# A Novel Synchronization Scheme for OFDM WLANs Using Cyclic-Shifted Preamble

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## ABSTRACT

In this paper, we present a novel cyclic-shifted preamble which is suitable for the synchronization in orthogonal frequency division multiplexing (OFDM) wireless local area networks (WLANs). Based on the novel cyclic-shifted preamble, we proposed two improved symbol timing estimators and a joint estimation method for carrier frequency and sampling clock offsets. It is shown that through simulations, the proposed scheme with low hardware complexity has a significantly improved performance than the other methods when a reasonable signal-to-noise ratio (SNR) is considered.

**Key Words :** OFDM, Synchronization, Symbol, Timing, Frequency, Offset

## Ⅰ. Introduction

Orthogonal frequency division multiplexing (OFDM) is a promising technique for high-speed data transmission over wireless communication channels, since it can provide high bandwidth efficiency and is more robust against frequency selective fading compared to other multiplexing techniques[1]. Due to its evident advantages, OFDM technology has been selected as standards for various wireless applications, such as digital audio broadcasting (DAB), digital video broadcasting (DVB), high performance local area networks (HIPERLAN/2), and IEEE 802.11awirelesslocalareanetworks(WLANs).

As is well known, OFDM systems are very vulnerable to synchronization errors[2,3]. For example, carrier frequency offsets, which are caused by the inherent instabilities of the transmitter and the receiver carrier frequency oscillators, can lead to severe system degradation due to intercarrier interference (ICI). Moreover, symbol timing offsets would cause both intercarrier interference and intersymbol interference (ISI) to degrade the performance of OFDM systems. In contrast to the symbol timing offsets, the sampling clock offsets not only introduce ICI but also result in a drift in the symbol timing which worsens the symbol synchronization problems. Hence, synchronization is one of the most important issues in an OFDM system.

In this paper, we present a novel cyclic-shifted preamble and a complete synchronization scheme based on it. The proposed scheme consists of two improved symbol timing estimators and a joint estimation method for carrier frequency and sampling clock offsets. The synchronization performance by simulations shows that our scheme is better than those of the approaches in [4,7].

This paper is organized as follows: Section 2 introduces the OFDM system model with synchronization errors. In Section 3 a novel cyclic-shifted preamble and corresponding synchronization methods are presented. Simulation results and discussion are provided in Section4. Finally, our work is summarized in Section5.

## Ⅱ. System Description

At the transmitter, the OFDM baseband signal, which is the output of the inverse fast Fourier

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transformation (IFFT), can be expressed in the form:

$$x_{n,i} = \frac{1}{N}\sum_{k=0}^{N-1} X_{k,i} e^{j\frac{2\pi}{N}kn} \quad (1)$$

Where $x_{n,i}$ is the discrete baseband signal of the i-th OFDM symbol at the n-th sample instant, $x_{k,j}$ is the transmitted quadrature amplitude modulation (QAM) symbol at the k-th subcarrier in the i-th OFDM symbol, and N is the fast Fourier transformation (FFT) blocksize.

At the receiver, in the presence of carrier frequency offset and sampling clock offset, the recovered QAM symbol can be written as

$$Y_{m,i} = \frac{1}{N}\sum_{n=0}^{N-1}\sum_{k=0}^{N-1} X_{k,i} H_{k,i} e^{j\frac{2\pi}{N}n(k+k\beta+\varepsilon-m)} e^{j\frac{2\pi}{N}\alpha_i(k\beta+\varepsilon)} + W_{m,i} \quad (2)$$

where $Y_{m,i}$ is the demodulated QAM symbol at the m-th subcarrier in the i-th OFDM symbol, $H_{k,i}$ is the channel response at the k-th subcarrier in the i-th OFDM symbol, $W_{m,i}$ is additive white Gaussian noise, $\varepsilon$ is the carrier frequency offset with respect to the carrier spacing, and $\beta$ denotes the relative sampling clock offset. That is, $\beta = (f_{sr} - f_{st})/f_{st}$ where $f_{sr}$ and $f_{sr}$ represent the transmitter and the receiver sampling clock respectively. The term $e^{j\frac{2\pi}{N}\alpha(k\beta+\varepsilon)}$ denotes the accumulated phase rotation of the first sample in i-th OFDM symbol, while the reference phase, the phase of first sample in the first symbol, is set to zero. And $\alpha_i = (i-1)N + iN_g$, where $N_g$ is the length of cyclic prefix (CP). Further, we can rewrite $Y_{m,i}$ as

$$\begin{aligned}
Y_{m,i} &= \frac{1}{N}\sum_{n=0}^{N-1}\sum_{k=0}^{N-1} X_{k,i} H_{k,i} e^{j\frac{2\pi}{N}n(k+k\beta+\varepsilon-m)} e^{j\frac{2\pi}{N}\alpha_i(k\beta+\varepsilon)} + W_{m,i} \\
&= X_{m,i} H_{m,i} e^{j\frac{\pi}{N}(m\beta+\varepsilon)(2\alpha_i+N-1)} \frac{\sin(\pi(m\beta+\varepsilon))}{N\sin(\frac{\pi}{N}(m\beta+\varepsilon))} \\
&+ \sum_{k=0,k\neq m}^{N-1} X_{k,i} H_{k,i} e^{j\frac{\pi}{N}(k\beta+\varepsilon)(2\alpha_i+N-1)} e^{-j\pi\frac{k-m}{N}} \frac{\sin(\pi(k\beta+\varepsilon))}{N\sin(\frac{\pi}{N}(k+k\beta+\varepsilon-m))} \\
&+ W_{m,i}
\end{aligned} \quad (3)$$

## Ⅲ. Proposed Synchronization

### 3.1 Novel Preamble Structure

In our approach, there are two consecutive time-domain preambles in the front of the OFDM symbols. We express the first OFDM preamble vector to be composed of $(N/N_g+1)N_g$ - dimensional vectors denoted by $x_0^l$ with $l = 0, 1, ..., N/N_g$. So the first preamble $x_0$ (including cyclic prefix) is given by

$$x_0 = [x_0^0 \; x_0^1 \; x_0^2 \ldots x_0^{N/N_g}] \quad (4)$$

where $N_g$is the number of sampling points in CP, $x_0^0 = x_0^{N/N_g}$ represents a vector of CP, and each $x_0^l$ is composed of $N_g$ sampling points.

Then the second preamble $x_1$ which is also composed of $(N/N_g+1)N_g$ - dimensional vectors denoted by $x_1^l$ with $l$=0, 1, ..., $N/N_g$ is designed to be the $(N+N_g)$ - th or $(-N_g)$ - th cyclic-shifted version of the first preamble $x_0$ due to the Time Shift Property in Fourier transform[8]i.e.

$$x(t \pm t_0) \quad \Leftrightarrow \quad X(f)\exp(\pm j2\pi f t_0) \quad (5)$$

So each component of $x_1^l$ can be easily formulated from the first frequency-domain preamble $X_0(k)$ as follows

$$x_1(n) = \frac{1}{\sqrt{N}}\sum_{k=0}^{N-1} X_0(k) e^{j2\pi k(n+N_g)/N}, \quad n = 0,1,\ldots,N-1 \quad (6)$$

From equations (5) and (6), the second preamble $x_1$ (including CP) can be expressed as

$$x_1 = [x_0^1 x_0^2 x_0^3 \ldots x_0^{N/N_g} x_0^1] \quad (7)$$

### 3.2 Symbol Timing Estimation Method A

This method uses the correlation between the two consecutive received preambles $y_0(n)$ and $y_1(n)$. Two correlation windows of length $N_g$ samples points are separated by $2N$ samples. And the correlation function $P_A(d)$ and received energy $R_A(d)$ are respectively defined as

$$P_A(d) = \sum_{k=0}^{N_g-1} y_0(d+k)\cdot y_1^*(d+k+N-N_g) \qquad (8)$$

and

$$R_A(d) = \frac{1}{2}\sum_{m=0}^{1}\sum_{k=0}^{N_g-1} \left|y_m(d+k+m(N-N_g))\right|^2 \qquad (9)$$

where d is a time index corresponding to the first sample in the correlation window, both $P_A(d)$ and $R_A(d)$ can be calculated iteratively.

From equations (8) and (9), the timing metric $M_A(d)$ can be given by

$$M_A(d) = M_A'(d-N_g) \qquad (10)$$

where

$$M_A'(d) = Z_A(d) - Z_A(d-N_g) \qquad (11)$$

and

$$Z_A(d) = \frac{|P_A(d)|^2}{(R_A(d))^2} \qquad (12)$$

Because the sampling point corresponding to the peak value of $M_A'(d)$ is the starting point of the whole OFDM symbol (including CP). So we should shift $M_A'(d)$ by $N_g$ points to get $M_A(d)$.

### 3.3 Symbol Timing Estimation Method B

Method B is based on Minn's sliding window method[5]. We use the same correlation function and received energy function given by equations (8)-(9) as method A. But the timing metric is simply averaged over a window of length $N_g + 1$ sampling points. Then the timing metric $M_B(d)$ is defined as

$$M_B(d) = \frac{1}{N_g+1}\sum_{k=-N_g}^{0} Z_B(d+k) \qquad (13)$$

where

$$Z_B(d) = \frac{|P_A(d)|^2}{(R_A(d))^2} \qquad (14)$$

Fig. 1 shows an example of the timing metrics under no noise and no channel distortion with 64 subcarriers and 16 cyclic prefix. The correct timing point (indexed 0 in the Fig. 1) is taken as the start of useful part of preamble (after cyclic prefix). The proposed timing metrics are compared to those of Schmidl's and Minn's[4,5]. As seen in the Fig. 1, Schmidl's method has a plateau, which will lead some uncertainty in practical case. But the timing metrics from Minn's method and proposed methods reduce the plateau, and yields a sharp timing metric which will help us to get an accurate estimation.

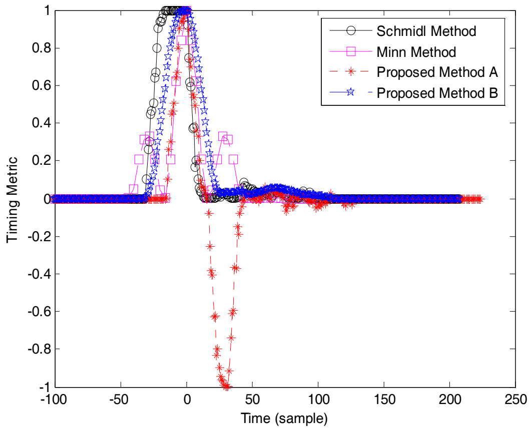


Fig.1. Timing metrics

### 3.4 Joint Carrier Frequency and Sampling Clock Offsets Estimation Method

According to equations (3) and (6), two consecutive preambles $Y_{m,0}$ and $Y_{m,1}$ after demodulation are given as

$$\begin{aligned} Y_{m,0} &= R_m + W_{m,0} \\ Y_{m,1} &= R_m e^{j\frac{2\pi}{N}mN_g} e^{j2\pi(m\beta+\varepsilon)} + W_{m,1} \quad m = 0,1,\ldots,N-1 \end{aligned} \qquad (15)$$

where $R_m$ represents the transmitted first preamble, $W_{m,0}$, $W_{m,1}$ are independent Gaussian noise with zero mean and variance $2\sigma_0^2$. We assume that the channel response function keeps same during the period of the two consecutive preambles.

Based on the Moose's method[6], we can get

$$\hat{\varepsilon}_1 = \frac{1}{2\pi} angle(\sum_{m=0}^{N-1} Y_{m,0}^{*} Y_{m,1}) \tag{16}$$

where $\hat{\varepsilon}_1$ is the estimated carrier frequency offset. If we consider a reasonable signal to noise ratio (SNR) and ignore the product of two independent Gaussian variables, the term $\sum_{m=0}^{N-1} Y_{m,0}^{*} Y_{m,1}$ can be expanded as

$$\sum_{m=0}^{N-1} Y_{m,0}^{*} Y_{m,1} = 2\delta_s^2 e^{j2\pi\varepsilon} e^{j\pi(N-1)(\frac{N_g}{N}+\beta)} \frac{\sin[\pi N(\frac{N_g}{N}+\beta)]}{\sin[\pi(\frac{N_g}{N}+\beta)]} + W_e \tag{17}$$

where $W_e$ is equivalent Gaussian noise, and we assume $R_m^{*} R_m = 2\sigma_s^2$. Approximately, if the estimated method given by equation (16) is applied in the presence of sampling clock offset $\beta$, in the presence of sampling clock offset $\beta$, we can get

$$\hat{\varepsilon}_1 \approx \varepsilon + \frac{N-1}{2}(\frac{N_g}{N} + \beta) \tag{18}$$

For sampling clock synchronization, we proposed a method as follows

$$\hat{\beta}_1 = \frac{1}{2\pi} angle(\sum_{m=0}^{N/2-1} Y_{m,0} Y_{m,1}^{*} Y_{m+N/2,0}^{*} Y_{m+N/2,1}) \tag{19}$$

where $\hat{\beta}_1$ is the estimated sampling clock offset. Considering the same procedure as equation (17), we can get

$$\sum_{m=0}^{N/2-1} Y_{m,0} Y_{m,1}^{*} Y_{m+N/2,0}^{*} Y_{m+N/2,1} = \sum_{m=0}^{N/2-1} R_m R_m^{*} R_{m+N/2}^{*} R_{m+N/2} e^{j\pi(N_g+N\beta)} + W_e \tag{20}$$

Comparing equations (19) and (20), a relationship between $\hat{\beta}_1$ and $\beta$ canbegivenas

$$\hat{\beta}_1 = \frac{N_g}{2} + \frac{N}{2}\beta \tag{21}$$

So after $\hat{\beta}_1$ is acquired, it is easy to get the accurate estimation of sampling clock offset $\hat{\beta}$ by

$$\hat{\beta} = \frac{2}{N}\hat{\beta}_1 - \frac{N_g}{N} \tag{22}$$

The reason why we can get more exact estimation $\hat{\beta}$ is that each product term in equation (19) has eliminated the impact imposed by the carrier frequency offset. This is verified by the simulation results shown in Fig. 2 where SNR=25dB. Here, the relative estimator error is define as $\frac{\left|\hat{\varepsilon}-\varepsilon\right|}{\varepsilon} \times 100\%$, or $\frac{\left|\hat{\beta}-\beta\right|}{\beta} \times 100\%$.

At last, we can get the correct estimation of carrier frequency offset $\hat{\varepsilon}$ by

$$\hat{\varepsilon} = \hat{\varepsilon}_1 - \frac{N-1}{2}(\frac{N_g}{N} + \hat{\beta}) \tag{23}$$

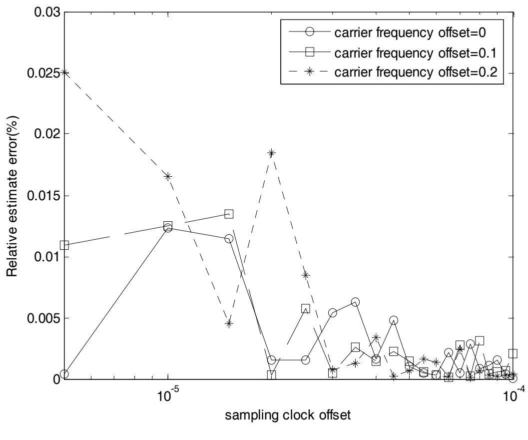


Fig. 2 The relative estimation error of the estimated sampling clock offset by the proposed method in the presence of carrier frequency offset

## Ⅳ. Simulation Results and Discussion

The performance of the proposed synchronization scheme has been investigated by computer simulations. The OFDM system parameters used are 1024 sub-carriers, 128 cyclic prefix and HIPERLAN/2 indoor channel model[9]. The sub-carrier modulation is quaternary phase-shift keying. Unless stated otherwise, 10 000 simulation runs will be applied.

Fig. 3 shows the means for the timing offset estimator in HIPERLAN/2 indoor channel A [9] when carrier frequency offset equals 0.2. We can see that the mean value of Schmidl's method has shifted to almost the middle of the cyclic prefix, whereas the mean value of the proposed two estimators and Minn's estimator is atroughly the correct timing point.

The mean square error (MSE) reflects both the bias and the variance of the estimation. Therefore, the performance of the proposed symbol timing estimation methods can be evaluated by MSE. As seen in Fig. 4, the proposed timing offset estimators under the condition that carrier frequency offset equals 0.2 have a much smaller MSE than the other estimators. Especially, a very accurate estimation can be gotten with our proposed estimators when the SNR is high.

As shown in Fig. 5, our proposed joint carrier frequency and sampling clock offsets estimation method generates a satisfactory estimation of carrier frequency offset, of which the relative estimate error is less than 3%. But for the conventional Moose method, the relative estimator error varies from 5%-53%, where SNR=25dB.

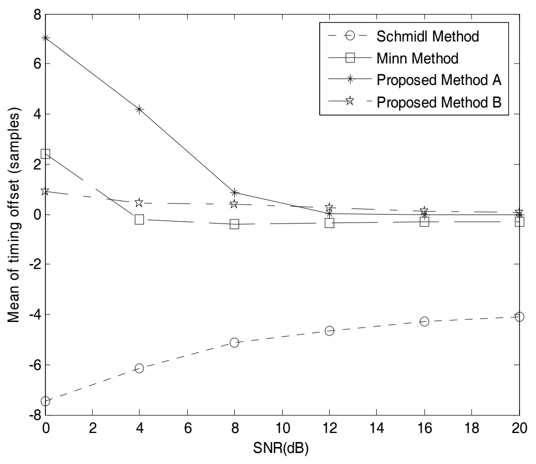


Fig. 3. Mean of timing offset estimators

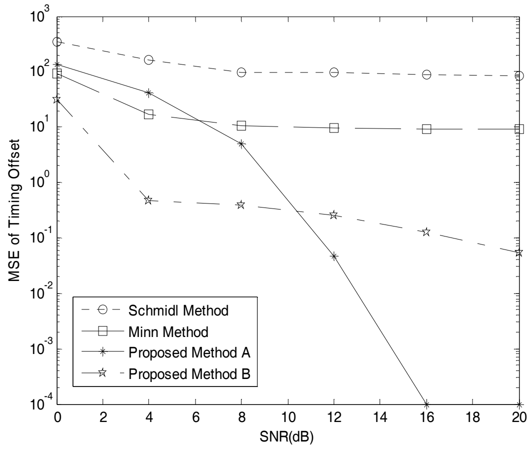


Fig. 4. MSE of timing offset estimators

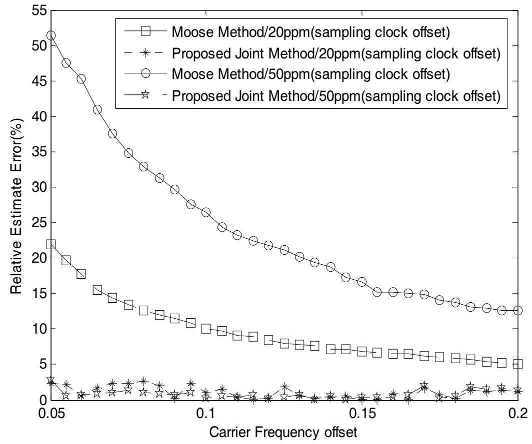


Fig. 5. The comparison of the proposed joint estimation method and the conventional Moose method in the presence of sampling clock offset

## V. Conclusion

A novel synchronization scheme based on the cyclic-shifted preamble was introduced in this paper. Compared with other synchronization methods, we can find that the proposed scheme gives a more accurate estimation and a significantly improved performance when a reasonable SNR is considered. At the same time, we can reduce the hardware complexity greatly because of easy algorithms and no special need for preamble's production. Therefore, the proposed scheme is suitable for improving the performance of the synchronization in orthogonal frequency division multiplexing (OFDM) based wireless local area networks (WLANs).

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