Optimal Sampling Phase Criterion for TDS - OFDM

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Abstract: **In time-domain synchronous OFDM (TDS-OFDM) system for digital television terrestrial multimedia broadcasting (DTMB) standard, the baseband OFDM signal is up sampled and shaping filtered by square root raised cosine (SRRC) filter before digital-to-analog converter (DAC). Much of the work in the area of timing synchronization for TDS-OFDM focuses on frame synchronization and sampling clock frequency offset recovery, which does not consider the sampling clock phase offset due to the up sampling and SRRC filter. This paper evaluates the bit-error-rate (BER) effect of sampling clock phase offset in TDS-OFDM system. First, we provide the BER for M-order quadrature amplitude modulation (M-QAM) in encoded TDSOFDM system. Second, under the condition of the optimal BER criterion and additive white Gaussian noise (AWGN) channel, we propose a near optimal sampling phase estimation criterion based on frequency-domain response. Simulations demonstrate that the proposed criterion also has good performance in actual TDS-OFDM system with channel coding over multipath channels, and it is superior to the conventional symbol timing recovery methods for TDS-OFDM system.**

Keywords: **time-domain synchronous OFDM (TDSOFDM), timing synchronization, square root raised cosine (SRRC) filter, bit-error-rate (BER).**

I.   INTRODUCTION

Time-domain synchronous OFDM (TDS-OFDM) has superior performance in terms of fast synchronization, accurate channel estimation and higher spectral efficiency compared with other OFDM solutions, and it has been adopted by the digital television terrestrial multimedia broadcasting (DTMB) standard [1], [2]. In TDS-OFDM system, the baseband OFDM signal is up sampled and shaping filtered by the square root raised cosine (SRRC) filter before digital-to-analog converter (DAC) [3]. In this way, the spectrum outside the band is effectively suppressed, the inter-symbol-interference within an OFDM data block is degraded, and the correlation based frame synchronization can be robust to carrier frequency offset (CFO) and multipath channels [4]–[7]. Previous work in the area of timing synchronization for TDS-OFDM focuses on frame synchronization and sampling clock frequency offset correction. [4] Proposed a symbol timing recovery (STR) method based on code acquisition (CA) to obtain frame synchronization and track the sampling clock frequency offset at the receiver. However, this CA based frame synchronization suffers from obvious performance loss when large CFO exists. Hence [5]–[8] proposed robust frame synchronization methods for TDS-OFDM. However, the STR method proposed in [4]–[8] cannot obtain the optimal sampling clock phase over multipath channels, which will be discussed in this paper. In respect of timing synchronization in other communication systems, [9] investigated the effect of frame synchronization error in general OFDM systems. [10] Proposed a pilot aided sampling frequency offset recovery method in cyclic prefix OFDM (CP-OFDM). [11] And [12] examined the effect of clock jitter in cooperative space-time coding multiple input single output systems (MISO). In contrast, to the best of our knowledge, this is the first paper to investigate the impact of sampling clock phase offset on system performance for TDS-OFDM system, which is superior to other OFDM solutions and has been adopted by DTMB standard. In this paper, we evaluate the effect of sampling clock phase offset owing to up sampling and SRRC filter shaping in the TDS-OFDM system after the perfect frame synchronization, sampling clock frequency offset recovery, and CFO elimination. Meanwhile, we also propose a near optimal sampling phase estimation criterion based on frequency domain response, which is different from the

conventional time-domain based timing synchronization methods for TDSOFDM. This paper focuses on three problems as follows. Whether the sampling phase offset does have a great influence on BER performance or not if yes, is there an optimal or near optimal criterion to solve this problem? Compared with the conventional synchronization methods, how much performance gain can be achieved by the proposed criterion? The rest of the paper is organized as follows. In Section II, we introduce the baseband model of DTMB system and the conventional synchronization methods for TDS OFDM. Meanwhile, the effect of sampling phase offset in TDS-OFDM is presented. In Section III, we provide the BER of encoded TDS-OFDM system and propose a near optimal sampling
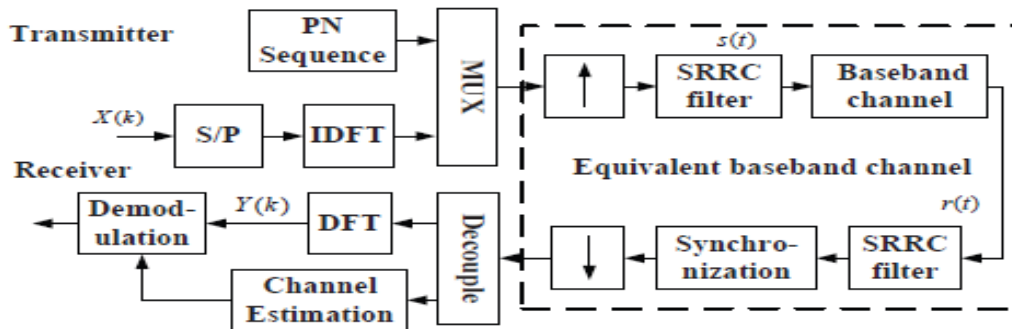

Fig.1. Baseband transceiver of the TDS-OFDM system,

Phase estimation criterion. In Section IV, simulation results are provided. In Section V, conclusions are drawn.

II.   SYSTEM MODEL OF TDS-OFDM BASED DTMB

The baseband transceiver of the TDS-OFDM system [3] is shown in Fig. 1. In the time domain, a TDS-OFDM symbol consists of a pseudo-noise (PN) sequence and the following OFDM data block. The PN sequence, serving as the guard interval, is inserted between the adjacent OFDM data blocks to eliminate the inter-block interference (IBI) over multipath channels. The OFDM data block is generated by inverse discrete Fourier transform (IDFT) of frequency domain data. Both the PN sequence and the OFDM data block share the same symbol rate $f_{sym} = 1/T_{sym} = 7.56$MHz. After multiplexing, the TDS-OFDM signal is processed by $N_{upsam} = 4$ times up sampling and SRRC filter shaping, thus the signal sampling rate becomes $f_{upsam} = 30.24$MHz. Finally, the signal is sent to DAC. At the receiver, the baseband sampling rate of the analog to digital converter (ADC) is $f_s = 1/T_s = 30.40$MHz, which is slightly higher than f_{upsam} to reduce the baseband signal information loss in the absence of synchronization [13], [14]. The synchronization module is aimed at frame synchronization, CFO elimination and the sampling clock frequency recovery. Sequentially, signal after synchronization is down sampled, and then PN and OFDM data block are decoupled, whereby PN is used for channel estimation and OFDM data block is sent to equalization and demodulation. The synchronization module consists of three parts: frame synchronization, CFO correction and sampling frequency offset recovery. [4] Proposed a STR method based on CA for TDS-OFDM. In CA stage, this method searches and tracks the correlation peak $R(k, e)$ of the PN sequences embedded in the signals to obtain the frame synchronization, where k is the time index of the correlation peak and e is the timing
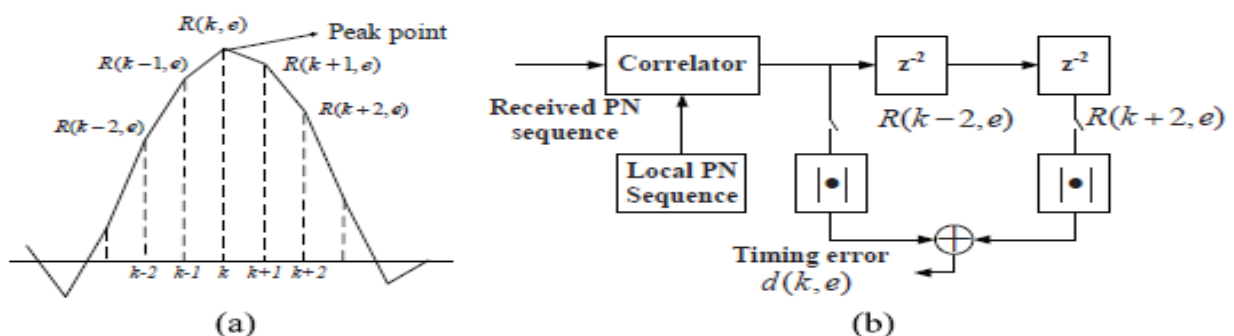

Fig.2. the conventional STR method in TDS-OFDM: (a) time-domain correlation; (b) timing error detector.
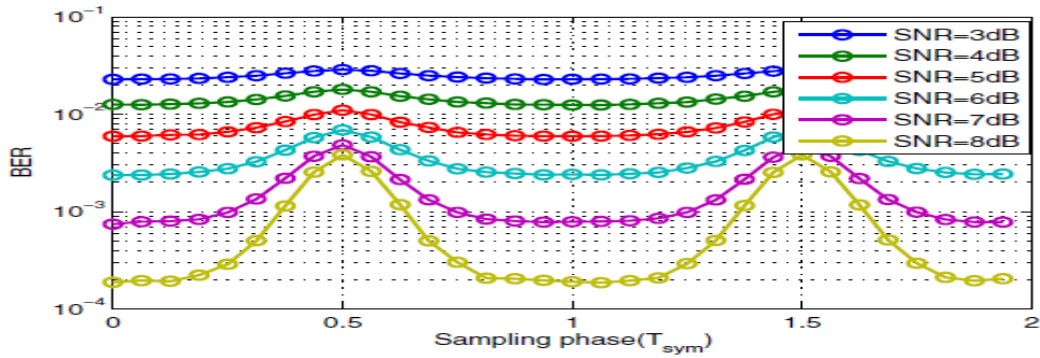
Fig.3. BER against different sampling phase offsets for TDS-OFDM using encoded BPSK modulation over AWGN channel.

Error as shown in Fig.2. (A). in this way, the timing error is reduced to no more than Tsym⁄2. Sequentially, the STR algorithm eliminates the residual timing error e by a STR feedback loop, which consists of a timing error detector, a loop filter, and a digital interpolator. The interpolator driven by the timing error signal is used to recover the received signal, and the loop filter is used to normalize the timing error signal and enhance its robustness to noise. Timing error signal is produced by the amplitude difference of adjacent side lobes of the acquired maximum correlation peak, as shown in Fig. 2 (b). Consequently, signal after the interpolator is adjusted to sampling frequency $f$upsam = 30.24MHz by a decimator. Nevertheless, the STR method based on the time-domain PN correlation aims at tracking the sampling clock frequency offset, and it cannot obtain the optimal sampling clock phase offset over multipath channels. Therefore, some questions appear. Does the sampling clock phase offset have a great influence on the BER performance? If yes, is there an optimal or near optimal criterion to solve this problem? Compared with the conventional STR methods, how much performance gain can be achieved by the proposed criterion? For the first problem, we provide the BER of TDS-OFDM system over additive white Gaussian noise (AWGN) channel with different sampling phase offsets, as shown in Fig. 3. Simulation assumes perfect frame synchronization; sampling clock frequency offset recovery, and CFO elimination. OFDM data block adopts encoded binary phase shift keying (BPSK) modulation with DFT length N = 4096, Nupsam = 4, and SRRC roll-off factor α = 0.05. From Fig. 3, we observe that the BER performance of different sampling phase offsets within a Tsym varies largely, and it appears Tsym periodicity. Even the optimal sampling phase offset is superior to the worst situation by about 3 dB performance gain.

III.   THE PROPOSED SAMPLING PHASE OFFSET CRITERION

In this section, we provide the BER in encoded TDS-OFDM system, and propose a near optimal sampling phase estimation criterion.
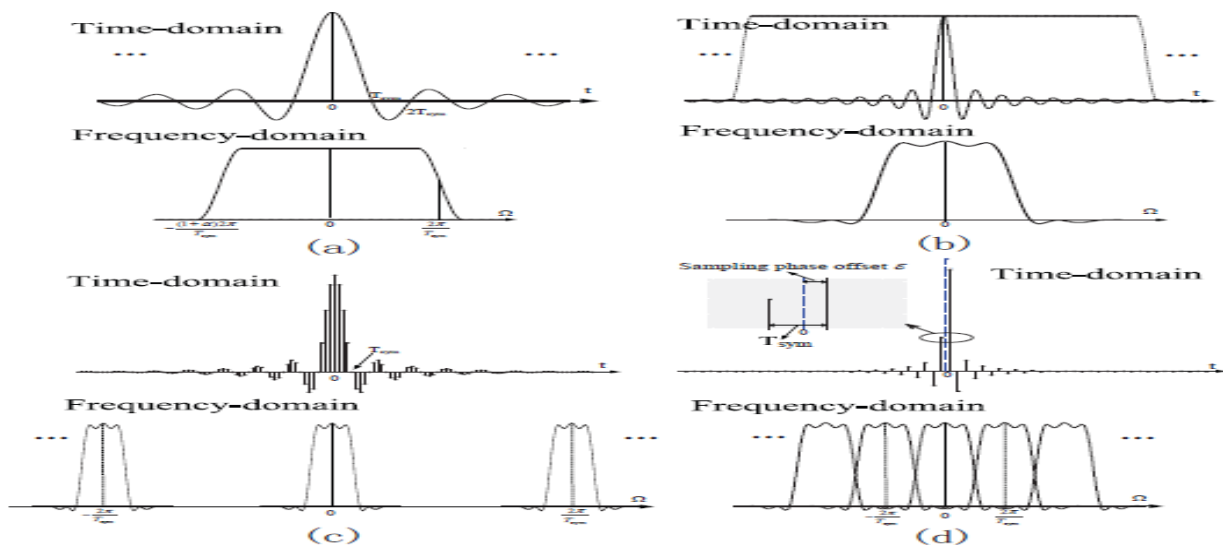

Fig. 4. Frequency-domain response and the corresponding time-domain response of the equivalent baseband channel. For convenience, we consider the AWGN channel, i.e. $HC(\Omega) = 1$. (a) $HC(\Omega)H$SRRC(Ω); (b) Hfinite(Ω); (c) $H\_(\Omega)$; (d) $H(\Omega)$.

A. *Frequency-Domain Response of The Equivalent Baseband Channel* In Fig. 1, we consider modules in the dashed box as the equivalent baseband channel. We assume that signal after synchronization module achieves perfect frame synchronization, sampling clock frequency offset recovery, and CFO elimination. The analog frequency-domain response of SRRC filters, including the transmit and receive shaping filters, is denoted as HSRRC (Ω), which can be expressed as

$$H_{\text{SRRC}}(\Omega) = \begin{cases} T_{\text{sym}}, & 0 \leq |\Omega| < \frac{(1-\alpha)\pi}{T_{\text{sym}}}, \\ \frac{T_{\text{sym}}}{2}[1 + \sin(\frac{T_{\text{sym}}}{2\alpha}(\frac{\pi}{T_{\text{sym}}} - \Omega))], \\ & \frac{(1-\alpha)\pi}{T_{\text{sym}}} \leq |\Omega| < \frac{(1+\alpha)\pi}{T_{\text{sym}}}, \\ 0, & |\Omega| > \frac{(1+\alpha)\pi}{T_{\text{sym}}}. \end{cases} \quad (1)$$

The analog frequency-domain response of baseband channel is denoted as $HC(\Omega)$. Fig.4. (a) Illustrates the frequency-domain and its corresponding time-domain response of $HC(\Omega)H$SRRC(Ω) under AWGN channel. In practice, the length of SRRC filters at the transmitter and receiver is finite, which means an equivalent time-domain rectangle windowing on the response, as shown in Fig. 4. (b). Since the window length is usually very long, the effect of equivalent time-domain windowing can be negligible. Consequently, the frequency-domain response after windowing is Hfinite(Ω) ≈ HSRRC$(\Omega)HC(\Omega)$. Next, the impact of up sampling at the transmitter on Hfinite(Ω) is that the spectrum becomes periodical and com-
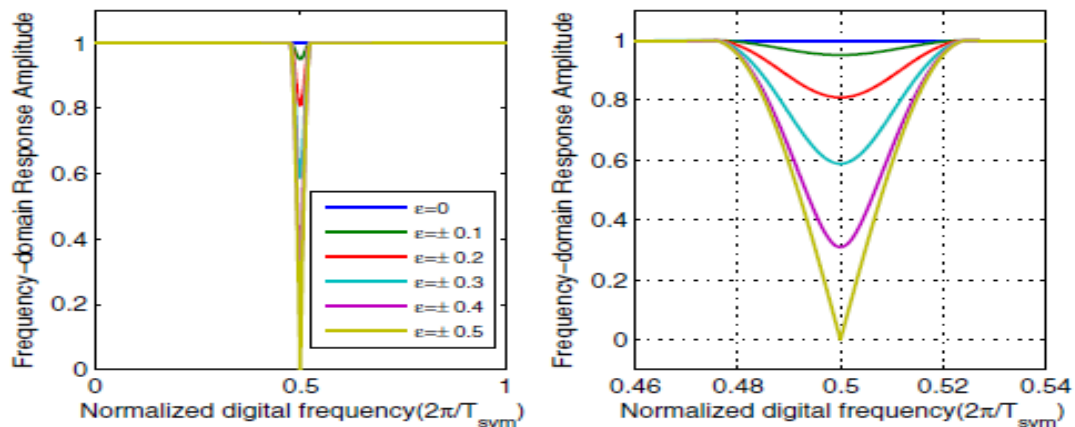

Fig.5. Frequency-domain response amplitude of $H(\Omega)$ with different sampling phase offsets. Here, we consider the AWGN channel, i.e. $HC(\Omega) = 1$, and $\alpha = 0.05$. The right figure is the details of the left figure in $f \in [0.46, 0.54]$. Pressed as illustrated in Fig.4. (c). the spectrum after up sampling can be expressed as

$$H'(\Omega) = \sum_{k=-\infty}^{+\infty} \frac{H_{\text{finite}}(\Omega - \frac{2\pi k}{T_{\text{sym}}})}{N_{\text{upsam}}T_{\text{sym}}} \approx$$
$$\sum_{k=-\infty}^{+\infty} \frac{H_C(N_{\text{upsam}}(\Omega - \frac{2\pi k}{T_{\text{sym}}}))H_{\text{SRRC}}(N_{\text{upsam}}(\Omega - \frac{2\pi k}{T_{\text{sym}}}))}{N_{\text{upsam}}T_{\text{sym}}}. \quad (2)$$

Finally, the influence of down sampling at the receiver on $H(\Omega)$ is spectrum aliasing, as illustrated in Fig. 4. (d). and the frequency-domain response of final equivalent baseband channel can be written as

$$H(\Omega) \approx$$
$$\sum_{k=-\infty}^{+\infty} \frac{H_C(\Omega - \frac{2\pi k}{T_{\text{sym}}})H_{\text{SRRC}}(\Omega - \frac{2\pi k}{T_{\text{sym}}})e^{j(\Omega - \frac{2\pi k}{T_{\text{sym}}})\varepsilon T_{\text{sym}}}}{T_{\text{sym}}},$$
$$(3)$$

Where $\varepsilon \in [-0.5, 0.5]$ is normalized sampling phase offset. Since $H(\Omega)$ appears Tsym periodical, we only need investigate $H(\Omega)$ of $\Omega \in [0, 2\pi/T$sym$]$ or $f \in [0, 1]$, where $f = \Omega T$sym 2π is the normalized digital frequency. Fig. 5 provides the frequency-domain response amplitude of $H(\Omega)$ with different sampling phase offsets, where $HC(\Omega) = 1$. It can be observed that with the sampling phase offset increasing, $/H(\Omega)/$ near $f = 0.5$ rapidly decreases. Obviously, sampling phase offset ε has a great influence on $/H(\Omega)/$. *B. BER and the Proposed Sampling Phase Criterion.* According to [15], the encoded symbol-error-rate (SER) of M-order quadrature amplitude modulation (M-QAM) over AWGN channel is

$$P_s = \frac{2(\kappa-1)}{\kappa}Q\left(\sqrt{\frac{6\log_2(\kappa)}{\kappa^2-1}\left(\frac{E_b}{N_0}\right)}\right)$$

where $\kappa = \sqrt{M}$, E_b is the 1 bit energy, N_0 is unilateral power spectral density of AWGN, and $Q(\bullet)$ is the tail probability of the standard normal distribution, i.e.

$$Q(x) = \frac{1}{\sqrt{2\pi}}\int_x^\infty e^{-\frac{u^2}{2}}du.$$

However, this SER equation is confined to AWGN channel. In fading channels with channel gain K_0, the received 1 bit energy is $K_0 E_b$. Therefore, in the case of OFDM with MQAM over fading channel, the SER of the ith subcarrier is

$$P_{s,i} = \frac{2(\kappa-1)}{\kappa}Q\left(\sqrt{\frac{6\log_2(\kappa)}{\kappa^2-1}|H_i|^2\left(\frac{E_b}{N_0}\right)}\right),$$

where H_i is the frequency domain response of the ith subcarrier.

In terms of OFDM data block with DFT length N and encoded M-QAM, SER can be expressed as

$$P_s = \frac{1}{N}\sum_{n=0}^{N-1}\frac{2(\kappa-1)}{\kappa}Q\left(\sqrt{|H(nf_0)|^2\frac{6\log_2(\kappa)}{\kappa^2-1}\left(\frac{E_b}{N_0}\right)}\right), \quad (4)$$

where $f_0 = 1/N$.

If M-QAM adopts Gray map, BER can be approximated as

$$P_e \approx P_s/\log_2(\kappa). \tag{5}$$

From (3)-(5), it is clear that $/H(f)/$ 2 has a great influence on SER or BER, and $H(f)$ can also be written as $H(f; \varepsilon)$. Therefore, the optimal sampling phase ε is to meet the minimum BER, i.e.

$$\varepsilon_{\text{opt}} = \arg\min_\varepsilon(P_e). \tag{6}$$

Obviously, to obtain εopt from (6) is very difficult. Hence, we use Chernoff Bound [15], i.e.

$$P_e \leq \frac{1}{N}\sum_{n=0}^{N-1}\lambda\exp(-|H(nf_0;\varepsilon)|^2\eta), \tag{7}$$

Where- $\lambda = \frac{2(\kappa-1)}{\kappa}$ and $\eta = \frac{6\log_2(\kappa)}{\kappa^2-1}\frac{E_b}{N_0}$

We consider that in AWGN channel, according to (1) and (3), $H(\Omega)$ can be written as

$$H(f) \approx H_{\text{SRRC}}(f) =$$
$$\begin{cases} e^{j2\pi\varepsilon f}, & 0 \leq f < 0.5(1-\alpha), \\ e^{j2\pi\varepsilon(f-0.5)}[\cos(\pi\varepsilon) + \sin(\frac{\pi}{\alpha}(0.5-f))\sin(\pi\varepsilon)j], \\ & 0.5(1-\alpha) \leq f < 0.5(1+\alpha), \\ e^{j2\pi\varepsilon(f-1)}, & 0.5(1+\alpha) \leq f < 1, \end{cases}$$
$$(8)$$

From (8), it is obvious that the sampling phase offset ε affects BER by affecting the $|H(\Omega; \varepsilon)|^2$ of $\Omega \in [0.5(1 - \alpha), 0.5(1 + \alpha)]$. Thus a sampling phase criterion in AWGN channel can be acquired based on (7)

$$
\begin{aligned}
\varepsilon_{\text{AWGN}} &= \arg\min_{\varepsilon}\Big(\sum_{n=\lceil 0.5N(1-\alpha)\rceil}^{\lfloor 0.5N(1+\alpha)\rfloor} \exp(-|H(nf_0)|^2\eta)\Big) \\
&= \arg\min_{\varepsilon}\Big(\sum_{n=\lceil 0.5N(1-\alpha)\rceil}^{\lfloor 0.5N(1+\alpha)\rfloor} \exp[-(\cos{(\pi\varepsilon)}^2 \\
&\quad + \sin{(\pi\varepsilon)}^2 \sin{\big(\tfrac{\pi}{\alpha}(0.5 - nf_0)\big)}^2)\eta]\Big),
\end{aligned}
\tag{9}
$$

Where $\lceil\cdot\rceil$ and $\lfloor\cdot\rfloor$ are integer ceiling and floor operators, respectively.

Furthermore, obtains the maximum value with $\varepsilon = \varepsilon\text{AWGN} = 0$, which is because the value of the first-order partial derivative $\pi \sin{(2\pi\varepsilon)} [\sin(\pi/\alpha(0.5 - f))^2 - 1]$ is positive when $\varepsilon < 0$, and negative when $\varepsilon > 0$. Therefore, (9) is optimal over AWGN channel and it can be also expressed by

$$
\begin{aligned}
\varepsilon_{\text{AWGN}} = \arg\max_{\varepsilon}\Big(\sum_{n=\lceil 0.5N(1-\alpha)\rceil}^{\lfloor 0.5N(1+\alpha)\rfloor} \cos{(\pi\varepsilon)}^2 + \\
\sin{(\pi\varepsilon)}^2 \sin(\tfrac{\pi}{\alpha}(0.5 - nf_0)^2).
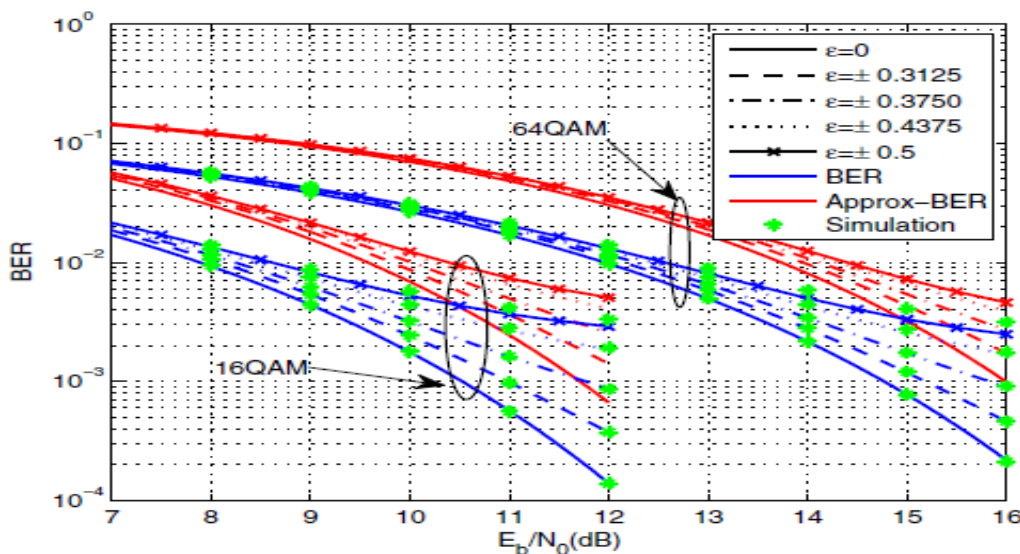\end{aligned}
\tag{10}
$$

Fig.6. BER of theoretical analysis, approximated BER, and BER of simulation with different sampling phase offsets over AWGN channel. Encoded 16QAM and 64QAM are used for OFDM data symbols. Compared with (6) and (9), (10) is more feasible, which only needs the sum of partial squared channel frequency domain gains. Moreover, this sampling phase criterion can also be extended to all channel conditions, i.e.

$$
\varepsilon_{\text{general}} = \arg\max_{\varepsilon}\Big(\sum_{n=\lceil 0.5N(1-\alpha)\rceil}^{\lfloor 0.5N(1+\alpha)\rfloor} |H(nf_0)|^2\Big).
\tag{11}
$$

(11) may not be optimal over multipath channels, but it is near optimal and its validity will be demonstrated in Section IV. In a sense, the correction of the sampling phase offset is a more fine synchronization operation. Therefore, this correction is implemented after the receiver achieves the perfect frame synchronization, CFO correction and sampling frequency offset elimination. Consequently, in practical application, the sampling phase correction module should be cascaded following the synchronization module shown in Fig. 1.

IV.   SIMULATION RESULTS

This section investigates the performances of the proposed sampling phase offset estimation criterion and the conventional STR method [4]. Simulations assume perfect frame synchronization; sampling clock frequency offset recovery, and CFO elimination. Dual PN-OFDM (DPN-OFDM) is adopted for TDS-OFDM transmission with OFDM data block length N = 4096, and PN sequence length L = 512. Up sampling factor and SRRC roll-off factor are the same with DTMB system, i.e. Nupsam = 4, α = 0.05. Additionally, the Brazil digital television field test 4th (Brazil-B) and 5th (Brazil-E) channel models [16] are selected. Fig. 6 shows the BERs of theoretical analysis based on (4), (5), (8) and the BERs of simulation with different sampling phase offsets (ε = 0, $\pm$0.3125, $\pm$0.3750, $\pm$0.4375, $\pm$0.5) over AWGN channel. Encoded 16QAM and 64QAM are used for OFDM data symbols. Additionally, the approximated BERs using Chernoff Bound are also plotted for comparison, denoted as Approx-BER. This figure strongly verifies the validity of the BER of theoretical analysis in Section III. The performance
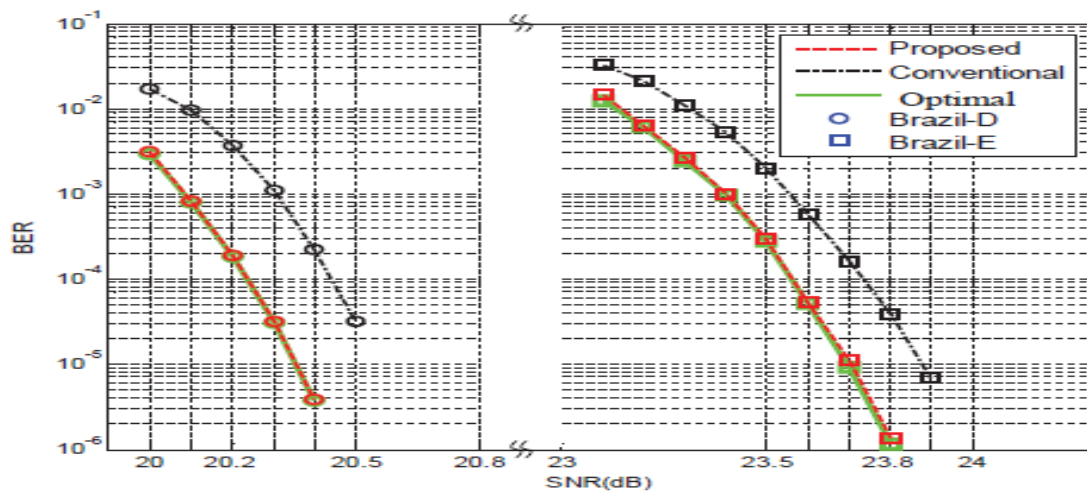

Fig. 7. BER performance comparison when 256QAM is adopted over multipath channels.

curves obtained via the theoretical approach are in good agreement with that of the Monte Carlo simulation results. Meanwhile, we also observe that Approx-BER is a good approximation of BER, since curves of Approx-BER are the horizontal axis shift versions of curves of BER. Therefore, these powerfully support the rationality of the sampling phase

Estimation criterion derived from the approximated BER using Chernoff Bound. In Fig. 6, it is obvious that, with the sampling phase offset increasing, the BER performance degrades rapidly. Here, the target BER of 3×10^{-3} is considered. To achieve the target BER, the best BER performance is superior to the worst situation by 2.5dB performance gain. Moreover, with Eb/N0 increasing, the BER performance differences of different sampling phases increase. Furthermore, from the BER curve tendency of different sampling phases, it can be observed that with the sampling phase increases, the BER floor phenomenon is more obvious. Fig. 7 compares the low-density parity check (LDPC) coded BER performance of the proposed criterion and the conventional STR method, where 256QAM is adopted and LDPC code rate is 0.6. Additionally, we adopts the grid search method to approach the BER of the optimal sampling phase. In this method, we divide the sampling period Tsym into 128 uniformly-spaced sampling phases, and consider the best performance of BERs associated with different sampling phases as the BER of the optimal sampling phase. In Fig. 7, the proposed sampling phase estimation criterion has superior BER performance to its counterpart. The performance gain can be 0.2dB over both Brazil-D and Brazil-E channels. In addition, the proposed criterion performs closely to the optimal sampling phase, which indicates the excellent performance of the proposed criterion. Therefore, the proposed method can be considered to be near optimal.

V.   CONCLUSION

In this paper, we derived the theoretical BER for encoded TDS-OFDM system over AWGN channel, and it can be observed that the sampling phase offset has a great impact on BER. Furthermore, we proposed a near optimal sampling phase estimation criterion. Compared with the conventional time domain based synchronization methods for TDS-OFDM, the proposed criterion obtains the near optimal sampling phase based on the frequency-domain response. Simulations

demonstrate that the proposed criterion has good performance in actual LDPC-coded TDS-OFDM system over multipath channels. The proposed criterion is superior to the conventional STR method and performs closely to the optimal sampling phase over multipath channels.

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