A Novel Time Synchronization Scheme for OFDM Systems Based on Variational Pseudo-Noise Preamble

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***Abstract*—The symbol timing synchronization inaccuracy can bring about phase offset (FO), inter-symbol interference (ISI) and inter-channel interference (ICI). To overcome these shortcomings as much as possible, in this paper, a novel symbol timing synchronizer is proposed for orthogonal frequency division multiplexing (OFDM) systems based on the properties of PN sequences. It is demonstrated by the computer simulations that the proposed scheme achieves accurate timing offset estimation with smaller error mean (EM) and standard deviation (SD) compared with the existing method under flat-fading channels and perform relatively good over frequency-selective fading (FSF) channels.**

I. INTRODUCTION

OFDM has been considered as an effective and promising approach in wireless communication systems and it is very sensitive to nonlinear distortion and synchronization errors. Numerous algorithms have been proposed on the subject of timing synchronization. The non-preamble-aided algorithm in [1] can only make a coarse estimation of symbol timing offset (STO). To further enhance the estimation precision, preamble-based algorithms have been widely researched. The methods in [2] utilize more than two OFDM symbols to realize timing synchronization, which result in a waste of spectrum resources. An OFDM symbol with a form of two identical preambles is used in [3], whereas there exists a plateau in the metric function with the length of N_{cp} samples. The training symbols in [4] are separated from two of the same data blocks into four of the same data blocks and the latter two pieces of blocks are weighted with the minus signs, but the declining curve is not steep enough, which gives rise to timing inaccuracy. Conjugate symmetry preamble is used in [5] which can solve the problem in [4]. Nevertheless, there exists side lobe outside the main lobe. Another estimator was proposed in [6] by employing a constant envelop preamble. Although the accuracy of STO is improved, whereas high precision still can not be achieved by this method.

II. SIGNAL MODEL

The input bit streams are mapped into a series of QPSK or QAM symbols which are then converted into N parallel signals in the transmitter. These parallel signals are modulated onto N different orthogonal subcarriers which are realized by inverse fast fourier transform (IFFT). Hence, the corresponding discrete-time OFDM symbol can be described as

$$\mathrm{x}(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X(k) e^{j2\pi kn/N}, n \in \{0, 1, \ldots, N-1\} \quad (1)$$

where $X(k)$ denotes the data transmitted over the kth subcarrier with a length of T_s. After S/P conversion, the duration of transmission time for useful part of each OFDM symbol is extended to $T = NT_s$. N is the IFFT window size. A cyclic prefix with the length of N_{cp} is inserted into the front of each OFDM symbol to reduce the impact of ISI between two consecutive OFDM symbols. In the receiver, the received signal can be represented as

$$y(n) = e^{j2\pi n\varepsilon/N} \sum_{l=0}^{L-1} h(l) x(n-\theta-l) + \omega(n)$$
$$n \in \{0, 1, \ldots, N+N_{cp}-1\} \quad (2)$$

where $h(l)$ is the channel impulse response. L is a lagged channel response samples including the path with no delay. The symbol timing offset θ and the normalized carrier frequency offset $\varepsilon$are modeled as a delayed signal in the receiver and a phase offset respectively. $\omega(n)$ is a complex Gaussian noise process with zero-mean and the variance of ${\sigma_{\mathrm{n}}}^2$.

III. PROPOSED SYMBOL SYNCHRONIZATION ALGORITHM

To reduce the error probability in the previous methods and to avoid burst wide range deviation and further enlarge the difference of the timing metric value around the correct starting point, we design a new preamble by introducing two different PN sequences with the same period. The double PN including PN1 and PN2 can be defined as

$$S_i, \quad i \in \{0, 1, \ldots, N/2-1\}, Q_i, \quad i \in \{0, 1, \ldots, N/2-1\}$$

The sequences of S_i and Q_i are mapped into $+1$ or -1 and the original discrete-time training sequences have the structure as follows

$$X_i = X_{i+N/4} = -X_{i+N/2} = -X_{i+3N/4}$$
$$i \in \{0, 1, \ldots, N/4-1\} \quad (3)$$

The new preamble can be defined as

$$X'_i = \begin{cases} X_i S_i, & i \in \{0, 1, \ldots, N/2-1\} \\ X_i Q_{i-N/2}, & i \in \{N/2, N/2+1, \ldots, N-1\} \end{cases} \quad (4)$$

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so the proposed preamble are designed to be the form

$$\text{Preamble}_{\text{pro.}} = [\mathrm{A}_{[\mathrm{N}/4]}\mathrm{S}_{[\mathrm{N}/4]} \quad \mathrm{A}_{[\mathrm{N}/4]}\mathrm{S}_{[\mathrm{N}/4]} \\ - A_{[N/4]}Q_{[N/4]} \quad - A_{[N/4]}Q_{[N/4]}]$$

where $A_{[N/4]}$ represents samples of $N/4$ length produced by IFFT of a PN sequence. $S_{[N/4]}$ and $Q_{[N/4]}$ are weighted PN sequences with $N/4$ samples. Note that the PN in the frequency-domain and the time-domain have completely different effects. The former is used to avoid high peak-to-average power ratio (PAPR) and the latter is employed to diminish the two adjacent values of the timing metric because it is shown that they share the same sum of the pairs of product, with the exception of only two product terms in the previous research [3], [4]. Therefore, to enlarge the difference between the two adjacent values of the timing metric and avoid sudden timing shift in [5] and non-robust under low SNR in [6], it is necessary to maximize the different pairs of product between them. Then the timing metric is given by

$$M_{\text{pro.}}(d) = \frac{\left|P_{\text{pro.}}(d)\right|^2}{(R_{\text{pro.}}(d))^2} \tag{5}$$

where

$$\begin{aligned} P_{\text{pro.}}(d) &= \sum_{i=0}^{N/4-1} S_i S_{i+N/4} y^*(d+i) y(d+i+N/4) \\ &+ \sum_{i=0}^{N/4-1} Q_i Q_{i+N/4} y^*(d+i+N/2) y(d+i+3N/4) \end{aligned} \tag{6}$$

$$R_{\text{pro.}}(d) = \frac{1}{2}\sum_{i=0}^{N-1}\left|y(d+i)\right|^2. \tag{7}$$

Obviously, we can see from (6) that the introduction of PN weighted factors ensures the proposed timing metric gets its maximum at the accurate timing point which is taken as the starting position of the useful part of the training symbol, whereas the remaining values at the other sample points are fairly low relative to the correct location. $P_{\text{pro.}}(d)$, and $R_{\text{pro.}}(d)$ can be calculated iteratively. The goal to minimize the correlation value between two partly related blocks and maximize different pairs of product between two adjacent values are achieved by the proposed method. Note that the randomness of PN sequences can be eliminated only at perfect timing position. When the FFT window size gets bigger, the length of PN weighted factors gets longer that closer to the random signal and then the autocorrelation properties of PN sequences becomes very sharp towards impulse form, the system performance will be further improved.

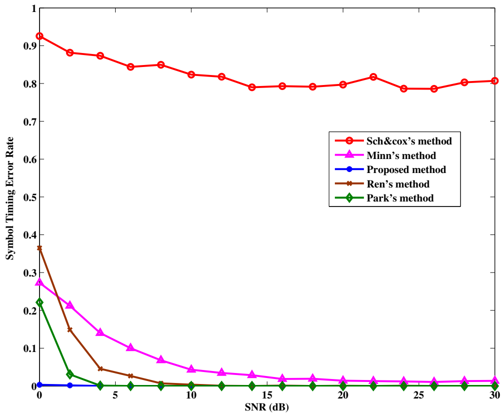

Fig. 1. Symbol timing error rate for the conventional and proposed algorithms versus SNR in flat-fading channels.

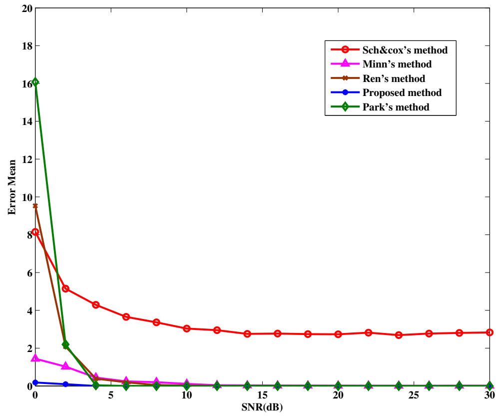

Fig. 2. Error mean of $\left|\hat{\theta} - \theta\right|$ for the conventional and proposed algorithms versus SNR in flat-fading channels.

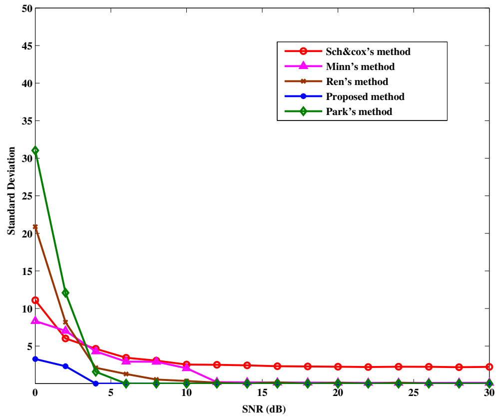

Fig. 3. SD of $\left|\hat{\theta} - \theta\right|$ for the conventional and proposed algorithms versus SNR in flat-fading channels.

IV. SIMULATION RESULTS

An OFDM system with 128 subcarriers $(N = 128)$ and CP length of 16 samples $(N_{cp} = 16)$ is considered to evaluate the performance of proposed estimator. We assume that the symbol timing offset and normalized carrier frequency offset

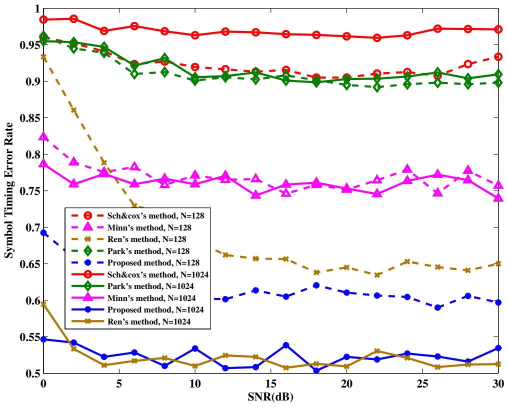

Fig. 4. Symbol timing error rate for the conventional and proposed algorithms versus SNR in frequency-selective fading channels ($L = 5$, $N = 128$ and $N = 1024$).

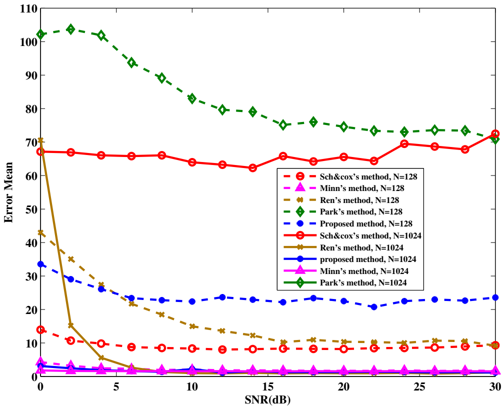

Fig. 5. Error mean of $\left|\hat{\theta} - \theta\right|$ for the conventional and proposed algorithms versus SNR in frequency-selective fading channels ($L = 5$, $N = 128$ and $N = 1024$).

are θ=3 and $\varepsilon = 0.2$, respectively. The system performance is evaluated in terms of symbol timing error rate, error mean and standard deviation. The error mean is the arithmetic average value of the error $\left|\hat{\theta} - \theta\right|$ and the standard deviation is the square root of the variance of $\left|\hat{\theta} - \theta\right|$ that used to measure the amount of variation or dispersion from the mean value. The algorithms in [3]- [6] were also implemented for comparison purpose. As seen in Fig. 1, the proposed algorithm performs better than the others especially at a low SNR because the further reduction of the influence of the side-peaks besides the correct position. The EM and SD performance are shown in Fig. 2 and Fig. 3 respectively. It indicates that the proposed method is almost the same as Park and Ren when the SNR is greater than 10 dB but greatly outperforms the others when SNR less than it. We do not use logarithmic coordinates in Fig. 3 for the reason that the proposed scheme achieves a zero SD except a few points in the beginning. The system performance is further tested when under frequency-selective fading channels as shown in Fig. 4, Fig. 5, and Fig. 6. Although there are some unsatisfactory results when the FFT window size is small, the performance is relatively better than others with the increases of FFT size in accordance with 3GPP LTE specifications [7].

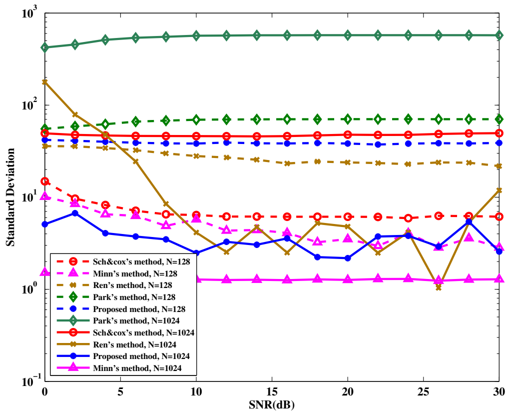

Fig. 6. SD of $\left|\hat{\theta} - \theta\right|$ for the conventional and proposed algorithms versus SNR in frequency-selective fading channels ($L = 5$, $N = 128$ and $N = 1024$).

V. CONCLUSION

A novel symbol timing synchronizer has been presented for OFDM systems. Simulations show that the proposed algorithm achieves precise timing offset estimation compared with the existing methods under flat-fading channels and shows good performance under FSF channels when FFT window is large.

VI. ACKNOWLEDGEMENT

This work was partly supported by the grant of scientific research No.26420339 from JSPS.

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