A PILOT-LESS SAMPLE-TIME SYNCHRONIZATION ALGORITHM FOR HIGH-MOBILITY DVB-T RECEIVING

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ABSTRACT

The OFDM system receiver is well-performed only when the transmitter and the receiver are precisely synchronized. The sampling clock offset causes a symbol-time drift and even introduces inter-carrier and inter-symbol interference (ICI and ISI), which become not negligible in the systems with large OFDM symbol sizes, e.g. in the terrestrial digital video broadcasting system (DVB-T). The sample-time synchronization (STS) algorithm becomes difficult in the high-mobility wireless transmission because of the Doppler Effect. In this paper, we propose an innovative STS algorithm especially suitable for DVB-T receiving in the high-mobility environment. Simulation results show that the proposed algorithm can accurately recover the sample-time when the DVB-T signals are received in a fast moving vehicle.

1. INTRODUCTION

The Orthogonal Frequency-Division Multiplexing (OFDM) is popular for broadband wireless transmission nowadays because of the ability of combating the multi-path channel problem. However, the advantages of the OFDM system are assured only when the transmitter and the receiver are precisely synchronized. The OFDM system is very sensitive to the symbol-timing error and the sampling clock offset (SCO). A tiny SCO will eventually cause a drift in symbol-timing and even infer non-negligible ICI and ISI effects especially in the broadcast-based systems with large symbol sizes [3]. Terrestrial Digital Video Broadcasting (DVB-T [1]) system adopts OFDM with large symbol sizes (2048/8192 samples) in order to provide adequate video transmission bandwidth. The sample-time recovery mechanism is therefore necessary in the DVB-T demodulator.

When the mobile wireless transmission is considered, synchronization maintenance will become more difficult than the fixed wireless transmission because of the Doppler Effect. Without properly solving this problem, the system performance will decay significantly in the high-mobility environment. Many sample-time synchronization (STS) algorithms have been proposed [3, 4, 5, 11] and successfully applied in DVB-T demodulators. However, most of them operate properly for the stationary and the low-mobility DVB-T receiving. To combat the severe Doppler Effect in the high-mobility environment, a novel and effective synchronization algorithm is necessary. In this paper, we aim at the STS in the high-mobility environment and attempt to propose an innovative algorithm, which is almost not affected by the severe fast fading caused by Doppler Effect. In the algorithm, we estimate SCO in the time domain without the assistance of pilot tones. Simulation results show that the proposed algorithm can lock the sample-time within 0.6 ppm when the mobile velocity is up to 300 km/hr and SNR is 30 dB.

2. OVERVIEW OF OFDM TRANSMISSION IN DIGITAL VIDEO BROADCASTING SYSTEMS

DVB-T allows many choices of transmission parameters in order to support the digital broadcast in different countries and regions. There are two kinds of bandwidths: 7 MHz and 8 MHz. The DVB-T transmitter can use OFDM with the two choices of the number of subcarriers (N): 2K mode ($N = 2048$) and 8K mode ($N = 8192$ mode). Each subcarrier can be modulated with either QPSK, 16 QAM, or 64 QAM. For the 2K and 8K modes operated in the 8 MHz bandwidth, the useful durations (excluding the cyclic prefix) of an OFDM symbol are T_u = 224 us and T_u = 896 us, respectively. The length of cyclic prefix (CP) can be 1/4, 1/8, 1/16, and 1/32 of T_u.

In DVB-T, there are two types of pilots: scattered pilots and continual pilots, as illustrated in Fig. 1. The value of each pilot is either +4/3 or -4/3 and determined only by the subcarrier location. In Figure 1, in the time direction, the locations of scattered pilots are periodic for every four OFDM symbols. In the frequency direction, there is one scattered pilot every twelve subcarriers.

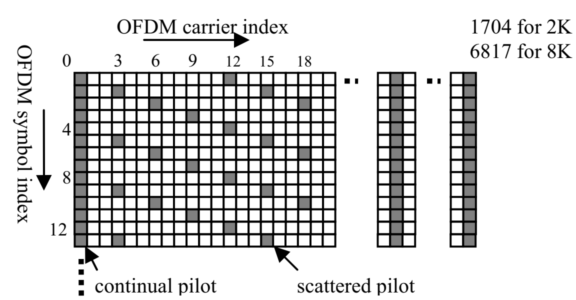

Fig. 1 The location diagram of scattered pilots in DVB-T.

3. SAMPLE-TIME SYCHRONIZATION ALGORITHMS FOR OFDM SYSTEMS

An approach of sample-time synchronization is to estimate SCO and then dynamically compensate the received subcarriers by phase rotation on the symbol-by-symbol basis [2]. This approach can successfully compensate the phase rotation caused by the sample-time phase shift. However, apart from the sample-time phase drift, the ICI caused by SCO (whose power is proportioned to the square of the subcarrier index [3]) should not be neglected in the system like DVB-T. The above phase rotation approach will not operate properly because the number of subcarrier can be large up to 8192 for DVB-T. The tracking loop approach [2] is the only alternative to provide good performance. The traditional way of SCO detection is based on the post-FFT (namely, the frequency-domain) algorithm [4, 5]. The received time-domain signal is converted to the frequency domain data in which the reference subcarriers (known as pilot tones) can be used to estimate SCO. We define the normalized carrier frequency offset as $\Delta f_c' = \Delta f_c \cdot T_u = (f_{c,rx} - f_{c,tx})T_u$, where $f_{c,rx}$ and $f_{c,tx}$ are the carrier center frequencies of the receiver and the transmitter, respectively. The normalized SCO is defined as $t_\Delta = (T'-T)/T$, where T' and T are the sampling periods of the receiver and the transmitter, respectively. Considering that only the kth subcarrier of the lth OFDM symbol (denoted as $X_{l,k}$) is transmitted, we can obtain the kth subcarrier of the l-th post-FFT in the receiver as [4],

$$R_{l,k} = (e^{j\pi\phi_k} \cdot e^{j2\pi\frac{lN_s+N_g}{N}\phi_k}) \cdot \sin c(\phi_k) X_{l,k} C_{l,k} + V_{l,k}, \tag{1}$$

where $\phi_k = \Delta f_c' + t_\Delta \cdot k$.

N_s and N_g are the number of samples of an OFDM symbol and the CP, respectively, $C_{l,k}$ is the channel frequency response at the kth subcarrier of the lth OFDM symbol. The noise term $V_{l,k}$ contains not only the channel noise but also the ICI noise.

We can use the reference subcarrier (pilot tone) data to extract the information of t_Δ. Usually, the scattered pilots instead of the continual pilots are used for SCO estimation in the DVB-T receiver [5] because the number of scattered pilots is much greater than that of the continual pilots. In addition, the locations of the scattered pilots are evenly spread among subcarriers and OFDM symbols so that the there are possibly more usable scatter pilots than usable continual pilots under the poor channel condition. In DVB-T, if k_p is the subcarrier index of a scatter pilot in the lth OFDM symbol, then the k_pth subcarrier of the (l-4) will be the scatter pilot, too. Let R_{l,k_p} with R_{l-4,k_p} denote the received post-FFT data at the k_pth subcarrier index in the lth and the (l-4)th OFDM symbols, respectively. By correlating R_{l,k_p} with R_{l-4,k_p}, we obtain

$$\begin{aligned} Y_{l,k_p} &= R_{l,k_p} \cdot R^*_{l-4,k_p} \\ &= e^{j2\pi\frac{4N_s}{N}\phi_{k_p}} \cdot (X_{l,k_p} \cdot X^*_{l-4,k_p}) \cdot (C_{l,k_p} \cdot C^*_{l-4,k_p}) \end{aligned} \tag{2}$$

In DVB-T, all the scattered pilots are real-valued. Therefore, the phase of Y_{l,k_p} is

$$\angle Y_{l,k_p} = 2\pi \frac{4N_s}{N}\phi_{k_p} + \angle(C_{l,k_p} \cdot C^*_{l-4,k_p}) \tag{3}$$

Under the slow-fading environment, the variation of the channel during four OFDM symbols is small and usually neglected. We can assume that the phase of the correlation of $C_{l-4,k}$ and $C_{l,k}$ is approximately equal to zero. Consequently, ϕ_{k_p} can be detected from the phase of $Y_{l,kp}$ as follows,

$$\phi_{k_p} = \frac{\angle Y_{l,k_p}}{2\pi \cdot 4N_s / N} \tag{4}$$

From eq. (4), ϕ_k is linear function of k with a slope of t_Δ. Hence, we can detect t_Δ by using Least-Squares (LS) criterion as follows [4], where N_p is the number of scattered pilots,

$$\hat{t}_\Delta = \frac{N_p \cdot \sum_{p=1}^{N_p} \phi_{k_p} k_p - (\sum_{p=1}^{N_p} k_p) \cdot (\sum_{p=1}^{N_p} \phi_{k_p})}{N_p \cdot \sum_{p=1}^{N_p} k_p^{\ 2} - (\sum_{p=1}^{N_p} k_p)} \tag{5}$$

This widely-used post-FFT algorithm of SCO estimation performs well in the motionless and the low-mobility environment. However, in the high-mobility environment, the fast channel variation makes the phase difference between $C_{l-4,k}$ and $C_{l,k}$ not negligible. It follows that the use of (4) for detecting ϕ_k will introduce errors. Therefore, it results in performance degradation of the post-FFT based sample-time synchronization algorithm when the DVB-T signal is received in a fast moving object.

4. PILOT-LESS SAMPLE-TIME SYCHRONIZATION ALGORITHM FOR OFDM SYSTEMS

To solve the erroneous SCO detection problem in the high-mobility environment, we propose a novel detection algorithm, which is based on the pre-FFT (time-domain) data and does not require pilots.

Provided that the lth OFDM symbol boundary has been synchronized correctly, the N_s samples of the lth time-domain symbol are $\{r_{l,0}, r_{l,1}, \ldots, r_{l,N_s-1}\}$. Their sample-times are $\{t_{l,0}, t_{l,1}, \ldots, t_{l,N_s-1}\}$, which can also be written as $\{t_{l,0}, t_{l,0}+T', \ldots, t_{l,0}+(N_s-1)T'\}$. Without loss of generality, we assume that the sampling period at the transmitter is equal to unity so that T' can be written as $T'=1+t_\Delta$, and the above sample-times can be further written as $\{t_{l,0}, t_{l,0}+(1+t_\Delta), \ldots, t_{l,0}+(N_s-1)(1+t_\Delta)\}$. Assume that the sample-time $t_{l,0}$ has been correctly synchronized with the corresponding sample-time of the transmitter. The correct sample-time of the nth sample of the lth OFDM symbol is $(t_{l,0}+n)$. Because of the mismatch of T and T', there is a sample-time error $n \cdot t_\Delta$ for the nth sample. For example, the sample-time of the nth sample of the lth OFDM symbol is $t_{l,n} = t_{l,0}+n(1+t_\Delta)$, but the correct sample-time should be $(t_{l,0}+n)$. The relative locations of these sample-times are illustrated in Fig. 2.

By using Taylor Series, we can express $r(t_{l,0}+i(1+t_\Delta))$ in terms of $r(t_{l,0}+i)$ as

$$r(t_{l,0}+n(1+t_\Delta)) = r(t_{l,0}+n) + r'(t_{l,0}+n)\cdot(nt_\Delta) + \frac{1}{2!}r''(t_{l,0}+n)\cdot(nt_\Delta)^2 + \ldots. \tag{6}$$

Because the SCO is usually small, we can approximate $r(t_{l,0}+n(1+t_\Delta))$ by neglecting the terms with $(nt_\Delta)^i$, for $i \geq 2$. Accordingly, the SCO can be estimated approximately as:

$$t_\Delta \approx \frac{r(t_{l,0}+n(1+t_\Delta)) - r(t_{l,0}+n)}{n \cdot r'(t_{l,0}+n)} \tag{7}$$

To use (7) to evaluate t_Δ, we let n be N (FFT size) and make use of the cyclic property in the guard interval: $r(t_{l,0}+N) = r(t_{l,0})$ and $r'(t_{l,0}+N) = r'(t_{l,0})$. We use all samples in the guard interval to estimate t_Δ as follows:

$$\hat{t}_\Delta = \sum_{i=0}^{N_g-1} \frac{r(t_{l,i}+N(1+t_\Delta)) - r(t_{l,i})}{N \cdot r'(t_{l,i})} \tag{8}$$

To calculate the derivation term, $r'(t_{l,i})$, we can apply a digital differentiator to the received samples. In our DVB-T demodulator, we adopt a B-spline differentiator [6].

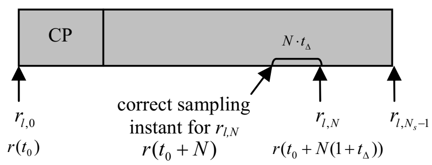

Fig. 2 The timing diagram for the samples in one OFDM symbol.

In (6), we have omitted the terms with $(nt_\Delta)^i$ for $i \geq 2$ in order to simplify the implementation. However, t_Δ may be large at the beginning of synchronization and may cause detection errors. To prevent this situation, we can first conduct coarse detection of SOC by using the post-FFT algorithm in the acquisition state and then apply the pre-FFT algorithm in the tracking state.

In OFDM system, the guard intervals are purposed to solve the ISI problem, but themselves are interfered by neighbor symbols. This interference may introduce a potential problem in the proposed pre-FFT algorithm. When the channel delay spread is not too long, we can use the ISI-free part of the guard interval to avoid this problem. In case the channel spread is as long as the guard interval, all points in the guard interval will be contaminated by the ISI noise. Nevertheless, this problem can be still relieved by properly adjusting the bandwidth of the loop filter.

5. PERFERMANCE EVAULATION FROM COMPUTER SIMULATION

In sample-time synchronization, we can adjust the sample-times of the time-domain signal by using a digital resampler. The structure of sample-time synchronization is shown in Fig. 3 (a). In our implementation, a digital resampler using the cubic B-spline interpolation is adopted in order to provide satisfactory performance without hardware complexity increment [6, 7]. From our analysis, when the oversampling rate (OSR) is 2, the cubic B-spline interpolation can provide around 45 dB SER (Signal to interpolation Error Ratio), and the fifth-order B-spline interpolation can even reach around 62 dB SER [7]. In addition, we use a cubic B-spline differentiator [6], as shown in Figure 3 (b). In our computer simulation, the loop filter in the sample-time recovery loop is a second-order loop filter with the transfer function (K_0+K_1/(z-1)) with K_0=0.004, and K_1=0.0016.

In the simulation, the system specifications are as follows: 2K mode, 8 MHz channel bandwidth, QPSK modulation, 1/8 CP length, and 500 MHz carrier frequency. The wireless channel is modeled as a Rayleigh multipath fading channel using the modified Jakes' fading model [8]. The power-delay profile is modeled by HT channel model in COST207 [9]. The carrier frequency offset Δf_c is 100 ppm. The sample clock offset (SCO) t_Δ is set to 100 ppm. To evaluate the performance of synchronization, we define a parameter, the residual clock offset, which is the averaged value (16 trials) of the difference between the tracking SCO in the steady state and the real SCO (t_Δ). In our DVB-T demodulator, the OFDM symbol boundary is estimated by making use of the cyclic property of the guard interval [10]. The carrier frequency offset estimation is completed by the aids of CP and pilots in DVB-T as used in [4, 5]. Finally, we estimate the channel responses by using the scattered pilots. The scattered estimated channel responses are linearly interpolated in time direction and sinc interpolated in frequency direction to obtain the full-range channel responses [11].

Fig. 4 shows the clock tracking performance of the traditional post-FFT and the proposed pre-FFT algorithms. When a slow-fading channel is applied, the residual clock offset can be reduced within 0.5 ppm and 1ppm for the post-FFT algorithm and the pre-FFT algorithm, respectively. In the mobile environment, the performance of the post-FFT algorithm gradually decays when the mobile velocity increases. With the speeds of 100, 200, and 300 km/hr, the post-FFT method will result in the residual clock offsets of around 7 ppm, 9 ppm, and 15 ppm, respectively. However, the proposed pre-FFT approach always has the same performance (within 1 ppm) no matter what speed is applied. The bit error rates (BER) of the inner receiver is plotted as functions of Es/No (SNR) in Fig. 5 (QPSK modulation) and in Fig. 6 (16-QAM modulation). The "ideal scheme" means the receiving with perfect synchronizations (both frequency and timing) and perfect channel estimation. In the traditional algorithm (the post-FFT), the BER performance decays when the mobile velocity increases, as shown in Fig. 5 (a) and 6 (a). On the other hand, the BER of the proposed pre-FFT algorithm is not affected even in the high-mobility environment, as shown in Fig. 5 (b) and 6 (b).

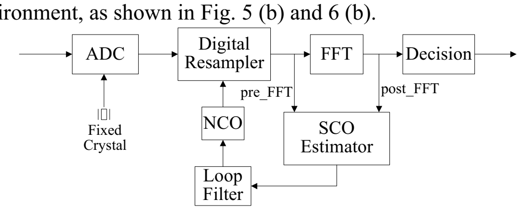

Fig. 3 (a) The structure of sample-time synchronization

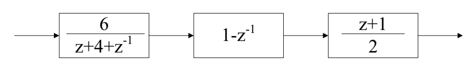

Fig. 3 (b) The structure of a cubic B-spline differentiator

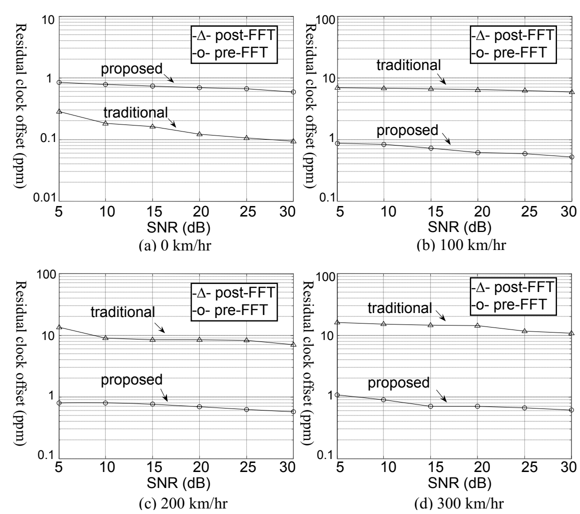

Fig. 4 Clock tracking performance comparison under the mobile velocity of 0, 100, 200, and 300 km/hr.

6. CONCLUSIONS

In this paper, we propose a sample-time synchronization algorithm that estimates the SCO based on the time-domain (pre-FFT) samples without using pilots. This algorithm is successfully applied to DVB-T demodulation in the situation of high-mobility receiving. By using computer simulation, the performance of the traditional post-FFT and the proposed pre-FFT algorithms is evaluated for comparison. The simulation results show that the proposed pre-FFT method has better performance than the post-FFT method when the receiver is on a fast-moving object. Due to the robustness of the pre-FFT method, it can be considered as an alternative for the high-mobility sample-time synchronization algorithm.

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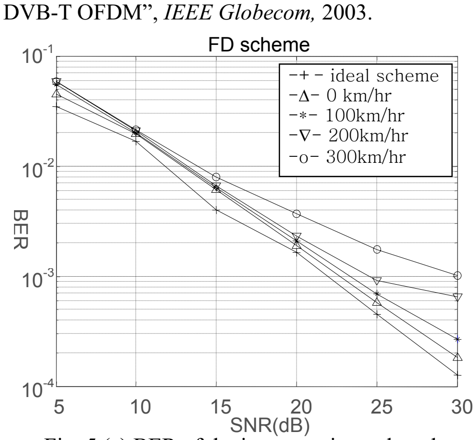

Fig. 5 (a) BER of the inner receiver when the traditional post-FFT algorithm is applied. (QPSK)

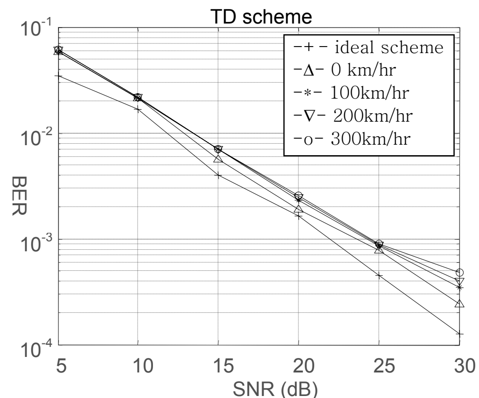

Fig. 5 (b) BER of the inner receiver when the proposed pre-FFT algorithm is applied. (QPSK)

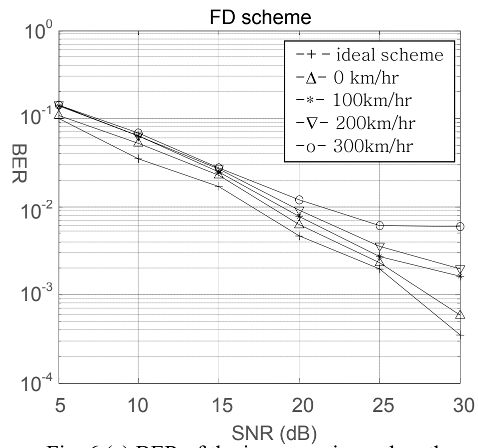

Fig. 6 (a) BER of the inner receiver when the traditional post-FFT algorithm is applied. (16QAM)

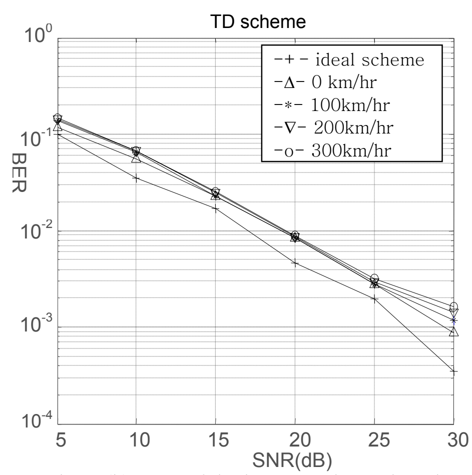

Fig. 6 (b) BER of the inner receiver when the proposed pre-FFT algorithm is applied. (16QAM)