On Channel Estimation and Equalization in TDS-OFDM based Terrestrial HDTV Broadcasting System

Bowei Song, Lin Gui, Yunfeng Guan and Wenjun Zhang

Abstract —*In TDS-OFDM (Time-Domain Synchronous Orthogonal Frequency Division Multiplexing) systems, pseudonoise (PN) sequences rather than cyclic prefixes are inserted as guard interval, between consecutive inverse discrete Fourier transformed (IDFT) symbol blocks. Since the PN sequences can also be used as training symbols, such system can provide higher spectrum efficiency. However, due to non-cyclic property of the signal, the simple channel estimation and equalization techniques for conventional cyclic prefixed OFDM (CP-OFDM) can not be applied to TDS-OFDM. In this paper, we propose a channel estimation and equalization method for TDS-OFDM. Channel estimation depends on time domain correlation and iterative interference cancellation techniques, while equalization is based on tail cancellation and cyclic restoration algorithm (TCCR). It is shown that our proposed method can provide satisfactory performance in TDS-OFDM based terrestrial high-definition television (HDTV) broadcasting system* [1]

Index Terms —Channel estimation, Correlation, DMB-T, Equalization, HDTV, Interference cancellation, OFDM.

I. INTRODUCTION

OFDM system is an attractive technique for broadband communications. By inserting cyclic prefix (CP) between different symbols, it can efficiently combat multipath channels and can be easily demodulated. However, in order to cope with long delayed multipath, longer CP is need; Furthermore, large amount of pilots are also inserted into data frames for system synchronization and channel estimation. These payloads will dramatically reduce bandwidth efficiency of the system.

In recent years, a technique called time-domain synchronous OFDM (TDS-OFDM) has received more and more interests. A typical application of TDS-OFDM is DMB-T system, a candidate terrestrial HDTV broadcasting standard in China [1-6]. In stead of CP, PN sequences are inserted as guard intervals. PN guard intervals can also be used as training symbols for synchronization and estimation purpose so there is no additional pilot symbol needed. For this reason, TDS-OFDM can provide higher system throughput than CP-OFDM.

However, with introducing PN guard interval, the cyclic property of the signal frame as in CP-OFDM does not hold any more, so the inter-symbol interference (ISI) is unavoidable. Neither conventional channel estimation algorithms nor simple one-tap frequency domain equalizer for CP-OFDM can be applied to TDS-OFDM directly.

In this paper, we will discuss the channel estimation and equalization technology for DMB-T system. Channel estimation is based on time domain correlation and iterative interference cancellation; for equalization, a technique called tail cancellation and cyclic restoration (TCCR) is proposed. By compared with the performance of DVB-T [7], it is shown that our proposed method can work well in DMB-T system.

This paper is organized as follows: In section II, we briefly overview the basic structure of the DMB-T system. In section III, we will discuss the details of channel estimation. After that, equalization method for DMB-T is proposed in section IV. In section V, computer simulation is presented and performance is compared with DVB-T. In section VI, we will conclude this paper finally.

II. DMB-T SYSTEM DESCRIPTION

The frame structure of DMB-T is shown in Fig.1. The Source bits can be mapped into QPSK, 16QAM, 64QAM according to different transmission mode. Each frame has one

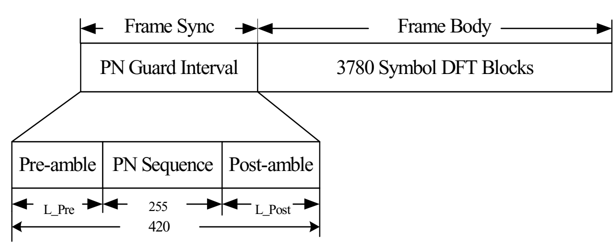

Fig. 1. The frame structure of DMB-T system [1]

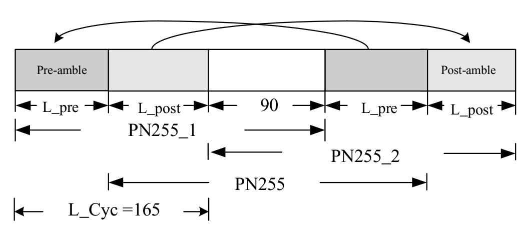

Fig.2. The structure of PN guard interval

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Frame Sync and one Frame Body. 3780 mapped symbols are grouped into one OFDM symbol and build up one Frame Body. A PN guard interval with length 420 is inserted before each Frame Body as Frame Sync. The detail of PN guard interval is shown in Fig.2. The PN guard interval consists of a pre-amble, a PN sequence with length 255 and a post-amble. PN255 is an m-sequence generated with 8-order linear feedback shift register. The last L_pre samples of PN255 compose the Pre-amble and the first L_post samples of PN255 compose the Post-amble. The whole 420-length guard interval is consequently in a quasi-cyclic form: PN255_1 and PN255_2 in Fig.2 are just shift vectors of PN255.

The symbol rate of DMB-T is 7.56M, the period of one OFDM symbol is 500 µs and the period of PN guard interval is 55.556 µs. The longest multipath delay that DMB-T system can combat is 55.556 µs. In the following part of this paper, we will discuss the channel estimation and equalization based on such frame structure.

III. Channel Estimation for DMB-T

The radio channel can be modeled as impulse response filter expressed by:

$$h(t) = \sum_{\tau=0}^{L-1} a(\tau)\delta(t - \tau T_s), \tag{1}$$

$a(\tau)$s are WSS, narrow band complex Gaussian processes, with L is the longest multipath delay. Time-varying channel is assumed quasi-static, i.e., constant during the transmission of an OFDM symbol.

Assume that the transmitted signal is $s(n)$ and the additive noise $n(n)$ is white Gaussian, then the sampled received signal is

$$y(n) = s(n) * h(n) + n(n) = \sum_{\tau=0}^{L-1} a(\tau)s(n-\tau) + n(n), \tag{2}$$

Here '*' denotes linear convolution.

The correlation between received signal and locally generated PN255 sequence $p(n)$ is:

$$\begin{aligned} R_{py}(n) &= \sum_{i=0}^{M-1} p(i)y(n+i) \\ &= \sum_{i=0}^{M-1} p(i)[\sum_{\tau=0}^{L-1} a(\tau)s(n+i-\tau) + n(n+i)] \\ &= \sum_{\tau=0}^{L-1} a(\tau)\sum_{i=0}^{M-1} p(i)s(n+i-\tau) + \sum_{i=0}^{M-1} p(i)n(n+i) \\ & \qquad n = 0,\dots,M-1 \end{aligned} \tag{3}$$

If the transmitted signal is PN sequence and satisfies:

$$s(n+i-\tau) = p((n+i-\tau)_M), \quad n,i = 1,\dots,M-1; \; \tau = 0,\dots,L-1 \tag{4}$$

M is 255 and $(\cdot)_M$ denotes modular M process. Then equation (3) can be rewritten as:

$$\begin{aligned} R_{py}(n) &= \sum_{\tau=0}^{L-1} a(\tau)\sum_{i=0}^{M-1} p(i)p((n+i-\tau)_M) + R_{pn}(n) \\ &= \sum_{\tau=0}^{L-1} a(\tau)R_{pp}(n-\tau) + R_{pn}(n) \quad , \quad n = 0,\dots,M-1 \end{aligned} \tag{5}$$

$R_{pp}(n)$ is the auto-correlation of PN255 and $R_{pn}(n)$ is cross-correlation between PN255 and noise. If $R_{pp}(n)$ is ideal δ function (impulse function), $R_{py}(n)$ is just the estimated channel impulse response.

However, the auto-correlation of m sequences $R_{pp}(n)$ is not ideal δ but with the following value:

$$R_{pp}(n) = \sum_{i=0}^{M-1} p(i)p((n+i)_M) = \begin{cases} M \; , & n=0 \\ -1 \; , & n \neq 0 \end{cases}, \quad n = 0,\dots,M-1 \tag{6}$$

There is a direct current value with $n\neq 0$ and this will introduce interference between different channel paths. Fortunately, the direct current value (minus one) is relatively small compared with M, and the interference can be removed by using following iterative 'correlation reshape' process:

1) First, we find the biggest peak value $\hat{a}(\tau_0)$ of $R_{py}(n)$, τ_0 is the location of this correlation peak. We can subtract the interference by:

$$R_{py}^{(1)}(n) = R_{py}(n) + \frac{\hat{a}(\tau_0)}{M}, \quad n = 0,\dots,M-1, \; n \neq \tau_0 \tag{7}$$

2) Then, we will find the second biggest peak value $\hat{a}(\tau_1)$ of $R_{py}^{(1)}(n)$, and subtract the interference of $\hat{a}(\tau_1)$ from $R_{py}^{(1)}(n)$ again. In general case, after removing interference of several strongest correlation peak value, we can get satisfactory estimation of channel impulse response (i is the iteration number):

$$\hat{h}(n) = R_{py}^{(i)}(n), \quad n = 0,\dots,L-1 \tag{8}$$

In the above discussion, a very important precondition is Equ.(4), which means a cyclic form of PN sequences. The PN guard interval in DMB-T is in a quasi-cyclic form: as shown in Fig.2, the first L_Cyc samples of PN guard interval can be viewed as a cyclic-prefix of PN255_2. But when channel delay is longer than L_Cyc, the cyclic property is destroyed, so with different channel impulse function length, the channel estimation algorithm should be different.

A. The situation 0≤L< 165

We first discuss the situation when longest channel delay L < L_Cyc=165. As shown in Fig.3, we denote the Pre-amble and Post-amble part as '2' and '1' respectively. 'Symbol_previous' and 'Symbol_current' represent two

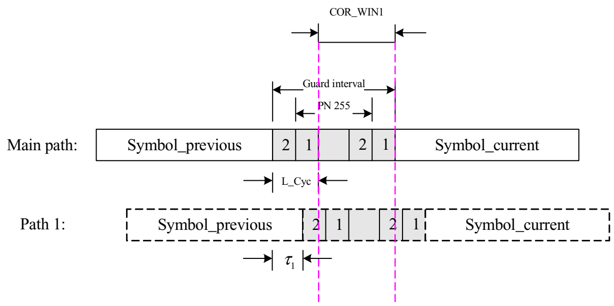

Fig. 3. The situation 0≤ L< 165

consecutive OFDM symbols. Assuming that the system is perfectly synchronized, we take the first path as main path, and Path1 as multipath with delay $0 \le \tau_1 < 165$. The cyclic correlation is performed between PN255_2 and received signal within correlation window 'COR_WIN1':

$$R_{py\_COR\_WIN1}(n) = \sum_{i=0}^{M-1} p_{PN\_255\_2}(i) y((n+i)_M), \quad n = 0,\ldots,M-1 \quad (9)$$

Where $p_{PN\_255\_2}(i)$ represents PN255_2 sequence. Since the condition in Equ.(4) holds, (9) can be rewritten:

$$R_{py\_COR\_WIN1}(n) = \sum_{\tau=0}^{L-1} a(\tau) R_{pp}(n-\tau) + R_{pn}(n), \; n = 0,\ldots,M-1 \quad (10)$$

We can get the channel estimation $\hat{h}(n)$:

$$\hat{h}(n) = Cor\_reshape(R_{py\_COR\_WIN1}(n)), \; n = 0,\ldots,M-1 \quad (11)$$

The function $Cor\_reshape(\cdot)$ represents the correlation reshape process in Eq.(7-8).

B. The situation 165≤L< 255

The situation when the longest channel delay is longer than L_Cyc but shorter than 255 is shown in Fig.4. Path1 and Path2 represent multipaths with respective delay $0 \le \tau_1 < 165$ and $165 \le \tau_2 < 255$.

It can be seen that in correlation window 'COR_WIN1', the

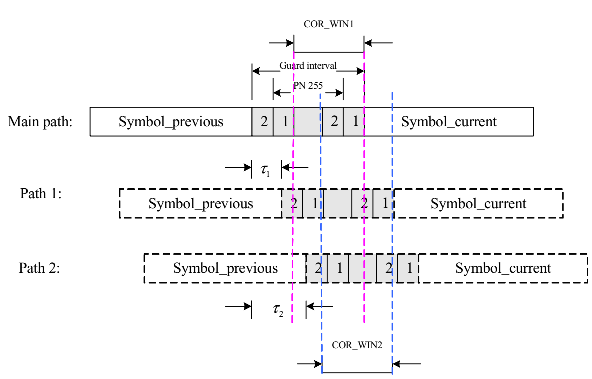

Fig. 4. The situation165≤ L<255

condition in Eq.(4) does not hold, the correlation result will contain interference from the previous OFDM data symbol. Such interference is in the form of random noise and will degrade the accuracy of estimation. In order to obtain proper estimation of channel, we should try to cancel the interference.

The interference cancellation is decision directed. After decision of the OFDM data symbols (in order to avoid large channel decoding delay, hard decision is used here), the decided values are remapped and transformed with IFFT again to regenerate the time domain signal. For convenience, we denote the regenerated time domain signal of 'Symbol_previous' and 'Symbol_current' as '$\hat{s}$_previous' and '$\hat{s}$_current' respectively.

The estimation process is divided into following steps:

1) Assuming that $\hat{h}^{(0)}(n)$, $n=0,\ldots,L-1$, is the initial channel estimation obtained in the period of previous frame, and '$\hat{s}$_previous' is the regenerated time domain signal of previous OFDM data symbol, we first set $\hat{s}$_current = $\boldsymbol{0}$, where $\boldsymbol{0}$ is a $1\times N$ zero vector.
2) define:

$$h_2(n) = \begin{cases} \hat{h}^{(0)}(n) \;, & 165 \le n < L-1 \\ 0 \;, \; 0 \le n < 165 \;\; and \;\; L \le n < 255 \end{cases} \quad (12)$$

and

$$D(n) = \begin{cases} \hat{s}\_\text{previous}, & 0 \le n < N \\ \text{PN guard interval}, & N \le n < N+420 \\ \hat{s}\_\text{current}, & N+420 \le n < 2N+420 \\ 0, & others \end{cases} \quad (13)$$

Here $N = 3780$. We perform convolution between $h_2(n)$ and $D(n)$:

$$C_1(n) = \sum_{\tau=0}^{255-1} h_2(\tau) D(n-\tau), \; n = 0,\ldots,2N+420+254 \quad (14)$$

Then we can subtract path2 from the received signal within window COR_WIN1 by:

$$y^{(1)}(n) = y(n) - C_1(n+N+165), \quad n = 0,\ldots,254 \quad (15)$$

Then, the cyclic correlation is performed within COR_WIN1 and the channel estimation is updated:

$$R_{py\_COR\_WIN1}(n) = \sum_{i=0}^{M-1} p_{PN\_255\_2}(i) y^{(1)}((n+i)_M), \quad (16)$$
$$n = 0,\ldots,M-1$$

$$\hat{h}^{(1)}(n) = \begin{cases} Cor\_reshape(R_{py\_COR\_WIN1}(n)), & 0 \le n < 165 \\ \hat{h}^{(0)}(n), & 165 \le n < L-1 \end{cases} \quad (17)$$

Next, we equalize the channel with $\hat{h}^{(1)}(n)$, make the decision of current OFDM data symbol 'Symbol_current' and regenerate the time domain signal '$\hat{s}$_current' with remapping and IFFT transform.

3) We define:

$$h_1(n) = \begin{cases} \hat{h}^{(1)}(n) \ , & 0 \le n < 165 \\ 0 \ , & 165 \le n < 255 \end{cases} \tag{18}$$

perform convolution between $h_1(n)$ and $D(n)$:

$$C_2(n) = \sum_{\tau=0}^{255-1} h_1(\tau) D(n-\tau), \ n = 0,\dots,2N+420+254 \tag{19}$$

To estimate the channel response longer than L_Cyc, we define a new correlation window 'COR_WIN2', as shown in Fig.4. We subtract Main path and path1 from the received signal within window COR_WIN2 by:

$$y^{(2)}(n) = y(n) - C_2(n+N+255), \quad n = 0,\dots,254 \tag{20}$$

The cyclic correction is performed within COR_WIN2:

$$R_{py\_COR\_WIN2}(n) = \sum_{i=0}^{M-1} p_{PN\_255\_2}(i) y^{(2)}((n+i)_M), \quad n = 0,\dots,M-1 \tag{21}$$

The channel estimation is updated as:

$$\hat{h}^{(2)}(n) = \begin{cases} \hat{h}^{(1)}(n), & 0 \le n < 165 \\ Cor\_reshape(R_{py\_COR\_WIN2}(n-165+\gamma)), & 165 \le n < L-1 \end{cases} \tag{22}$$

Where γ=2×L_Cyc-255=75, is a shift value.

Hereto, the whole channel estimation is updated. We equalize the channel and make hard decision of current OFDM data symbol again. After remapping and transform with IFFT, the '$\hat{s}$_current' is updated. We replace $\hat{h}^{(0)}(n)$ with $\hat{h}^{(2)}(n)$ and repeat step 2), 3). After several iterations (in general case, it only needs 1 or 2 iterations), the whole process converges and we will get the final channel estimation.

4) In step 2) ,3), an initial channel estimation $\hat{h}^{(0)}(n)$ and decision of previous OFDM data symbol '$\hat{s}$_previous' is needed. However, at the beginning of the work, there is no such initial information available. In order to obtain this initial channel estimation $\hat{h}^{(0)}(n)$, we will perform coarse channel estimation.

As we have discussed before, the correlation within COR_WIN1 will suffer the interference from path2. Although there exist interference, the correlation corresponding to path2 can still generate correlation peaks, the correlation gain is at least L_Cyc. The amplitude of interference is relatively small compared to these correlation peaks. We can use the 'leak' technique to partially eliminate the interference. The 'leak' process is defined as:

$$leak(x) = \begin{cases} x, & |x| \ge Tr \\ 0, & |x| < Tr \end{cases} \tag{23}$$

Where Tr is the threshold value and is determined by simulation.

The correlation gain of PN255 sequences is 255. For the path whose delay τ is in the range 165≤ τ<255, the corresponding correlation gain is only 420-τ, so there should be a scale correction process: the amplitude of such correlation peak will be multiplied with a scalar 255/(420-τ).

After that, the correlation result is processed by correlation reshape function and the coarse channel estimation $\hat{h}^{(0)}(n)$ is obtained finally.

With coarse channel estimation, we can make the decision of the previous OFDM data symbol and regenerate time domain signal '$\hat{s}$_previous'.

C. The situation 255≤L< 420

As shown in Fig.5, Path1, Path2, Path3 represent multipaths with respective delay 0≤τ_1<165, 165≤τ_2<255 and 255≤τ_3<420. The correlation within COR_WIN1 will suffer interference from path2 and path3. Since the longest delay τ_3 is longer than 255, which is the size of correlation, this situation becomes much more complex: interference from path3 includes not only noise-like interference but also a parasitical correlation peak on the location τ_3-255 of the correlation result. This parasitical correlation peak will be detected as a multipath which does not exist at all and will introduce larger estimation error than those noise-like interferences.

In order to estimate Path3 accurately, we define another correlation window COR_WIN3, as shown in Fig.5. The correlation within COR_WIN2 will suffer interference from

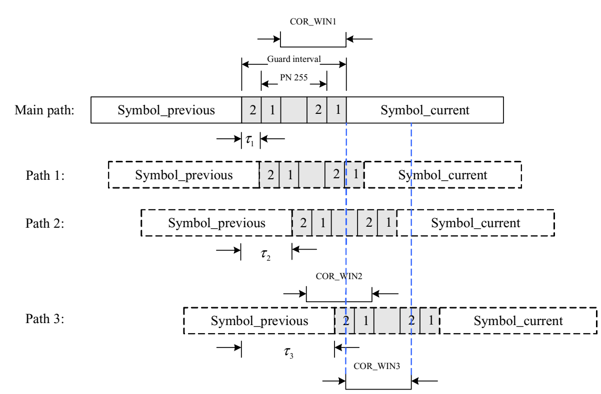

Fig. 5. The situation 255≤ L<420

Main path, Path1 and Path3; the correlation within COR_WIN3 will suffer interference from Main path, Path1 and Path2. We will try to cancel these interferences in different step:

1) First, as in part B, we set $\hat{s}$_current = $\boldsymbol{0}$ and assume there exists initial channel estimation $\hat{h}^{(0)}(n)$ and '$\hat{s}$_previous'.
2) Define:

$$h_{12}(n)=\begin{cases}\hat{h}^{(0)}(n), & 0\le n<255\\ 0, & 255\le n<420\end{cases}$$
$$h_{23}(n)=\begin{cases}\hat{h}^{(0)}(n), & 165\le n<L\\ 0, & 0\le n<165 \ \ and \ \ L\le n<420\end{cases} \tag{24}$$
$$h_{13}(n)=\begin{cases}\hat{h}^{(0)}(n), & 0\le n<165 \ \ and \ \ 255\le n<L\\ 0, & 165\le n<255 \ \ and \ \ L\le n<420\end{cases}$$

The convolution between $h_{23}(n)$ and $D(n)$ is:

$$C_{23}(n)=\sum_{\tau=0}^{255-1}h_{23}(\tau)D(n-\tau),\ \ n=0,...,2N+420 \tag{25}$$

Then Path2 and Path3 are removed from the received signal within COR_WIN1:

$$y^{(1)}(n)=y(n)-C_{23}(n+N+165),\quad n=0,...,254 \tag{26}$$

The correlation is performed within COR_WIN1 and the channel estimation is updated with Eq.(16-17). After equalization, decision of current OFDM symbol is made and '$\hat{s}$_current' is generated.

3) Similar to 2), we can remove the Main Path, Path1 and Path3 from the received signal within COR_WIN2 by Eq(27-28):

$$C_{13}(n)=\sum_{\tau=0}^{255-1}h_{13}(\tau)D(n-\tau),\ \ n=0,...,2N+420 \tag{27}$$
$$y^{(2)}(n)=y(n)-C_{13}(n+N+255),\quad n=0,...,254 \tag{28}$$

After correlation as in Eq.(21), the channel estimation can be updated:

$$\hat{h}^{(2)}(n)=\begin{cases}\hat{h}^{(1)}(n), & 0\le n<165 \ \ and \ \ 255\le n<L\\ Cor\_reshape(R_{py\_COR\_WIN2}(n-165+\gamma)), & 165\le n<255\end{cases} \tag{29}$$

Again, remove Main Path, Path1 and Pathe2 from the received signal within COR_WIN3 by Eq(30-31):

$$C_{12}(n)=\sum_{\tau=0}^{255-1}h_{12}(\tau)D(n-\tau),\ \ n=0,...,2N+420 \tag{30}$$
$$y^{(3)}(n)=y(n)-C_{12}(n+N+420),\quad n=0,...,254 \tag{31}$$

The correlation in COR_WIN3 is performed:

$$R_{py\_COR\_WIN3}(n)=\sum_{i=0}^{M-1}p_{PN\_255\_2}(i)y^{(3)}(n+i),\quad n=0,...,M-1 \tag{32}$$

The channel estimation is updated:

$$\hat{h}^{(3)}(n)=\begin{cases}\hat{h}^{(2)}(n), & 0\le n<255\\ Cor\_reshape(R_{py\_COR\_WIN3}(n-255)), & 255\le n<L\end{cases} \tag{33}$$

After the whole channel impulse function is updated, OFDM data symbol is equalized and decided again. Then we replace $\hat{h}^{(0)}(n)$ with $\hat{h}^{(3)}(n)$ and repeat step 2-3).

4) At the beginning of the work, no initial channel estimation $\hat{h}^{(0)}(n)$ and regenerated time domain signal '$\hat{s}$_previous' available. So coarse channel estimation should be obtained.
Such coarse channel estimation is mainly based on correlation within window COR_WIN1 and COR_WIN3. The correlation result in COR_WIN1 will suffer the interference from Path2 and Path3. As we discussed in part B, noise-like interference can be partially eliminated by using 'leak' process, but the interference from Path3 also contain a parasitical correlation peak, which should be further removed by using the correlation result of COR_WIN3.
Without removing Main path, Path1 and Path2 from the received signal, the correlation result of COR_WIN3 will contain serious interference. However, such interference is noise-like. By using 'leak' technique, we can at least detect the large correlation peaks which represent the strong multipaths with delay between 255 and 420 samples. Then, the coarse channel estimation can be expressed as:

$$\hat{h}^{(0)}(n)=\begin{cases}Z_{COR\_WIN1}(n)-Z_{COR\_WIN3}(n)\dfrac{165-n}{255}, & 0\le n<165\\ Z_{COR\_WIN1}(n)\dfrac{255}{420-n}, & 165\le n<255\\ Z_{COR\_WIN3}(n-255), & 255\le n<420\end{cases} \tag{34}$$

Where

$$Z_{COR\_WIN1}(n)=Cor\_reshape(leak(R_{COR\_WIN1}(n)));$$
$$Z_{COR\_WIN3}(n)=Cor\_reshape(leak(R_{COR\_WIN3}(n))); \tag{35}$$

$R_{COR\_WIN1}(n)$ and $R_{COR\_WIN3}(n)$ represent correlation result within COR_WIN1 and COR_WIN3 respectively.

IV. Channel Equalization for DMB-T

With the estimated channel information, in this section, we will talk about the equalization technique for DMB-T system.

Since there is no cyclic prefix in DMB-T data frame, conventional one-tap frequency equalizer can not be directly applied to DMB-T receivers. In this section, we propose a tail cancellation and cyclic restoration (TCCR) technique, which can reconstruct the cyclic convolution relationship between Frame body and channel impulse function. The process of TCCR is shown in Fig.6.

In Fig.6, G1 and G2 represent two consecutive guard intervals. $\hat{h}$ is the estimated channel impulse response. We first make linear convolution between guard interval and $\hat{h}$. The convolution result G has the length of 420+L-1, we then perform calculations as:

$$\begin{aligned} &\mathrm{X}_1(m) = \text{Frame body}(m) - \mathrm{G}(m+420) \\ &\mathrm{X}_2(m) = \mathrm{G2}(m) - \mathrm{G}(m), \qquad m = 0,\ldots,L-2 \\ &\mathrm{X}(n) = \begin{cases} \mathrm{X}_1(n) + \mathrm{X}_2(n), & 0 \le n < L-1 \\ \text{Frame body}(n), & L-1 \le n < N \end{cases} \end{aligned} \tag{36}$$

X(n) is the OFDM data symbol after TCCR. Then the one-tap frequency equalization can be used:

$$S = \mathrm{FFT}(\mathrm{X}) / \mathrm{FFT}(\hat{h}) \tag{37}$$

S is the equalized data symbol in frequency domain.

Since there is only ±1 in PN guard interval, the linear convolution and processes in Eq.(36) only involve addition and subtraction calculations. So TCCR can be easily implemented.

V. SIMULATION RESULTS

Computational simulation is performed to verify above channel estimation and equalization algorithms. We will compare the performance between DMB-T system and DVB-T system under wireless frequency selective environment. Some simulation parameters of two systems are shown in Table.I.

The longest multipath delay that DMB-T can handle is 55.556 µs. In order to combat same channel delay, the cyclic prefix of DVB-T should be at least 1/4 and 1/16 in 2K and 8K mode respectively.

We chose COST207 Typical Urban (TU) channel model and China Test 8th (CT8) channel model in our simulation. The profiles of these two channel model is shown in Table.II and Table.III. The longest delay of TU channel and CT8 channel is 5 µs and 31.8µs respectively. Besides, when in single frequency network (SFN) application, there will be strong multipath with much longer delay. In order to test such situation, we define another SFN channel in our simulation.

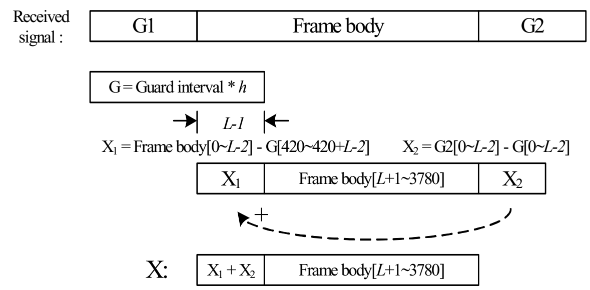

Fig. 6. The process of tail cancellation and cyclic restoration(TCCR)

TABLE I
THE PARAMETERS OF SIMULATION SYSTEM

	DMB-T	DVB-T 2K	DVB-T 8K
Subcarrier number	3780	2048	8192
Subcarrier for Data	3780	1529	6116
Training Subcarrier or symbol	420	176	700
Sample period (µs)	0.1323	0.1094	0.1094
Bandwidth (MHz)	7.56	7.61	7.61
Cyclic Prefix Ratio	0	1/4	1/16
Transmission efficiency	0.9	0.673	0.841

TABLE II
THE PROFILE OF TYPICAL URBAN (TU) CHANNEL

	Tap1	Tap2	Tap3	Tap4	Tap5	Tap6	unit
Delay	0	0.2	0.5	1.6	2.3	5	µs
Power	-3	0	-5	-6	-8	-10	dB

TABLE III
THE PROFILE OF CHINA TEST 8 (CT8) CHANNEL

	Tap1	Tap2	Tap3	Tap4	Tap5	Tap6	unit
Delay	0	-1.8	0.15	1.8	5.7	30	µs
Power	0	-18	-20	-20	-10	0	dB

TABLE IV
THE PROFILE OF SFN CHANNEL

	Tap1	Tap2	Tap3	unit
Delay	0	19	52	µs
Power	0	-5	-1	dB

The profile of our defined SFN channel is shown in Table.IV. DVB-T channel estimation is based on pilot. In our simulation, least squared (LS) criterion and cubic interpolation is used [8-9]. Since there are many different channel estimation algorithms for DVB-T, we can only give a rough comparison in this paper.

The mean squared errors (MSE) of channel estimation under different signal to noise ratio (SNR) are shown in Fig.7-9, for each channel respectively. For further comparison, the performance of bit error rates (BER) before channel decoding for 64QAM are also given out in Fig.10-12.

It is shown that the performances of DMB-T and DVB-T are almost same in TU and SFN channel. While in CT8 channel, DMB-T works better than DVB-T 2K mode but worse than DVB-T 8K mode. This phenomenon is reasonable:

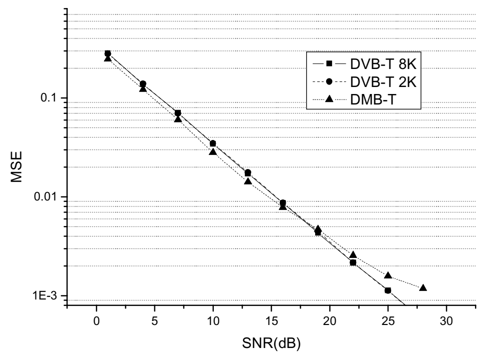

Fig.7. The MSE performance under TU channel

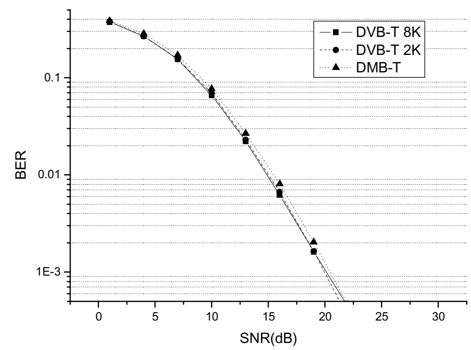

Fig.10. The BER performance under TU channel (64QAM before channel decoding)

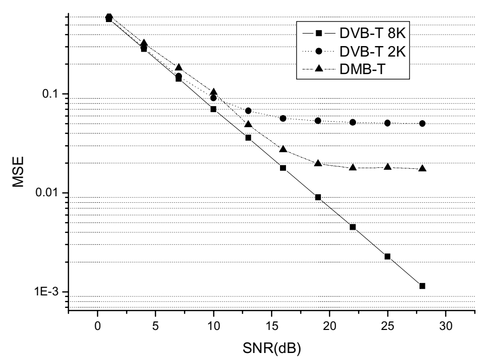

Fig. 8. The MSE performance under CT8 channel

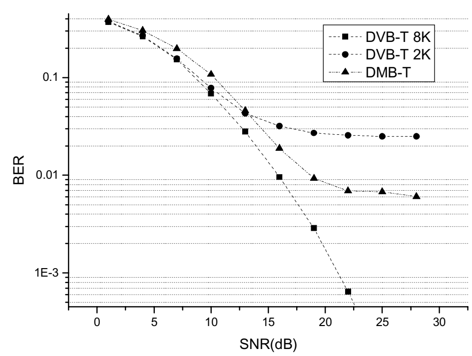

Fig. 11. The BER performance under CT8 channel (64QAM before channel decoding)

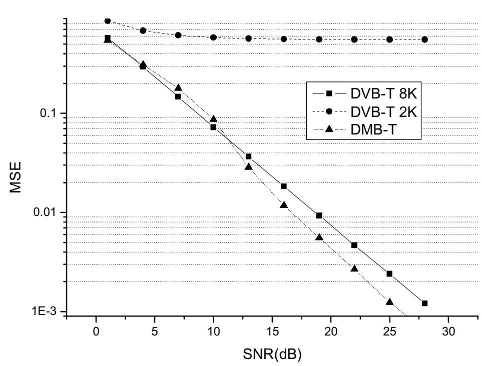

Fig. 9. The MSE performance under SFN channel

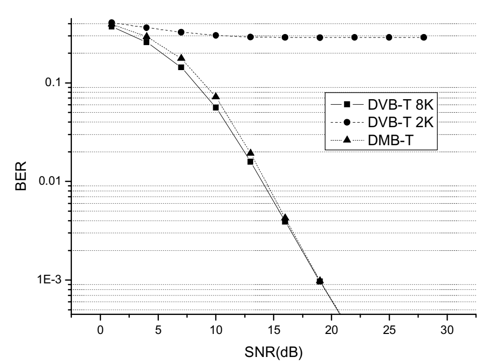

Fig. 12. The BER performance under SFN channel (64QAM before channel decoding)

there are much more subcarriers in DVB-T 8K mode, so DVB-T 8K mode has relatively higher pilot's frequency domain resolution and consequently has better estimation accuracy than in DMB-T and DVB-T 2K mode.

DMB-T frame structure is in some sense a trade off between DVB-T 2K and 8K mode. Although DVB-T 8K mode has better ability to combat the frequency selective channel, it is much more sensitive to the time selective property of the channel, such as frequency offset, phase noise and Doppler shift. With help of powerful foreword error control encoding techniques such as TURBO code and LDPC code, our channel estimation and equalization method for DMB-T can provide satisfactory system performances under all above three channels.

VI. CONCLUSION

In This paper, we propose a channel estimation and equalization method for DMB-T, a TDS-OFDM based terrestrial HDTV broadcasting system. Our proposed channel estimation algorithm is based on time domain correlation and iterative interference cancellation techniques. In order to equalize the channel with simple one-tap frequency domain equalizer, a tail cancellation and cyclic restoration (TCCR) technique is used. Simulation results show that such method can provide satisfactory performance for DMB-T system.

REFERENCES

[1] J. Wang; Z.X. Yang; C.Y. Pan; M. Han; L. Yang, "A combined code acquisition and symbol timing recovery method for TDS-OFDM," *IEEE Trans. Broadcasting,* Vol. 49, pp. 304-308, Sept. 2003.

[2] D. Huang; L. Yang; H. Xing; Z. Zhou, "Multiplexing guard intervals and time domain pilots in OFDM systems", *Consumer Electronics, 2001. ICCE. International Conference on*, pp. 68-69, June 2001.

[3] Z.X. Yang; L. Tong; L. Yang, "Outage probability comparison of CP-OFDM and TDS-OFDM for broadcast channels", *IEEE Global Telecommunications Conference, 2002. GLOBECOM '02.*, Vol. 1, pp. 594-598, Nov. 2002.

[4] Z.W. Zheng; Z.X. Yang; C.Y. Pan; Y.S. Zhu, "Robust phase noise suppression scheme for TDS-OFDM-based digital television terrestrial broadcasting systems", *IEEE Trans. Consumer Electronics*, Vol. 50, pp. 436-442, May. 2004.

[5] Z.W. Zheng; Z.X. Yang; C. Y. Pan; Y.S. Zhu, "Synchronization and channel estimation for TDS-OFDM systems", *IEEE 58th Vehicular Technology Conference. VTC 2003-Fall*, Vol. 2, pp. 1229-1233, Oct. 2003.

[6] "Terrestrial Digital Multimedia/Television Broadcasting System," P.R. China Patent 00 123 597.4 filed Aug.25, 2000, issued Mar. 21,2001.

[7] ETSI, "Digital Video Broadcasting: Framing Structure, Channel coding, and Modulation for Digital Terrestrial Television," European Telecommunication Standard EN300744, Aug. 1997.

[8] S. Coleri, M. Ergen, A. Puri and A. Bahai, "Channel Estimation Techniques Based on Pilot Arrangement in OFDM Systems," *IEEE Trans. Broadcasting,* Vol.48, No. 3, September 2002.

[9] M.H. Hsieh and C.H. Wei, "Channel Estimation for OFDM systems Based on Comb-Type Pilot Arrangement in Frequency selective Fading Channels," *IEEE Trans. Consumer Electronics,* vol. 44, no. 1, Feb. 1998

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