Blind Symbol Synchronization Based on Cyclic Prefix for OFDM Systems

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***Abstract*—In this paper, a blind symbol synchronization algorithm is presented for orthogonal frequency-division multiplexing (OFDM) systems, and a new timing function based on the redundancy of the cyclic prefix (CP) is introduced. It proves that the maximum of this function necessarily points to the correct timing offset, irrespective of channel conditions when the signal-to-noise ratio is high. Using the timing function, the timing offset is estimated through a searching algorithm. Channel power profile and channel length information are unnecessary. Simulation results show that the proposed algorithm is robust and outperforms the existing CP-based algorithms, particularly in frequency-selective fading channels.**

***Index Terms*—Orthogonal frequency-division multiplexing (OFDM), symbol synchronization.**

I. Introduction

IN ORTHOGONAL frequency-division multiplexing (OFDM) systems, synchronization issues are of great importance since synchronization errors might destroy the orthogonality among all subcarriers and, therefore, introduce intercarrier interference (ICI) and intersymbol interference (ISI) [1]–[3]. Generally, synchronization tasks include sampling clock synchronization, symbol synchronization, and carrier frequency synchronization [4]. Sampling clock synchronization mitigates the sampling clock errors due to the mismatch of the crystal oscillators, while carrier frequency synchronization eliminates the carrier frequency offset arising from transceiver oscillator mismatches and/or Doppler shifts. In this paper, symbol synchronization is the focus. The objective is to find the correct starting position [i.e., the fast Fourier transform (FFT) window] of the OFDM symbol for FFT demodulation. It is equivalent to estimate the timing offset between the transmitter and receiver. This is also referred to as timing synchronization. Note that the timing offset is an integer and can be anywhere within an OFDM symbol.

Symbol synchronization may be performed at the receiver using a predesigned preamble [4]–[7]. Although accurate estimation can be achieved, the bandwidth efficiency is inevitably reduced. To eliminate this reduction, algorithms using the redundancy introduced by the cyclic prefix (CP) have been proposed [4], [8]–[12]. The most representative one is the maximum likelihood (ML) symbol synchronization algorithm [8]. However, good performance is achieved only under flat-fading channels. When the system is operating under frequency-selective fading channels, the ML algorithm exhibits significant fluctuation in the estimated timing offset. This is because the maximum of the timing function depends on channel conditions and does not necessarily point to the correct timing offset due to ISI under frequency-selective fading channels. Recently, a number of algorithms have been proposed in [9]–[12] to combat the effect of ISI and mitigate the fluctuation. Unfortunately, the double correlation method in [9] works well only when the first path channel response is the strongest. The correlation derivative algorithm in [10] makes use of the property that the correlator output is the sum of a set of triangular functions and is sensitive to the carrier frequency offset, as well as the filter length used [13]. The rank method based on minimum description length (MDL) criterion in [11] is computation intensive. The asymptotic ML algorithm in [12] uses the correlation length equal to the sum of the channel and CP lengths. It is rather impractical since exact knowledge of the channel power profile and channel length is required. The simplified minimum mean square error and maximum correlation algorithms in [12] set the correlation length to twice the CP length and still include ISI in the timing function, resulting in fluctuation and poor performance.

In this paper, a blind symbol synchronization algorithm is proposed for OFDM systems, and a new timing function is designed based on the redundancy introduced by the CP. Using the proposed timing function, the timing offset is estimated through a 2-D search. It is proved that under a high signal-to-noise ratio (SNR) assumption, the maximum of the timing function necessarily points to the correct timing offset, irrespective of channel conditions, even under frequency-selective fading channels. Since the position uncertainty in the maximum point is removed, the fluctuation of the estimated timing offset is mitigated, and the estimation accuracy is thus improved. In addition, there is no restriction on the strength of the first path channel response, and channel information (including channel power profile and channel length) is not required. Computer simulations show that the proposed algorithm achieves an unbiased estimation of the timing offset and that the mean square of the estimation error is significantly reduced compared with the existing CP-based algorithms when the system is operating under frequency-selective fading channels. It is also verified that the proposed algorithm is robust to frequency-selective fading channels.

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The rest of this paper is organized as follows. In Section II, the OFDM system model is introduced. The CP-based blind symbol synchronization algorithm is presented in Section III, and in Section IV, the performance of the proposed algorithm is demonstrated by computer simulations. Finally, Section V draws the conclusion.

II. System Model

Consider a baseband discrete-time OFDM system with N subcarriers. The independent identically distributed (i.i.d.) complex signals $d(n)$, $n \in \{0, 1, \ldots, N-1\}$, are modulated onto the N subcarriers by means of the inverse discrete Fourier transform (IDFT). The CP with length N_{cp} is then inserted at the beginning of each OFDM symbol to avoid ISI and preserve the orthogonality among subcarriers. The OFDM symbol $s(n)$, $n \in \{0, 1, \ldots, N+N_{\mathrm{cp}}-1\}$, is finally serially transmitted through a frequency-selective fading channel, which is generally modeled as an $(L+1)$-tap finite-impulse response filter. Here, the channel length L is assumed to be shorter than the CP length N_{cp}.

At the receiver, assuming perfect sampling clock synchronization has been achieved, the received sampled signal can be expressed as

$$r(n) = e^{j2\pi\varepsilon n/N} \sum_{l=0}^{L} h(l)s(n-l-\theta) + w(n) \qquad n = 0, 1, \ldots, N+N_{\mathrm{cp}}-1 \quad (1)$$

where ε denotes the carrier frequency offset (which is normalized by the subcarrier spacing) due to the mismatch of the transceiver oscillators and/or the Doppler frequency shift, θ represents the timing offset to be estimated, $\theta \in \{0, 1, \ldots, N+N_{\mathrm{cp}}-1\}$, $h(l)$ is the response of the frequency-selective fading channel, $s(n)$ is the OFDM signal after CP insertion, and $w(n)$ denotes the additive white Gaussian noise (AWGN) with zero mean and variance σ_n^2, which is independent of the OFDM signals $s(n)$.

III. CP-Based Blind Symbol Synchronization

Now, consider the correlation between the received sampled signals $r(n)$ and $r(n+N)$. From (1), it can be written as

$$E\{r(n)r^*(n+N)\} = E\{a(n) + b(n)\} \quad (2)$$

where

$$a(n) = e^{-j2\pi\varepsilon} \left(\sum_{l=0}^{L} h(l)s(n-l-\theta)\right) \times \left(\sum_{l=0}^{L} h^*(l)s^*(n+N-l-\theta)\right) \quad (3)$$

$$b(n) = e^{j2\pi\varepsilon n/N}\left(\sum_{l=0}^{L} h(l)s(n-l-\theta)\right) w^*(n+N) + e^{-j2\pi\varepsilon(n+N)/N}\left(\sum_{l=0}^{L} h^*(l)s^*(n+N-l-\theta)\right) w(n) + w(n)w^*(n+N). \quad (4)$$

Since the zero mean AWGN noise $w(n)$ is independent of the OFDM signals, $E\{b(n)\} = 0$. On the other hand, the OFDM signal $s(n)$ is the IDFT of the i.i.d. signals $d(n)$, and the signals in the CP $\{s(0), s(1), \ldots, s(N_{\mathrm{cp}}-1)\}$ are a copy of the last N_{cp} samples in the body of the OFDM symbol. The correlation of the OFDM signal $s(n)$ therefore satisfies

$$E\{s(n_1)s^*(n_2)\} = \begin{cases} \sigma_s^2, & n_1 = n_2 \text{ or } n_1 = n_2 \pm N \\ 0, & \text{otherwise} \end{cases} \qquad n_1, n_2 \in \{0, 1, \ldots, N+N_{\mathrm{cp}}-1\} \quad (5)$$

where $\sigma_s^2 \triangleq E\{|s(n)|^2\}$. As a result, the correlation between $r(n)$ and $r(n+N)$ is equivalent to $E\{r(n)r^*(n+N)\} = E\{a(n)\} = e^{-j2\pi\varepsilon} \sum_{l=0}^{L} |h(l)|^2 E\{s(n-l-\theta)s^*(n+N-l-\theta)\}$ and has the following property:

$$E\{r(n)r^*(n+N)\} = E\{a(n)\} = \begin{cases} \sigma_x^2 e^{-j2\pi\varepsilon}, & n \in \{\theta+L, \theta+L+1, \ldots, \theta+N_{\mathrm{cp}}-1\} \\ P_1(n)e^{-j2\pi\varepsilon}, & n \in \Omega_1 \\ P_2(n)e^{-j2\pi\varepsilon}, & n \in \Omega_2 \\ 0, & \text{otherwise} \end{cases} \quad (6)$$

where $\sigma_x^2 \triangleq E\{|r(n)|^2\} - \sigma_n^2 = \sum_{l=0}^{L} |h(l)|^2 \sigma_s^2$, and

$$P_1(n) = \sum_{l=0}^{n-\theta} |h(l)|^2 \sigma_s^2 \qquad \Omega_1 \triangleq \{\theta, \theta+1, \ldots, \theta+L-1\} \quad (7)$$

$$P_2(n) = \sum_{l=n-\theta-N_{\mathrm{cp}}+1}^{L} |h(l)|^2 \sigma_s^2 \qquad \Omega_2 \triangleq \{\theta+N_{\mathrm{cp}}, \theta+N_{\mathrm{cp}}+1, \ldots, \theta+N_{\mathrm{cp}}+L-1\}. \quad (8)$$

Obviously, $P_1(n)$ and $P_2(n)$ are positive, with $P_1(n) < \sigma_x^2$ and $P_2(n) < \sigma_x^2$.

A new timing function is now proposed as follows:

$$\Lambda(k,m) \triangleq \left| \sum_{n=m}^{N_{\rm cp}-1} E\left\{ r(n+k) r^*(n+N+k) \right\} \right| - \frac{\rho^2}{2} \sum_{n=m}^{N_{\rm cp}-1} \left[E\left\{ |r(n+k)|^2 \right\} + E\left\{ |r(n+N+k)|^2 \right\} \right]$$
$$k \in \{0,1,\ldots,N+N_{\rm cp}-1\}; \qquad m \in \{0,1,\ldots,N_{\rm cp}-1\} \tag{9}$$

where $\rho \triangleq (E\{|r(n)|^2\} - E\{|w(n)|^2\})/E\{|r(n)|^2\} = \sigma_x^2/(\sigma_x^2+\sigma_n^2)$. From the definition of $\Lambda(k,m)$, we have the following property.

Property 1: The maximum of the function $\Lambda(k,m)$ occurs at the point ($k=\theta$, $m=L$) when $\sigma_x^2 \gg \sigma_n^2$ (high SNR) and $\sigma_n^2 \neq 0$.

Proof: Since $E\{|r(n)|^2\} = \sigma_x^2 + \sigma_n^2$, the function $\Lambda(k,m)$ is equivalent to

$$\Lambda(k,m) = \left| \sum_{n=m}^{N_{\rm cp}-1} E\left\{ r(n+k) r^*(n+N+k) \right\} \right| - (N_{\rm cp}-m)\sigma_x^4 / \left(\sigma_x^2+\sigma_n^2\right). \tag{10}$$

When $\sigma_x^2 \gg \sigma_n^2$, the second term in (10) can be approximated as $(N_{\rm cp}-m)\sigma_x^2$.

When $k \neq \theta$, it follows from (6) and (10) that the function $\Lambda(k,m)$ satisfies

$$\Lambda(k\neq\theta, m<L) < (N_{\rm cp}-m)\sigma_x^2 - (N_{\rm cp}-m)\sigma_x^2 = 0 \tag{11}$$
$$\Lambda(k\neq\theta, m=L) < (N_{\rm cp}-L)\sigma_x^2 - (N_{\rm cp}-L)\sigma_x^2 = 0 \tag{12}$$
$$\Lambda(k\neq\theta, m>L) \leq (N_{\rm cp}-m)\sigma_x^2 - (N_{\rm cp}-m)\sigma_x^4/(\sigma_x^2+\sigma_n^2) = (N_{\rm cp}-m)\Delta(\sigma_x^2,\sigma_n^2) \tag{13}$$

where $\Delta(\sigma_x^2,\sigma_n^2) = \sigma_x^2\sigma_n^2/(\sigma_x^2+\sigma_n^2)$.

Similarly, when $k=\theta$, it yields

$$\Lambda(k=\theta, m<L) < (N_{\rm cp}-m)\sigma_x^2 - (N_{\rm cp}-m)\sigma_x^2 = 0 \tag{14}$$
$$\Lambda(k=\theta, m=L) = (N_{\rm cp}-L)\sigma_x^2 - (N_{\rm cp}-L)\sigma_x^4/\left(\sigma_x^2+\sigma_n^2\right) = (N_{\rm cp}-L)\Delta\left(\sigma_x^2,\sigma_n^2\right) \tag{15}$$
$$\Lambda(k=\theta, m>L) = (N_{\rm cp}-m)\sigma_x^2 - (N_{\rm cp}-m)\sigma_x^4/\left(\sigma_x^2+\sigma_n^2\right) = (N_{\rm cp}-m)\Delta\left(\sigma_x^2,\sigma_n^2\right). \tag{16}$$

Since $\sigma_n^2 \neq 0$, which means that $\Delta(\sigma_x^2,\sigma_n^2) > 0$, $(N_{\rm cp}-L)\Delta(\sigma_x^2,\sigma_n^2)$ in (15) is greater than zero and $(N_{\rm cp}-L)\Delta(\sigma_x^2,\sigma_n^2)$ in (15) is also greater than $(N_{\rm cp}-m)\Delta(\sigma_x^2,\sigma_n^2)$ in (13) and (16), where $m>L$. Consequently, it can be concluded from (11)–(16) that the function $\Lambda(k,m)$ achieves its maximum at the point ($k=\theta$, $m=L$), and the proof follows.

Based on Property 1, the timing offset θ can be estimated from the maximum point of the timing function $\Lambda(k,m)$ as

$$\hat{\theta} = \arg\max_k \max_m \Lambda(k,m)$$
$$m \in \{0,1,\ldots,N_{\rm cp}-1\}; \qquad k \in \{0,1,\ldots,N+N_{\rm cp}-1\}. \tag{17}$$

The implementation of this symbol synchronization algorithm is then summarized as follows: 1) Compute the timing function $\Lambda(k,m)$ for $\{k \in \{0,1,\ldots,N+N_{\rm cp}-1\}, m \in \{0,1,\ldots,N_{\rm cp}-1\}\}$, and 2) choose k corresponding to the maximum of $\Lambda(k,m)$ as the estimated timing offset. Notice that two parameters (i.e., k and m) are included in the proposed timing function $\Lambda(k,m)$. This is different from the existing CP-based algorithms [8]–[12], where the timing functions have only one parameter. The parameter k corresponding to the maximum of $\Lambda(k,m)$ gives an estimation of the timing offset. Meanwhile, the parameter m corresponding to the maximum of $\Lambda(k,m)$ also yields an estimation of the channel length L. Since channel length estimation is outside the scope of this paper, it will not be further discussed here. On the other hand, it is noticed from the definition of the timing function (9) that channel information (including channel power profile and channel length) is unnecessary here, and there is no restriction on the strength of the first path channel response $h(0)$. ■

Remark 1: In the existing CP-based algorithms [8], [10], [12], due to the existence of ISI, the maximum of the timing function fluctuates in the range of $[\theta, \theta+L]$, where L is the frequency-selective fading channel length, and the maximum depends on channel conditions. This position uncertainty in the maximum will cause significant fluctuation of the estimated timing offset. Unlike the existing algorithms, the maximum of the proposed timing function $\Lambda(k,m)$ necessarily occurs at the point ($k=\theta$, $m=L$). That is, the maximum of $\Lambda(k,m)$ necessarily points to the correct timing offset θ, irrespective of channel conditions. Since the position uncertainty is removed, the fluctuation of the estimated timing offset is mitigated, which means that an accurate estimation is obtained.

Remark 2: From Property 1, it is apparent that the maximum value of $\Lambda(k,m)$ is $(N_{\rm cp}-L)\Delta(\sigma_x^2,\sigma_n^2)$ (15), while the value of $\Lambda(k,m)$ at $m=L+1$ may be $(N_{\rm cp}-L-1)\Delta(\sigma_x^2,\sigma_n^2)$ (13), (16). They differ only by a factor of $\Delta(\sigma_x^2,\sigma_n^2)$. In the finite-sample scenario, the position of the maximum might be perturbed, which would degrade the estimation performance. To avoid this perturbation, a high-resolution analog-to-digital converter is thus required in practice.

Apparently, from the timing function in (9), $N_{\rm cp}$ correlation windows are included by means of the parameter m, and the timing offset is estimated through a 2-D search. The number of complex multiplications required here is therefore $3\sum_{i=1}^{N_{\rm cp}} i = 3N_{\rm cp}(N_{\rm cp}+1)/2$, while that required in the ML algorithm in [8] and the correlation derivative algorithm in [10] is $3N_{\rm cp}$ and $N_{\rm cp}$, respectively. That is, the complexity of the proposed algorithm in terms of the number of complex multiplications is respectively $(N_{\rm cp}+1)/2$ and $3(N_{\rm cp}+1)/2$ times that of [8]

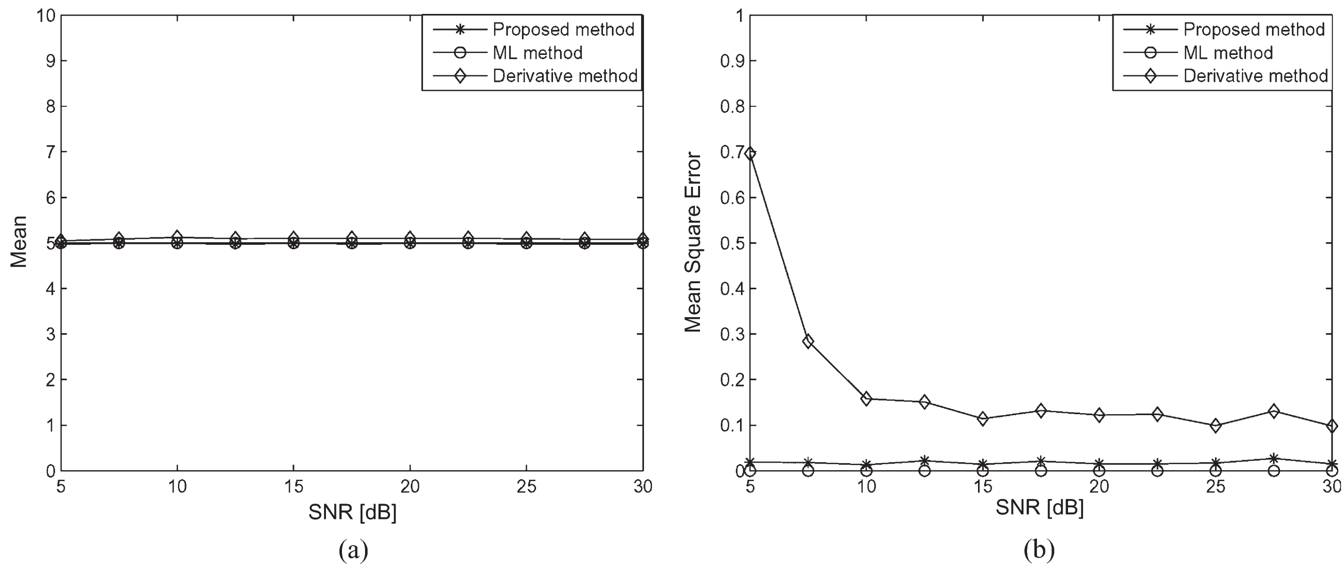


Fig. 1. Performance versus SNR for flat-fading channels ($L = 0$).

and [10]. On the other hand, the proposed algorithm can work well, irrespective of channel conditions, while the algorithms in [8], [10] do not yield accurate estimation in frequency-selective fading channels. This will be demonstrated by computer simulations in the following section.

IV. Simulation Results

Computer simulations have been conducted to investigate the performance of the proposed algorithm. In the simulations, an OFDM system with 128 subcarriers ($N = 128$) and 12.5% CP ($N_{\mathrm{cp}} = 16$) was considered. The carrier frequency offset was assumed to be one-third of the subcarrier spacing ($\varepsilon = 1/3$). The timing offset was set to five sample periods ($\theta = 5$). The performance was evaluated in terms of the mean of the estimated timing offset and the mean square of the estimation error (MSE). All these performance measures were computed by averaging the results over 1000 Monte Carlo realizations. In each run, 1) a data packet with 20 random OFDM symbols was transmitted; 2) all transmitted signals were modulated by the quaternary phase-shift keying scheme; and 3) slow fading was considered here, and the channel responses were randomly generated and time invariant during each run. SNR was defined as

$$\mathrm{SNR} \triangleq E\left\{|r(n) - w(n)|^2\right\} / E\left\{|w(n)|^2\right\} = \sigma_x^2/\sigma_n^2. \quad (18)$$

The expectation $E\{\cdot\}$ in $\Lambda(k, m)$ was calculated by averaging the term in brackets over 20 OFDM symbols. With the noise variance σ_n^2 estimated by the methods in [11], [14], and [15], the parameter ρ in the proposed algorithm could be computed as $\rho = (E\{|r(n)|^2\} - \sigma_n^2)/E\{|r(n)|^2\}$. In the simulation, it was assumed that ρ was perfectly known. For comparison purposes, the ML algorithm in [8] and the correlation derivative algorithm in [10] were also implemented here. Note that in the correlation derivative algorithm [10], the filter length used was set to 16, and the hard replacement rule was adopted.

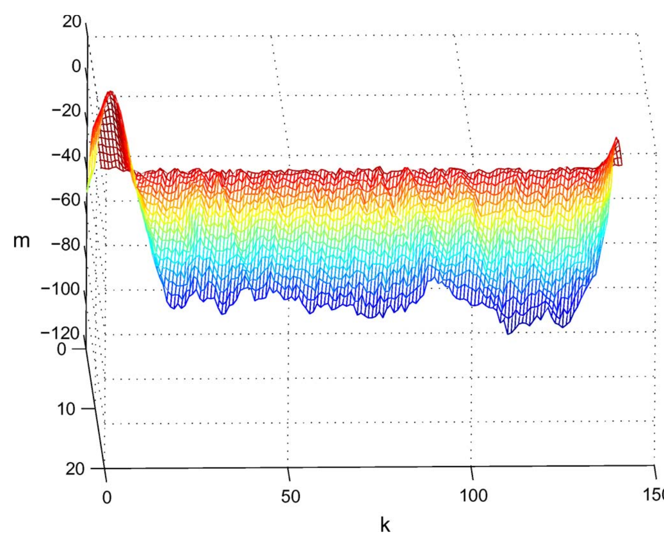
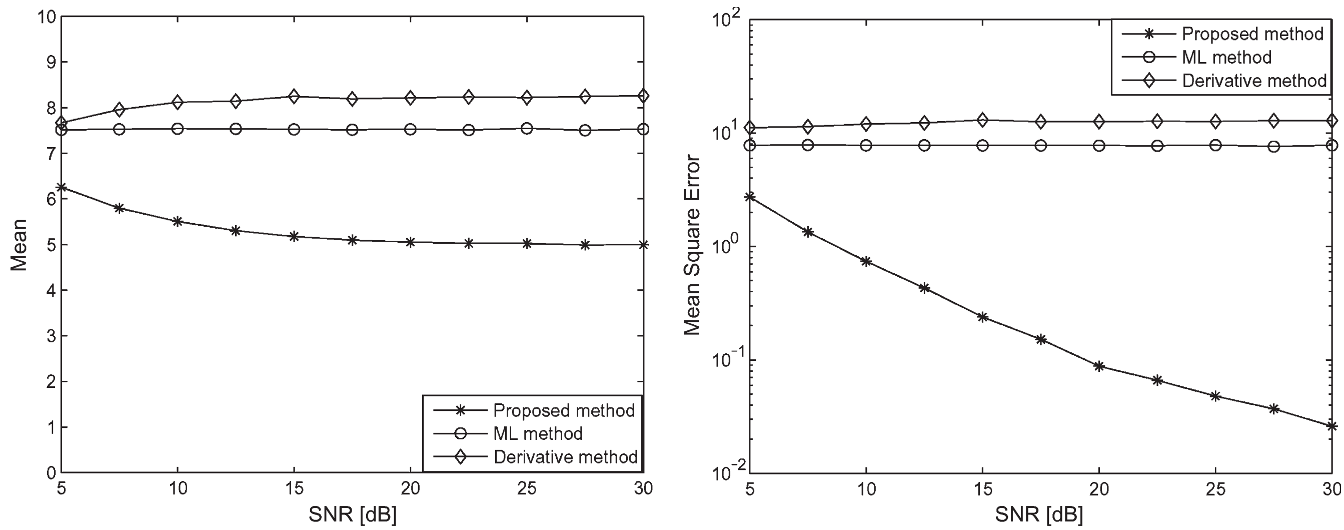


Fig. 2. $\Lambda(k, m)$ ($L = 11$, SNR $= 10$ dB).

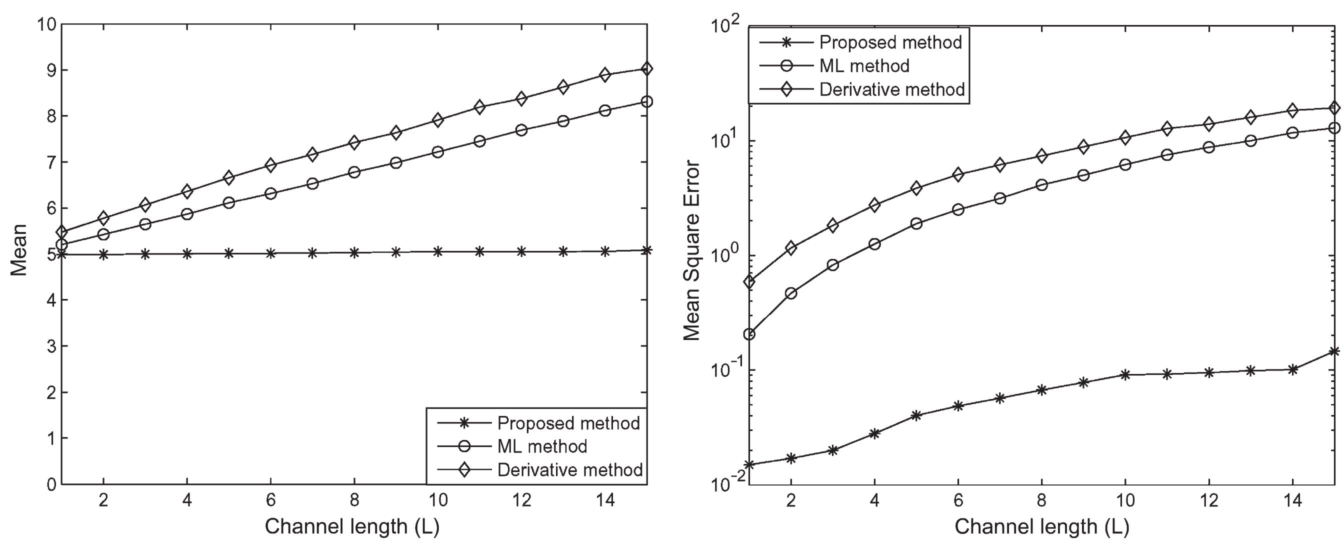
Flat Rayleigh-fading channels ($L = 0$) were first considered, and the channel response $h(0)$ was randomly generated as a complex Gaussian variable with zero mean and a variance of 2. Fig. 1(a) and (b) shows the mean and MSE of the timing offset estimated by various algorithms under consideration. It is obvious that all of these algorithms yield an unbiased estimation of the timing offset, while the MSE for the proposed algorithm is lower than that of the derivative algorithm and is close to that of the ML algorithm. Notice that the ML algorithm is the best under flat-fading channels, and its MSE approaches zero since no interference exists in the received signals.

Next, frequency-selective fading channels with 12 independent sample-spaced Rayleigh-fading taps ($L = 11$) were considered. In detail, the channel responses $h(0)$–$h(11)$ were modeled as independent and complex Gaussian random variables with zero mean and variances of [1.9560, 1.8287, 1.6321, 1.3868, 1.1172, 0.8481, 0.6007, 0.3911, 0.2281, 0.1136, 0.0434, 0.0090], respectively. The function $\Lambda(k, m)$ ($\mathrm{SNR} = 10$ dB) is first shown in Fig. 2, since it is essential for the



(a) (b)

Fig. 3. Performance versus SNR for frequency-selective fading channels ($L = 11$).



(a) (b)

Fig. 4. Performance versus channel length ($\text{SNR} = 20$ dB).

derivation of the proposed algorithm. It is shown that the maximum of $\Lambda(k, m)$ points to the correct timing offset $(k = 5)$, which clearly verifies Property 1. The mean of the timing offset estimated by the three aforementioned methods are then illustrated in Fig. 3(a). It shows that the proposed algorithm is still able to achieve an unbiased estimation of the timing offset, while the ML and derivative algorithms cannot. The MSE performance is finally shown in Fig. 3(b). Obviously, the proposed algorithm has a much smaller MSE than the other two algorithms. In addition, the MSE of the proposed algorithm decreases when SNR increases, while there is little variation in that of the other two algorithms. This can be explained as follows: The dominant factor of the MSE in the proposed algorithm is the additive noise as the maximum of $\Lambda(k, m)$ necessarily points to the correct timing offset. On the other hand, the other two algorithms yield a timing offset estimation falling within θ and $\theta + L$ in frequency-selective fading channels, which mainly depends on channel conditions and is therefore little affected by the SNR.

Finally, the channel length L was varied from 1 to 15 $(\text{SNR} = 20 \text{ dB})$ to further test the effect of frequency-selective fading channels on the synchronization algorithms. Results are shown in Fig. 4(a) and (b). Unbiased estimation of the timing offset is achieved by the proposed algorithm irrespective of channel length, whereas the mean of the estimated timing offset in the ML and derivative algorithms increases linearly with the channel length L. In terms of MSE, the performance of the proposed algorithm slightly degrades when the channel length increases. However, for any given channel length under consideration, the proposed algorithm significantly outperforms the other two algorithms. These results demonstrate that the proposed algorithm is robust to frequency-selective fading channels and can significantly reduce the MSE compared with the ML and derivative algorithms.

V. Conclusion

A CP-based blind symbol synchronization algorithm for OFDM systems has been proposed in this paper. A new timing function in which the maximum necessarily points to the correct timing offset, irrespective of channel conditions, has been designed. The proposed algorithm has overcome fluctuations in the existing CP-based algorithms, particularly when the system is operating under frequency-selective fading channels. Channel information is unnecessary, and there is no restriction on the first path channel response, rendering the proposed algorithm more practical. It has been verified by computer simulations that the proposed algorithm achieves an unbiased estimation and significantly reduces the MSE of the estimated timing offset compared with the existing CP-based algorithms when the system is operating under frequency-selective fading channels.

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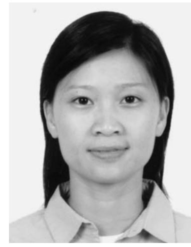

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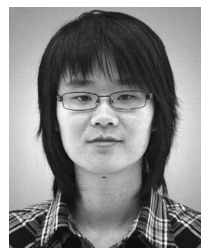

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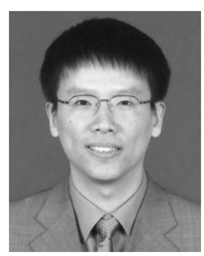

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