}_k \approx \frac{R_{rc}(\tau_k)}{G} \tag{13}$$

It is applicable especially when the maximum time spread $\tau_{\max}$ is less than L_{pre}. Fig. 4 demonstrates the amplitude of $R_{rc}(n)$ in a multipath channel with a short-delay 0dB echo. And channel noise is not considered here for simplicity. It is

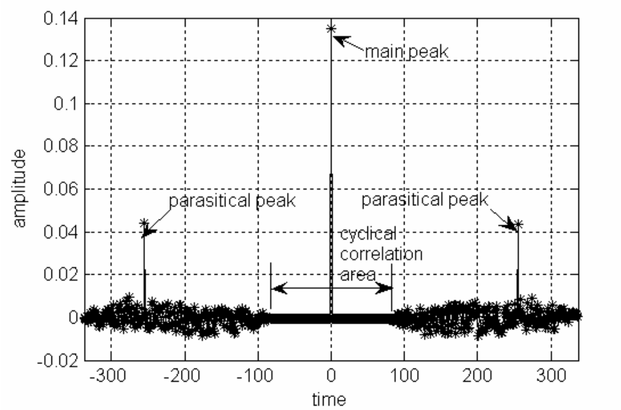

Fig. 3. Linear correlation between the frame head and the local PN

so clear in the cyclical correlation area that we can estimate the CIR using (13) if multipath correlation peaks are distinguished form parasitical peaks. In the presence of long-delay and large-amplitude echoes (as shown in Fig. 5), the central correlation area would be ruined completely, which make it so difficult to detect the short-delay and small-amplitude echoes. In this case, the CIR estimation using (13) would result in considerable performance loss. Hence, it is quite necessary that the interference is canceled as much as possible before the CIR estimation.

On the basis of the correlation in (12), we can estimate the CIR using the ITD algorithm as the following steps:

(1) Set the initially relative threshold $TH_{relative}$ and make $R_{rc}^{iter=1}(n) = R_{rc}(n)$. The relative threshold is defined as the ratio of the minimum multipath amplitude aimed for detection to the maximum multipath amplitude.

(2) Searching for the largest peak value $\hat{a}(\tau_{\max})$ of $R_{rc}^{iter=1}(n)$ with $\tau_{\max}$ denoting the position of the peak, the initial peak-detection threshold is set to $th_1 = \hat{a}(\tau_{\max})/M$, where the constant M is chose as to make th_1 larger than the maximum parasitical peaks whose amplitudes can be computed as $\hat{a}(\tau_{\max})L_{pre}/L_{PN}$ or $\hat{a}(\tau_{\max})L_{post}/L_{PN}$.

(3) And for the l-th iteration, finding the peaks higher than

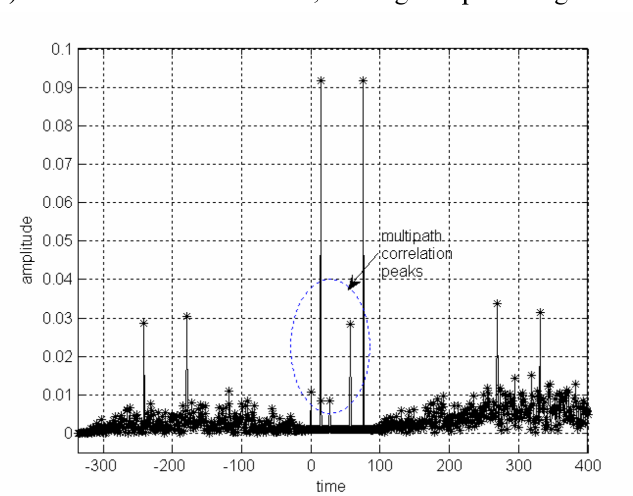

Fig. 4. Linear correlation between the received signal and the local PN

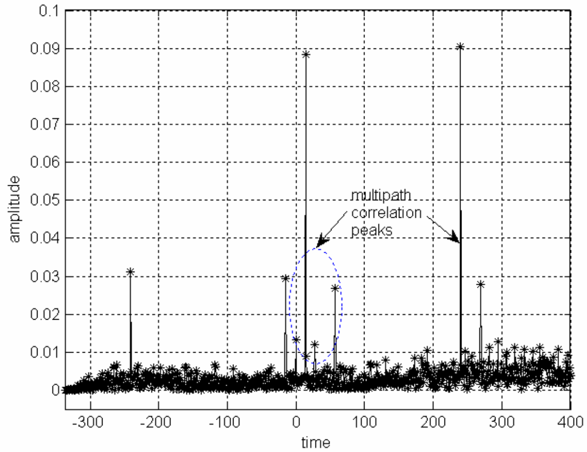

Fig. 5. Linear correlation between the received signal and the local PN

the threshold th_l in $R_{rc}^{iter=l}(n)$, we can get the observation vector $\mathbf{A}_l = [\hat{a}(\tau_0), \hat{a}(\tau_1), \cdots, \hat{a}(\tau_{Q_l-1})]^T$ with Q_l and $(\cdot)^T$ denoting the number of peaks detected in the l-th iteration and the matrix transpose operation respectively.

(4) The CIR amplitude is estimated based on the observation vector $\mathbf{A}_l$. Provided only the interference of type I in (12) is considered here, we can rewrite (12) as the following matrix form:

$$\mathbf{B}_l \hat{\mathbf{H}}_l = \mathbf{A}_l \tag{14}$$

where $\hat{\mathbf{H}}_l = [\hat{h}_0, \hat{h}_1, \cdots, \hat{h}_{Q_l-1}]^T$ is a $Q_l \times 1$ vector, and $\mathbf{B}_l = (b_{ij})_{Q_l \times Q_l}$ is a $Q_l \times Q_l$ matrix with

$$b_{ij} = \begin{cases} G, & i = j \\ I(\tau_i - \tau_j), & i \neq j \end{cases} \tag{15}$$

The interference coefficient matrix $\mathbf{B}_l$ is known for the receiver, so we can estimate the CIR vector as

$$\hat{\mathbf{H}}_l = \mathbf{B}_l^{-1} \mathbf{A}_l \tag{16}$$

where $(\cdot)^{-1}$ denotes the matrix inversion.

(5) On the basis of $\hat{\mathbf{H}}_l$, we can subtract the interference of type I partially as (17) for the purpose of detecting the smaller multipaths. The interference cancellation at the l-th iteration is performed as

$$\begin{aligned} R_{rc}^{iter=l+1}(n) = R_{rc}^{iter=l}(n) &- \sum_{k=0}^{Q_l-1} \hat{h}_k I(n - \tau_k) \\ &- \sum_{k=0}^{Q_l-1} \hat{h}_k G \delta(n - \tau_k) \end{aligned} \tag{17}$$

(6) If current threshold value th_l is larger than $\hat{a}(\tau_{\max}) TH_{relative}$, then $th_{l+1}=th_l/M$ and go back to step (3), else jump to step (7).

(7) Vector $\hat{\mathbf{H}}_l$ in all iterations can be formed as the final CIR estimate. Padding the estimated CIR with zero to the length of N_c, we can obtain the channel frequency response by means of the DFT of the padding-zero CIR vector.

C. DFE Based Random-Data Interference Cancellation

In the presence of long-delay and large-amplitude echoes, the cyclical correlation area would be ruined severely by the interference of type I as well as type II. Accordingly the mean squared error (MSE) performance of channel estimation would get worse, and even many short-delay and small-amplitude paths could not be detected. It is quite necessary in this case to reduce the interference of type II in (12).

Before the channel estimation using the ITD algorithm, we can cancel the interference introduced by random data of the last OFDM symbol. In (7), we can set the data vector as

$$d_i(n) = \begin{cases} \hat{s}_{i-1}(n + N_c), & n < 0 \\ 0, & \text{else} \end{cases} \tag{18}$$

where $\hat{s}_{i-1}(n)$ is the frame body of the last OFDM symbol derived from the last OFDM demodulation and modulation. So the correlation data used to estimate the CIR can be calculated as

$$R_{rc}^{iter=1}(n) = R_{rc}(n) - R_{dc}(n) \text{e } \hat{h}_{i-1}(n) \tag{19}$$

where $\hat{h}_{i-1}(n)$ is the last CIR estimate. If the channel state changes rapidly, computation in (19) could introduce extra error, and this step cannot be performed.

The CIR estimate based on the ITD algorithm is the initial result for demodulating the frame-body data, and hard-decision data is used to form the transmitted frame-body $\hat{s}_i(n)$. Then we can update the data vector and reduce the data interference from the receiver's correlation.

$$d_i(n) = \begin{cases} \hat{s}_{i-1}(n + N_c), & n < 0 \\ \hat{s}_i(n), & L_{PN} \le n < L_{PN} + N_c \\ 0, & 0 \le n < L_{PN} \end{cases} \tag{20}$$

$$R_{rc}^{iter=1}(n) = R_{rc}(n) - R_{dc}(n) \text{e } \hat{h}_i(n) \tag{21}$$

Note that the term $\hat{h}_i(n)$ in (21) is the initial CIR estimate in the current OFDM symbol period, which would be reestimated after the cancellation of random-data interference.

IV. ISI Cancellation and Channel Equalization

In conventional CP-OFDM systems, the CP can be discarded directly in the receiver due to the inherently cyclical convolution between OFDM symbols and CIR taps. Nevertheless, this method is not applicable in TDS-OFDM receiver, since the inserted frame head is not the copy of tails of OFDM symbols.

As shown in Fig. 6, the periodically inserted frame head $p(n)$ acts as time guard interval (TGI) in practice. Due to channel time-delay spread, the frame body $s_i(n)$ of the i-th OFDM symbol $r_i(n)$ is affected by the tail of current frame head, meanwhile the tail of $s_i(n)$ interferers the frame head of the next OFDM symbol $r_{i+1}(n)$. In TDS-OFDM receivers, the

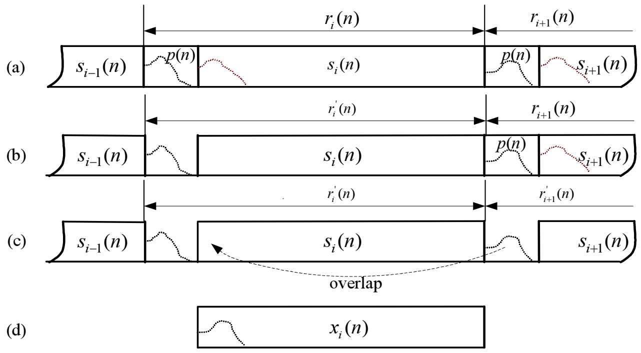

Fig. 6. Illustration of frame-head removal and the recovery of cyclical convolution. (a) is the received signal frames; (b) demonstrates the frame-head removal of current frame; (c) demonstrates the frame-head removal of next frame; (d) is the recovered frame body of current frame.

purpose of frame-head removal is to cancel current frame head's interference to $s_i(n)$ and overlap the tail of $s_i(n)$ to its start, which recovers the cyclical convolution between $s_i(n)$ and CIR tap $h_i(n)$.

We can summarize the process of frame-head removal as follows (in Fig. 6):

(1) According to CIR estimate $\hat{h}_i(n)$, compute the linear convolution between the frame head $p(n)$ and $\hat{h}_i(n)$ as

$$R_{pc}^{(i)}(n) = p(n) \otimes \hat{h}_i(n) \tag{22}$$

and subtract $R_{pc}^{(i)}(n)$ from $r_i(n)$ as

$$r_i^{'}(n) = r_i(n) - R_{pc}^{(i)}(n) \tag{23}$$

(2) Similarly, for the (i+1)-th OFDM symbol $r_{i+1}(n)$, compute

$$R_{pc}^{(i+1)}(n) = p(n) \otimes \hat{h}_{i+1}(n) \tag{23}$$

$$r_{i+1}^{'}(n) = r_{i+1}(n) - R_{pc}^{(i+1)}(n) \tag{24}$$

(3) For the i-th OFDM symbol, we can compute the ISI-free frame body $\{x_i(n)\}_{n=0}^{N_c-1}$ as (25) by adding the time spread of current frame body to its start for maintaining cyclical convolution relation.

$$x_i(n) = \begin{cases} r_i^{'}(n+L_{PN}) + r_{i+1}^{'}(n), & 0 \le n \le \hat{n}_{\max} \\ r_i^{'}(n+L_{PN}), & \hat{n}_{\max} < n < N_c \end{cases} \tag{25}$$

where $\hat{n}_{\max}$ represents the estimated maximum delay spread.

After the ISI cancellation and the recovery of cyclical convolution relation, the receiver can perform the channel equalization as CP-OFDM systems.

$$\hat{S}_i(k) = X_i(k) / \hat{H}_i(k) \tag{26}$$

where $\hat{H}_i(k)$'s denote the estimated channel frequency response, and $X_i(k)$'s are the FFT of $\{x_i(n)\}_{n=0}^{N_c-1}$. Over doubly selective channels, the equalization method in [7] can obtain much greater performance improvements than zero-forcing equalization as (26).

V. Numerical Simulation

Based on the TDS-OFDM system with parameters in Table I, we now evaluate the symbol-error-rate (SER) performance of the proposed algorithm. Two typical broadcast channels are chosen: Fixed Reception F1 version of DVB-T channel model [2] as well as China Test 8th (CT8) channel model in Table II. In DVB-T F1 channel, there are no long-delay echoes more than 10 μs, so the interference to the CIR detection is mainly the inter-path interference (Type I). While in CT8 channel, there is a long-delay 0dB echo in order to verify the interference-cancellation performance of DFE iteration.

Fig. 7 shows the SER performance in DVB-T F1 channel. The corresponding SER curves of ideal channel estimation and using the method in [6] with 3 iterations are provided for reference. From Fig. 7, it follows that the SER performance with only ITD method (N_{dfe}=0) is so close to the performance provided in [6]. The ITD method with one DFE iteration (N_{dfe}=1) can improve the SNR about 3dB at SER=0.1, which is very close to the ideal performance curve. Comparing to Fig. 7, the performance gain in Fig. 8 gained by the DFE iteration is much more obvious in the presence of long-delay and large-amplitude echo. In this case, the random-data interference (type II) makes it so difficult to find the short-delay and small-amplitude echoes. Accordingly, the ITD method even with 2 DFE iterations cannot make the SER performance close to the ideal channel estimation, nevertheless at SER=0.1 there exists about 2.5dB improvement with respect to the method in [6] with 3 iterations. It is also shown by Fig. 7 and Fig. 8 that the ITD method with one DFE iteration (N_{dfe}=1) is sufficient to the performance improvement, and more DFE iterations could not produce the desired improvements.

TABLE I
SYSTEM SIMULATION PARAMETERS

Symbol Rate f_s	7.56MSPS
Signal Constellation	64QAM
FFT size N_c	3780
Subcarrier Spacing	2KHz
Frame Head Length L_{PN}	420
PN Sequence Length L_m	255
Pre-amble Length L_{pre}	83
Post-amble Length L_{post}	82

TABLE II
CT8 CHANNEL PROFILE

Path Number	Time Delay (μs)	Relative Power (dB)
1	0.0	0
2	-1.8	-18
3	0.15	-20
4	1.8	-20
5	5.7	-10
6	30.0	0

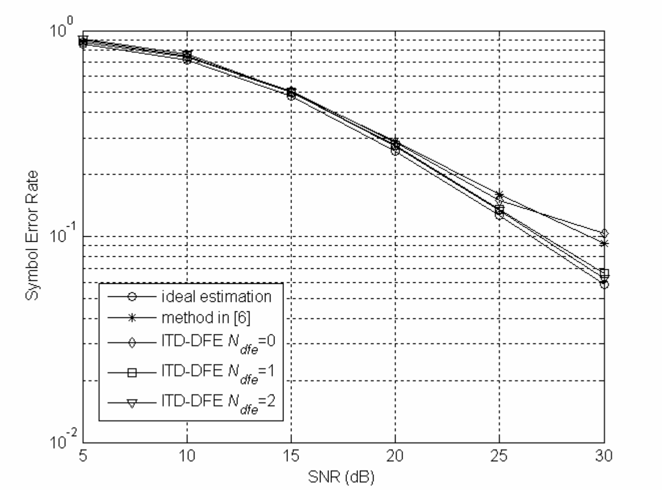

Fig. 7. The system SER performance in DVB-T F1 channel

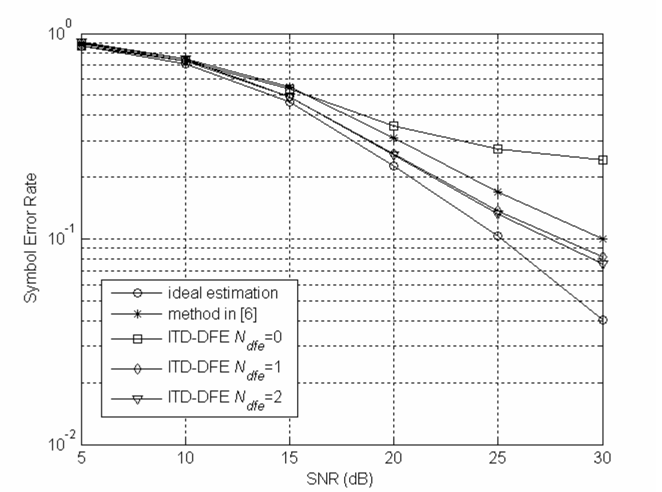

Fig. 8. The system SER performance in DVB-T F1 channel

VI. Conclusion

In this paper, the ITD-DFE algorithm is proposed to estimate channel impulse response in time domain on the basis of the linear correlation between the received samples and the local PN sequence. The iterative threshold values are used to search for multipath correlation peaks and the interference is canceled iteratively, which makes more accurate CIR estimate. The ITD method associated with DFE improves the system performance in multipath environment especially with long-delay and large-amplitude echoes. With the excellent performance and the lower complexity, the ITD-DFE algorithm is applicable in the practical TDS-OFDM systems.

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Guanghui Liu (M'06) received the M.Sc. and Ph.D. degrees in electronic engineering from University of Electronic Science and Technology of China (UESTC), Chengdu, Sichuan, China, in 2002 and 2005 respectively. Dr. Liu serves as a reviewer for the IEEE Transactions on Communications, the IEEE Transactions on Broadcasting and the IEEE Transactions on Wireless Communications. As a senior engineer, he is now with Samsung Electronics. His research interests include synchronization, channel estimation and equalization, phase-noise suppression in OFDM communication systems.

Junling Zhang received his Master and Ph.D degrees from School of Electronic Engineering, University of Electronic Science and Technology of China, in 1999 and 2003 respectively. From 1999 to 2001 he was a research assitant with City University of Hong Kong and The Hong Kong Polytechnic University, Hong Kong. From 2003 to 2006 he was the senior engineer with Samsung Electronics, South Korea. Since 2006, he has been with Philips Research East Asia, Shanghai China, as a senior scientist. His research interests include OFDM systems, wide band wireless communications and mobile communications.