무선 이동 네트워크에서 OFDM 시스템을 위한 적은 계산량의 타이밍 추정

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A Novel Timing Estimation with Low-Complexity for OFDM Systems in Mobile Wireless Networks

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요 약

In this paper, we present a novel timing estimation method with low complexity for orthogonal frequency division multiplexing systems in mobile fading channels. The proposed method is aimed to reduce the complexity of conventional methods in mobile fading channels. Our training sequences which have low peak-to-average power ratio consist of the complementary sets.

I. Introduction

Orthogonal frequency division multiplexing(OFDM) systems are very sensitive to symbol timing and carrier frequency offset [1]. Several approaches have been proposed to estimate time and frequency offset either jointly or individually[2]-[4]. The most popular of the pilot-aided algorithms is the method proposed by Schmidl[5]. His method uses a preamble containing the same two halves to estimate the symbol timing and frequency offset. Schmidl's estimator provides simple and robust estimates for symbol timing and carrier frequency offset. However, the timing metric of Schmidl's method has a plateau, which causes a large variance in the timing estimate. To reduce the uncertainty arising from the timing metric, Minn proposed a method as a modification to Schmidl's approach[6]. Park proposed a new timing synchronization method for OFDM timing estimation which produces an even sharper timing than Schmidl's and Minn's[7]. But these methods require very high complexity in mobile wireless networks. In this paper we proposed a novel preamble structure using complementary sets. By utilizing the property of expansion of complementary set an inverse discrete Fourier transform, we can achieve timing estimation with a much lower complexity. In Section II we introduce the OFDM signal model and describe the existing timing estimation methods. Section III covers the proposed preamble and timing offset estimation method, in Section IV, the performances of the proposed estimator and the other estimator is compared in terms of mean-square error using computer simulation results, finally conclusion is drawn in Section V.

II. System Description

A. OFDM Signal Description

Consider a general case of a OFDM system, using the standard complex-valued baseband equivalent signal model. The nth received sample has the standard form

$$y[n]=\sum_{k=0}^{L-1}h[k]x[n-k] \tag{1}$$

where $h[n]$ is the channel impulse response, whose memory is denoted by L . $x[n]$ is the time-domain OFDM signal expressed by

$$x[n]=\sum_{k=0}^{N-1}c_k\exp\left(j2\pi kn/N\right) \tag{2}$$

where N is the number of sub-carriers and the c_k's are the complex information symbols. At the receiver, timing offset is modeled as a delay in the received signal and frequency offset is modeled as a phase distortion of the received data in the time domain. These two uncertainties and the AWGN $w[n]$ yield the received signal

$$r[n] = y[n - n_\varepsilon]\exp\left(2\pi\theta_\varepsilon n / N + \phi\right) + w[n] \tag{3}$$

where n_ε is the integer-valued unknown arrival time of a symbol, θ_ε is the frequency offset and ϕ is the initial phase.

B. OFDM Timing Synchronization

The goal of OFDM timing synchronization is to estimate n_ε. Before we proceed, let us briefly describe the timing offset estimation methods presented in [6] and [7].

1) Minn's Method: In order to alleviate the uncertainty caused by the timing metric plateau and to improve the timing offset estimation, Minn proposed a modified preamble. Minn's preamble has the following form:

$$P_{\text{Minn}} = \left[\mathbf{B}_{N/4} \quad \mathbf{B}_{N/4} \quad -\mathbf{B}_{N/4} \quad -\mathbf{B}_{N/4}\right] \tag{4}$$

where $\mathbf{B}_{N/4}$ represents a PN sequence of length $N/4$. Then the timing metric is expressed as

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

where

$$R_{\text{Minn}}(d) = \sum_{m=0}^{1}\sum_{k=0}^{N/4-1} r^*\left(d + k + \frac{N}{2}m\right)\cdot r\left(d + k + \frac{N}{2}m + \frac{N}{4}\right) \tag{6}$$

$$P_{\text{Minn}}(d) = \sum_{m=0}^{1}\sum_{k=0}^{N/4-1} \left| r\left(d + k + \frac{N}{2}m + \frac{N}{4}\right)\right|^2 \tag{7}$$

In Schmidl's method, the timing metric has its peak for the entire interval of the cyclic prefix. The Minn's method has its peak at the correct starting point for the OFDM symbol, since correlation of some samples results in negative values. For this reason, Minn's method eliminates the peak plateau of the timing metric, hence resulting in a smaller MSE.

2) Park's Method: In spite of the reduction of the timing metric plateau, it is observed that the MSE of Minn's estimator is quite large in ISI channels from the results in [6]. In order to increase the performance of the estimator, Park proposed a preamble and correlation method to obtain an impulse-shaped timing metric. The samples of the preamble are designed to be of the form

$$P_{\text{Park}} = \left[\mathbf{C}_{N/4} \quad \mathbf{D}_{N/4} \quad \mathbf{C}^*_{N/4} \quad \mathbf{D}^*_{N/4}\right] \tag{8}$$

where $\mathbf{C}_{N/4}$ represents samples of length $N/4$ generated by IFFT of a PN sequence, and $\mathbf{C}^*_{N/4}$ represents a conjugate of $\mathbf{C}_{N/4}$. To get impulse-shaped timing metric, $\mathbf{D}_{N/4}$ is designed to be symmetric with $\mathbf{C}_{N/4}$. The timing metric is expressed as

$$M_{\text{Park}}(d) = \frac{\left|R_{\text{Park}}(d)\right|^2}{\left(P_{\text{Park}}(d)\right)^2} \tag{9}$$

where

$$R_{\text{Park}}(d) = \sum_{k=0}^{N/2} r(d-k)\cdot r(d+k) \tag{10}$$

$$P_{\text{Park}}(d) = \sum_{k=0}^{N/2} \left| r\left(d+k\right)\right|^2 \tag{11}$$

Park's method produces an even sharper timing than Schmidl's and Minn's. But these methods above mentioned require very high complexity in mobile wireless environments.

III. PROPOSED SYMBOL TIMING METHOD

We propose a novel preamble structure using complementary sets. By utilizing the property of expansion of complementary set an inverse discrete Fourier transform, we can achieve timing estimation with a much lower complexity. The proposed preamble are designed as follow

$$P_{\text{Pro}} = \left[\overline{\mathbf{A}}_{N/8}\ \overline{\mathbf{B}}_{N/8}\ \overline{\mathbf{C}}_{N/8}\ \overline{\mathbf{D}}_{N/8}\ \overline{\mathbf{A}}_{N/8}\ \overline{\mathbf{B}}_{N/8}\ \overline{\mathbf{C}}_{N/8}\ \overline{\mathbf{D}}_{N/8}\right] \tag{12}$$

where $\overline{\mathbf{A}}_{N/8}$, $\overline{\mathbf{B}}_{N/8}$, $\overline{\mathbf{C}}_{N/8}$ and $\overline{\mathbf{D}}_{N/8}$ represent samples of length $N/8$ generated by IFFT of a complementary sets, respectively. Let us describe how to generate the one of the four complementary sets. First above all we have to obtain the basis complementary sets $\boldsymbol{\alpha}_1$ and $\boldsymbol{\alpha}_2$ which consist of two elements, then the mate of these complementary sets can be obtained by

$$\boldsymbol{\beta}_1 = \tilde{\boldsymbol{\alpha}}_2^*, \quad \boldsymbol{\beta}_2 = -\tilde{\boldsymbol{\alpha}}_1^*. \tag{13}$$

where $\tilde{\alpha}_i$ represents the reverse sequence of $\boldsymbol{\alpha}_i$. From these sets, we can obtain the expanded complementary sets as follow

$$\boldsymbol{\chi}_1 = \boldsymbol{\alpha}_1 \otimes \boldsymbol{\beta}_1, \quad \boldsymbol{\chi}_2 = \boldsymbol{\alpha}_2 \otimes \boldsymbol{\beta}_2. \tag{14}$$

where $\otimes$ represents the interleaving expansion between $\boldsymbol{\alpha}_i$ and $\boldsymbol{\beta}_i$ [8]~[10]. After $\boldsymbol{\alpha}_1 = \boldsymbol{\chi}_1$ and $\boldsymbol{\alpha}_2 = \boldsymbol{\chi}_2$, until the number of elements of $\boldsymbol{\chi}_i$ is $N/16$, this expansion is repeatedly conducted. We can obtain the complementary sets which are mutually orthogonal as below

$$\begin{aligned} \boldsymbol{\varphi}_1 &= \boldsymbol{\chi}_1\boldsymbol{\chi}_2 \\ \boldsymbol{\varphi}_2 &= \boldsymbol{\chi}_1(-\boldsymbol{\chi}_2) \\ \boldsymbol{\varphi}_3 &= \tilde{\boldsymbol{\varphi}}_2 = -\tilde{\boldsymbol{\chi}}_2\tilde{\boldsymbol{\chi}}_1 \\ \boldsymbol{\varphi}_4 &= -\tilde{\boldsymbol{\varphi}}_1 = (-\tilde{\boldsymbol{\chi}}_2)(-\tilde{\boldsymbol{\chi}}_1). \end{aligned} \tag{15}$$

Finally $\bar{\mathbf{A}}_{N/8}$, $\bar{\mathbf{B}}_{N/8}$, $\bar{\mathbf{C}}_{N/8}$ and $\bar{\mathbf{D}}_{N/8}$ are obtained by

$$\begin{aligned} \bar{\mathbf{A}}_{N/8} &= \mathrm{IFFT}[\boldsymbol{\varphi}_1] \\ \bar{\mathbf{B}}_{N/8} &= \mathrm{IFFT}[\boldsymbol{\varphi}_3] \\ \bar{\mathbf{C}}_{N/8} &= \mathrm{IFFT}[\boldsymbol{\varphi}_2] \\ \bar{\mathbf{D}}_{N/8} &= \mathrm{IFFT}[\boldsymbol{\varphi}_4]. \end{aligned} \tag{16}$$

Our training sequence has the following property which arise during the generating processes

$$\begin{aligned} \bar{\mathbf{A}}_{\mathrm{L}} &= \bar{\mathbf{B}}_{\mathrm{R}} \\ \bar{\mathbf{A}}_{\mathrm{R}} &= \bar{\mathbf{B}}_{\mathrm{L}} \\ \bar{\mathbf{C}}_{\mathrm{L}} &= \bar{\mathbf{D}}_{\mathrm{R}} \\ \bar{\mathbf{C}}_{\mathrm{R}} &= \bar{\mathbf{D}}_{\mathrm{L}}. \end{aligned} \tag{17}$$

where $(\cdot)_{\mathrm{L}}$ represents the left-sided half of $(\cdot)$ and $(\cdot)_{\mathrm{R}}$ represents the right-sided half of $(\cdot)$.

The proposed timing metric is expressed as

$$M_i(d) = \frac{|R_i(d)|^2}{(P_i(d))^2}, \quad 0 \le i \le 3, \ 0 \le d \le L_b - 1 \tag{18}$$

Where

$$R_i(d) = \sum_{m=0}^{1}\sum_{k=0}^{L_b/2-1}\left\{\begin{aligned} & r(d+d_1+k)\,r^*\left(d+d_1+\frac{3L_b}{2}+k\right)+ \\ & r\left(d+d_2+\frac{L_b}{2}+k\right)r^*(d+d_2+L_b+k) \end{aligned}\right\} \tag{19}$$

$$P_i(d) = \sum_{m=0}^{1}\sum_{k=0}^{L_b-1}|r(d+d_1+k+L_b)|^2 \tag{20}$$

where $L_b = N/8$, $d_1 = 2L_b(m+2)$, $d_2 = 2L_b(2m+1)$. From timing metric, we search the maximum frame point as follow

$$[\eta, \kappa] = \arg\max_{0 \le i \le 3,\ 0 \le d \le L_b - 1} M_i(d) \tag{21}$$

Then the maximum point can be obtained by

$$\hat{d} = \frac{N}{2}\eta + \kappa \tag{22}$$

The timing metric for the final timing estimate point can be rewritten by

$$M_l^{\mathrm{CP}} = \frac{|R_l^{\mathrm{CP}}|^2}{(P_l^{\mathrm{CP}})^2}, \quad 0 \le l \le 2 \tag{23}$$

Where

$$R_l^{\mathrm{CP}} = \sum_{k=0}^{N_G-1} r^*(\hat{d}+k-2L_b l-N_G)\cdot r\left(\hat{d}+k-2L_b l-N_G+N\right) \tag{24}$$

$$P_l^{\mathrm{CP}} = \sum_{k=0}^{N_G-1}\left|r\left(\hat{d}+k-2L_b l-N_G+N\right)\right|^2 \tag{25}$$

After searching the maximum value, we can determine the timing estimate as follow

$$\bar{d} = \hat{d} - 2L_b l - N_G \tag{26}$$

From these processes, our proposed method has eventually much lower complexity than other conventional methods.

IV. SIMULATION RESULTS

The performance of the proposed estimator is evaluated by computer simulations. 1) Simulation I: Carrier frequency is 2.3GHz and the channels bandwidth is 10MHz which is divided equally among 1024 tones and CP size is 128 and RMS channel delay is 2.5μs. 2) Simulation II: Carrier frequency is 2GHz and the channels bandwidth is 20MHz which is divided equally among 2048 tones and CP size is 256 and RMS channel delay is 2.5 μs. The channel is based on COST 207[9]. We assume that frequency offset is 0. The performance of the proposed timing estimator is evaluated

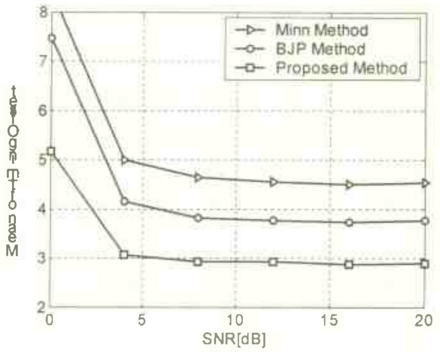

Figure 1. Performance in simulation I

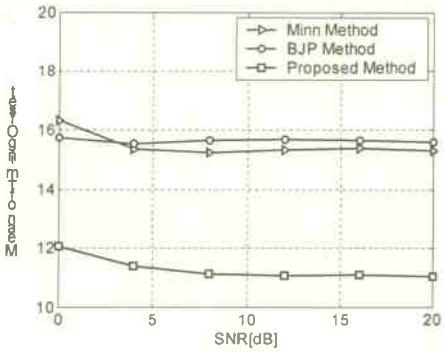

Figure 2. Performance in simulation II

by mean, and is compared with those of Minn's and Park's methods. From the several simulations, we can obtain the training sequences whose PAPR is about 3dB or 2.3dB. Figs. 1, 2 and 3 show the means for the timing offset estimators in mobile fading channels. The various simulation results make it clear that the performance of the proposed estimator performs better than the other two estimators in mobile fading environments.

V. CONCLUSIONS

A training sequence and a timing offset estimator are presented in this paper. The proposed timing offset estimator reduces the complexity. The proposed timing synchronization method makes it possible to estimate a more accurate timing offset. In mobile environments, our method with lower complexity is very powerful in fast timing estimation.

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