Combined Time Synchronization And Channel Estimation For MB-OFDM UWB Systems

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Abstract

Symbol timing error amounts to a major degradation in the system performance. Conventionally, timing error is estimated by predefined preamble on both transmitter and receiver. The maximum of the correlation result is considered the start of the OFDM symbol. Problem arises when the prime path is not the strongest one. In this paper, we propose a new combined time and channel estimation method for multi-band OFDM ultra wide-band (MB-OFDM UWB) systems. It is assumed that a coarse timing has been obtained at a stage before the proposed scheme. Based on the coarse timing, search interval is set (or time candidates). Exploiting channel statistics that are assumed to be known by the receiver, we derive a maximum a posteriori estimate (MAP) of the channel impulse response. Based on this estimate, we discern for the timing error. Timing estimation performance is compared with the least squares (LS) channel estimate in terms of mean squared error (MSE). It is shown that the proposed timing scheme is lower in MSE than the LS method.

Keywords: Time synchronization, timing error, MB-OFDM UWB

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1. Introduction

OFDM has been receiving a great interest and it is adopted in many standards such as [1] as a base-band modulation technique in the form of MB-OFDM UWB to achieve rates of 110 Mbits/s at a distance of 10 meters. The Multi-Band OFDM system divides the UWB spectrum (3.1 to 10.6 GHz) into 528 MHz wide sub-bands and uses OFDM modulation to transmit the information in each sub-band.

Due to the sensitivity of OFDM signal against the timing error, a reliable synchronization algorithm is required in order to provide reliable communication in harsh channel conditions. The synchronization algorithms in literatures are mainly based on sending a training sequence which has a periodic property. At the receiver, autocorrelation is performed to detect the start of the OFDM symbol through maximum value searching.

In this paper, we propose combined time synchronization and channel estimation suitable to work in harsh channel environment such as MB-OFDM UWB systems channel models [2] in an attempt to enhance the MSE. We assume in our work, as in the work in [3], that the time frequency code [1] (TFC) is already known. The main difference between our work and the work in [3] is that there is no clue about the exact start time of the FFT window. Therefore, the received signal model is considered over one band as the receiver is not aware of the exact time to hop its frequency. In other words, we are unable to receive the signal in three bands unless we are aware of the exact symbol start time.

This paper is organized as follows. In section II, the system model is presented including IEEE 802.15-03 systems [1] preamble and channel models [2]. In section III, the proposed algorithm is explained. Performance analysis is given in section IV. Finally, section V is dedicated to the conclusion.

2. System Model

2.1 System Preamble Structure

The Physical Layer Convergence Protocol (PLCP) preamble in the IEEE 802.15-03 proposal consists of three parts [1]: Packet Synchronization (PS) sequence, Frame Synchronization (FS) sequence, and Channel Estimation (CE) sequence. PS sequence consists of 21 repeated periods denoted as PS0, PS1, …, PS20 as shown in **Fig. 1**.

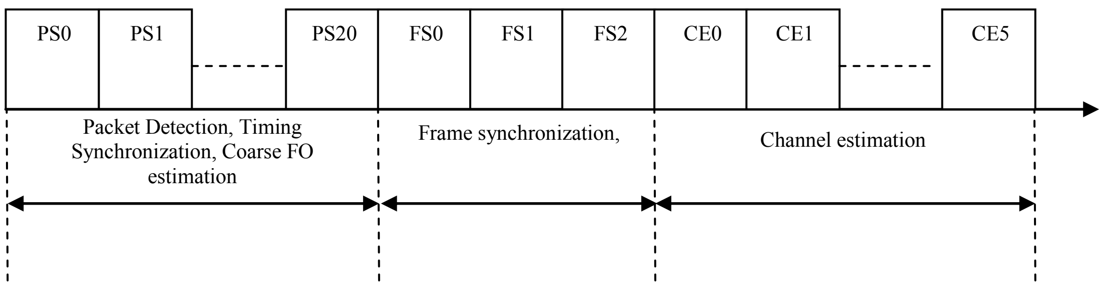

Fig. 1. Preamble structure in [1]

Each piconet uses a distinct time domain sequence (four piconets). After repetition, each period is pre-appended by 32 zero samples and by appending a guard interval of 5 zero samples resulting in M = 165 samples of one PS period. This portion of the preamble can be used for packet detection, coarse NFO estimation and coarse symbol timing. The FS sequence part consists of 3 periods denoted as FS0, FS1, and FS2. Each FS period is a PS but with negative sign. This part of the preamble can be used for frame synchronization, and Time Frequency Code (TFC) identification. Finally, the CE sequence consists of 6 CE periods denoted as CE0, CE1, …, CE5. This part is used to estimate the channel frequency response, fine frequency offset estimation, and fine symbol timing.

2.2 The Received Signal Model

After the transmitted signal is time-frequency interleaved in three bands, a coarse time estimate n_c is obtained at the receiver in earlier stage. Therefore, the received baseband signal (PLCP preamble) during the preamble period in the qth band is described as:

$$r_q(n) = e^{\frac{j2\pi\vartheta_q n}{N}} \sum_{l=0}^{L-1} h_{q,l} s(n - l - n_\epsilon) + w_q(n)$$
$$N_{min} \le n \le N_{max} \qquad (1)$$

where $h_{q,l}$ is the amplitude of the channel lth path in the qth band with variance $\sigma_{h_l}^2$, ϑ_q is the carrier frequency offset (CFO) normalized by the subcarrier spacing in the qth band, $s(n)$ is the OFDM nth sample of the PLCP preamble in time domain, n_ϵ is the time error, and $w_q(k)$ is a zero-mean complex Gaussian noise term in the qth band with variance σ_W^2. We assume that the received signal is of length confined to the interval $[N_{min}, N_{max}]$, where N_{min} and N_{max} are integers depends on the variance of n_c such that $n_c = \frac{N_{max} - N_{min}}{2} + N_{min}$.

We assume that the reception is only upon one band because the exact time of any symbol is unknown. As a result, the receiver is unaware of the time that it has to hop its frequency. Hence, the subscript q is omitted as we assume that we work only on one band. Therefore, (1) can be expressed in matrix notation as follows

$$\mathbf{r}_\epsilon = diag(\mathbf{d}_{\vartheta,\epsilon})\mathbf{S}_\epsilon \mathbf{h} + \mathbf{w}_\epsilon \qquad (2)$$

where $\epsilon = n - n_\epsilon$, and

$$\mathbf{S}_\epsilon = \begin{bmatrix} s_\epsilon & s_{\epsilon-1} & \cdots & s_{\epsilon-L} \\ s_{\epsilon+1} & s_\epsilon & \cdots & s_{\epsilon+1-L} \\ \vdots & \vdots & \ddots & \vdots \\ s_{\epsilon+N-1} & s_{\epsilon+N-2} & \cdots & s_{\epsilon+N-L} \end{bmatrix} \qquad (3)$$

$\mathbf{r}_\epsilon = [r_\epsilon \; r_{\epsilon+1} \; \cdots \; r_{\epsilon+N-1}]^T$, $\mathbf{h} = [h_0 \; h_1 \; \cdots \; h_{L-1}]^T$, $\mathbf{d}_{\vartheta,\epsilon} = [e^{\frac{j2\pi\vartheta\epsilon}{N}} \; e^{\frac{j2\pi\vartheta(\epsilon+1)}{N}} \; \cdots \; e^{\frac{j2\pi\vartheta(\epsilon+N-1)}{N}}]$, $\mathbf{w}_\epsilon = [w_\epsilon \; w_{\epsilon+1} \; \dots \; w_{\epsilon+N-1}]^T$, and $diag(\mathbf{x})$ is a diagonal matrix formed by the vector $\mathbf{x}$. Note that $S_n = 0$ if $n \notin I$, and I is the set of all the points of the same band. Therefore, when $n = n_\epsilon = 0$ we have

$$\mathbf{S} = \begin{bmatrix} s_0 & 0 & \cdots & 0 \\ s_1 & s_0 & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ s_{N-1} & s_{N-2} & \cdots & s_{N-L} \end{bmatrix} \tag{4}$$

Another received signal model can be formulated as $\ddot{\mathbf{r}}_\epsilon = [\mathbf{r}_\epsilon;\ \mathbf{r}_{\epsilon+KM}]$ where K is the number of symbols separating two periods in one band. $K = 3$ for TFC number 1 and 2. Ignoring the frequency offset error, this can be written as $\ddot{\mathbf{r}}_\epsilon = \ddot{\mathbf{S}}_\epsilon \mathbf{h} + \mathbf{w}_\epsilon$ where $\ddot{\mathbf{S}}_\epsilon = [\mathbf{S}_\epsilon;\ \mathbf{S}_\epsilon]$, $\mathbf{w}_\epsilon$ is the corresponding noise term. Also, when $n = n_\epsilon = 0$, $\ddot{\mathbf{S}} = [\mathbf{S};\ \mathbf{S}]$. In the upcoming discussion, the timing error ϵ will be incorporated in the channel vector $\mathbf{h}$ rather than $\mathbf{S}$, thus the channel vector will be denoted as $\mathbf{h}_\epsilon$.

3. Proposed Timing Estimation

In this section henceforth, it is assumed that the frequency error had been estimated and compensated. Hence, $\vartheta = 0$. The log-likelihood function [4] given the timing error ϵ is expressed as

$$l_{\mathbf{h}}(\mathbf{h}_{\tilde{\epsilon}}) = C_1 + \hat{\mathbf{h}}_{\tilde{\epsilon}}^T \boldsymbol{\Lambda}_{\mathbf{h}}^{-1} \hat{\mathbf{h}}_{\tilde{\epsilon}} \tag{5}$$

where C_1 is a constant independent of ϵ, $\tilde{x}$ is a trial value of the random variable x, $\hat{\mathbf{h}}_\epsilon$ is an estimate of h_ϵ, and Λ_{h} is a diagonal matrix of the elements $[\sigma_{\mathrm{h}_0}^2\ \sigma_{\mathrm{h}_1}^2\ \cdots\ \sigma_{\mathrm{h}_{L-1}}^2]$. The maximum likelihood timing estimation is given as

$$\hat{\epsilon} = \max_{\epsilon} l_{\mathbf{h}}(\mathbf{h}; \tilde{\epsilon}) \tag{6}$$

From (6) it is apparent that to obtain an estimate of ϵ the channel vector $\mathbf{h}$ should be available, which is not possible always. Therefore, we will derive the MAP estimator of the channel vector $\mathbf{h}_\epsilon$, i.e. $\check{\mathbf{h}}_{\tilde{\epsilon}}$, and we substitute it in (6) to obtain the maximum likelihood timing estimation. The MAP equation of $\mathbf{h}_\epsilon$ given $\mathbf{r}$ is [4]

$$l_{\mathbf{h}_\epsilon;\mathbf{r}}\left(\tilde{\mathbf{h}}_{\tilde{\epsilon}}; \ddot{\mathbf{r}}\right) = l_{\mathbf{r};\mathbf{h}_\epsilon}\left(\ddot{\mathbf{r}}; \tilde{\mathbf{h}}_{\tilde{\epsilon}}\right) + l_{\mathbf{h}}\left(\tilde{\mathbf{h}}_{\tilde{\epsilon}}\right) - l_{\mathbf{r}}(\ddot{\mathbf{r}}) \tag{7}$$

where

$$l_{\mathbf{r};\mathbf{h}_\epsilon}\left(\mathbf{r}; \tilde{\mathbf{h}}_{\tilde{\epsilon}}\right) = C_2 + \left(\ddot{\mathbf{r}} - \ddot{\mathbf{S}}\tilde{\mathbf{h}}_{\tilde{\epsilon}}\right)^{\mathbf{T}} \boldsymbol{\Lambda}_{\ddot{\mathbf{r}}}^{-1} (\ddot{\mathbf{r}} - \ddot{\mathbf{S}}\tilde{\mathbf{h}}_{\tilde{\epsilon}}) \tag{8}$$

C_2 is constant independent of $\mathbf{h}_\epsilon$, $l_{\mathbf{r}}(\ddot{\mathbf{r}})$ is the logarithm of the probability density function of the received signal, and since it is independent of $\mathbf{h}_\epsilon$, it will be omitted in the next derivations, and $\boldsymbol{\Lambda}_{\ddot{\mathbf{r}}}$ is the covariance matrix of size $2N \times 2N$ of the received signal $\ddot{\mathbf{r}}$ given $\mathbf{h}_{\tilde{\epsilon}}$, hence, it is a diagonal matrix of the elements $[\sigma_w^2 \cdots \sigma_w^2]$

We can see that the timing error is incorporated in the channel vector rather than the matrix $\ddot{\mathbf{S}}$. $\check{\mathbf{h}}_{\tilde{\epsilon}}$ is given by solving

$$\frac{\partial l_{\mathbf{h}_\epsilon;\mathbf{r}}\left(\tilde{\mathbf{h}}_{\tilde{\epsilon}}; \ddot{\mathbf{r}}\right)}{\partial \tilde{h}_{l,\tilde{\epsilon}}} = 0 \tag{9}$$

After straight forward mathematical derivations of (7) with respect to $\tilde{\mathrm{h}}_{l,\breve{\epsilon}}$, we obtain

$$\underbrace{\begin{bmatrix} \mathbf{z}_0^T \ddot{\mathbf{S}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-1} \ddot{\mathbf{S}} + \mathbf{z}_0^T \mathbf{\Lambda}_{\mathbf{h}}^{-1} \\ \mathbf{z}_1^T \ddot{\mathbf{S}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-1} \ddot{\mathbf{S}} + \mathbf{z}_1^T \mathbf{\Lambda}_{\mathbf{h}}^{-1} \\ \vdots \\ \mathbf{z}_{L-1}^T \ddot{\mathbf{S}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-1} \ddot{\mathbf{S}} + \mathbf{z}_{L-1}^T \mathbf{\Lambda}_{\mathbf{h}}^{-1} \end{bmatrix}}_{\ddot{\mathbf{A}}} \mathbf{h}_{\breve{\epsilon}} = \ddot{\mathbf{S}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-1} \ddot{\mathbf{r}} \tag{10}$$

where $\mathbf{z}_i$ is a zeros row vector with the ith element equals one, and $\mathbf{z}_i^T \ddot{\mathbf{S}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-1} \ddot{\mathbf{S}} + \mathbf{z}_i^T \mathbf{\Lambda}_{\mathbf{h}}^{-1}$ is a row vector of L elements. The matrix $\ddot{\mathbf{A}}$ can be written as $\ddot{\mathbf{S}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-1} \ddot{\mathbf{S}} + \mathbf{\Lambda}_{\mathbf{h}}^{-1}$. This leads to

$$\check{\mathbf{h}}_{\breve{\epsilon}} = \ddot{\mathbf{A}}^{-1} \ddot{\mathbf{S}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-1} \ddot{\mathbf{r}} \tag{11}$$

The MSE of the channel estimation is obtained as follows

$$MSE = E\left[\left|\mathbf{h} - \check{\mathbf{h}}\right|^2\right] = E[\mathbf{h}^T \mathbf{h}] - 2E\left[\mathbf{h}^T \check{\mathbf{h}}\right] + E\left[\check{\mathbf{h}}^T \check{\mathbf{h}}\right] \tag{12}$$

where $E\left[\mathbf{h}^T \check{\mathbf{h}}\right] = E\left[\mathbf{h}^T \mathbf{A}^{-1} \ddot{\mathbf{S}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-\mathbf{1}} (\mathbf{S}\mathbf{h} + \mathbf{w})\right]$, keeping in mind that the noise and the channel are independent, this results in

$$E\left[\mathbf{h}^T \check{\mathbf{h}}\right] = \sum_{l=0}^{L-1} \frac{2(N-l)^2 \sigma_{h_l}^4}{2(N-l)\sigma_{h_l}^2 + \sigma_w^2} \tag{13}$$

where the following approximation is utilized

$\ddot{\mathbf{A}}^{-1} = diag\left(\left[\frac{\sigma_w^2 \sigma_{h_0}^2}{2N\sigma_{h_0}^2 + \sigma_w^2} \; \frac{\sigma_w^2 \sigma_{h_1}^2}{2(N-1)\sigma_{h_1}^2 + \sigma_w^2} \cdots \frac{\sigma_w^2 \sigma_{h_{L-1}}^2}{2(N-L+1)\sigma_{h_{L-1}}^2 + \sigma_w^2}\right]^T\right)$, and $\ddot{\mathbf{S}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-\mathbf{1}} \ddot{\mathbf{S}} = \frac{\ddot{\mathbf{S}}^T \ddot{\mathbf{S}}}{\sigma_w^2}$. The other term $E\left[\check{\mathbf{h}}^T \check{\mathbf{h}}\right]$ in (12) can be expressed as

$$\begin{aligned} E\left[\check{\mathbf{h}}^T \check{\mathbf{h}}\right] &= E\left[\left(\ddot{\mathbf{S}}\mathbf{h} + \ddot{\mathbf{w}}\right)^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-1} \ddot{\mathbf{S}} \ddot{\mathbf{A}}^{-2} \ddot{\mathbf{S}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-1} \left(\ddot{\mathbf{S}}\mathbf{h} + \ddot{\mathbf{w}}\right)\right] \\ &= E\left[\mathbf{h}^T \ddot{\mathbf{S}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-1} \ddot{\mathbf{S}} \ddot{\mathbf{A}}^{-2} \ddot{\mathbf{S}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-1} \ddot{\mathbf{S}}\mathbf{h}\right] \\ &\quad + E\left[\ddot{\mathbf{w}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-1} \ddot{\mathbf{S}} \ddot{\mathbf{A}}^{-2} \ddot{\mathbf{S}}^T \mathbf{\Lambda}_{\ddot{\mathbf{r}}}^{-1} \ddot{\mathbf{w}}\right] \end{aligned} \tag{14}$$

which can be approximated as

$$\begin{aligned} E\left[\check{\mathbf{h}}^T \check{\mathbf{h}}\right] = \sum_{l=0}^{L-1} \frac{4(N-l)^2 \sigma_{h_l}^6}{\left(2(N-l)\sigma_{h_l}^2 + \sigma_w^2\right)^2} + \sum_{k=0}^{l} \frac{\sigma_w^2 \sigma_{h_k}^4}{\left(2\sigma_{h_k}^2 (N-k) + \sigma_w^2\right)^2} \\ + (2N-L) \frac{\sigma_w^2 \sigma_{h_l}^4}{\left(2\sigma_{h_l}^2 (N-l) + \sigma_w^2\right)^2} \end{aligned} \tag{15}$$

Finally, $E[\mathbf{h}^T \mathbf{h}] = \sum_{l=0}^{L-1} \sigma_{h_l}^2$.

Another estimator that utilizes only one PLCP preamble can be expressed as follows

$$\check{\mathbf{h}}_{\tilde{\varepsilon}} = \mathbf{A}^{-1}\mathbf{S}^{T}\boldsymbol{\Lambda}_{\mathbf{r}}^{-1}\mathbf{r} \qquad (16)$$

where $\boldsymbol{\Lambda}_{\mathbf{r}}$ is alike to $\boldsymbol{\Lambda}_{\ddot{\mathbf{r}}}$ but with size $N \times N$, and $\mathbf{A}$ is the appropriate matrix. Similarly, the MSE of the estimator in (16) can be expressed as

$$\begin{aligned} MSE = \sum_{l=0}^{L-1} \sigma_{h_l}^2 - 2 \times \frac{(N-l)^2\sigma_{h_l}^4}{\left((N-l)\sigma_{h_l}^2 + \sigma_w^2\right)^2} \\ + \frac{(N-l)^2\sigma_{h_l}^6}{\left((N-l)\sigma_{h_l}^2 + \sigma_w^2\right)^2} + \sum_{k=0}^{l} \frac{\sigma_w^2\sigma_{h_k}^4}{\left(\sigma_{h_k}^2(N-k) + \sigma_w^2\right)^2} \\ + (N-L)\frac{\sigma_w^2\sigma_{h_l}^4}{\left(\sigma_{h_l}^2(N-l) + \sigma_w^2\right)^2} \end{aligned} \qquad (17)$$

The incorporation of S or S̈ in the matrix A renders the estimation insensitive to timing error for high signal to noise ratios (SNR). This fact will deteriorate the estimation in (6) but will be advantageous in channel estimation.

To summarize the proposed algorithm, the receiver calculates $\check{\mathbf{h}}_{\tilde{\varepsilon}}$ from equation (11) or (12) for $N_{min} \leq n \leq N_{max}$, and substitute the estimated vector in equation (6). These steps are repeated until we obtain the maximum from (6). The vector $\check{\mathbf{h}}_{\tilde{\varepsilon}}$ that generates the maximum in equation (6) will be considered the estimated channel coefficients.

If we approximate $\mathbf{S}^{\mathbf{T}}\mathbf{S}$ as $\mathbf{S}^{\mathbf{T}}\mathbf{S} = N\mathbf{I}_{\mathbf{L}}$ our estimator will resemble the estimator in [4]. In our case, we approximate $\mathbf{S}^{\mathbf{T}}\mathbf{S}$ as

$$\mathbf{S}^{\mathbf{T}}\mathbf{S} = \begin{bmatrix} N & 0 & \cdots & 0 \\ 0 & N-1 & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & N-L \end{bmatrix} \qquad (18)$$

and

$$\ddot{\mathbf{S}}^{\mathbf{T}}\ddot{\mathbf{S}} = \begin{bmatrix} 2N & 0 & \cdots & 0 \\ 0 & 2(N-1) & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & 2(N-L) \end{bmatrix} \qquad (19)$$

As shown in **Fig. 2**, the approximation renders the timing metric (log-likelihood function (8)) more sensitive to timing error. **Fig. 2(a)** illustrates the timing metric with no approximation. It is apparent in this figure that there is a flat area which imposes ambiguity for the start of the OFDM symbol. While in **Fig. 2(b)**, we can see there is a sharp maximum value which refers to the start of OFDM symbol. We will see in the simulation results section that this approximation improves the timing estimation in terms of MSE.

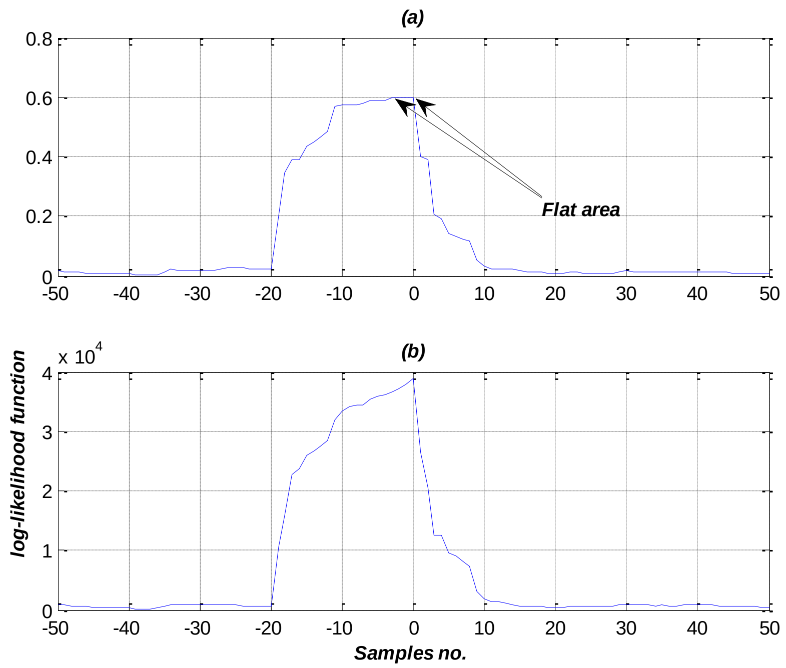

Fig. 2. log-likelihood function in SNR=20dB (a) without approximation. (b) with approximation

4. Simulation Results and Discussion

Our simulation parameters are as in [3] with the extreme UWB dispersive multipath channel models CM1 and CM4 [1]. For the proposed method uncoded bit error rate (BER) performance evaluation, 1000 OFDM symbols of 100 bits (DPSK modulation) are averaged to obtain the BER. While for the MSE evaluation of timing and channel estimation error, 10000 OFDM symbols are averaged to obtain the MSE. The channel is considered constant for the OFDM packet. Independent channel vector is generated each iteration.

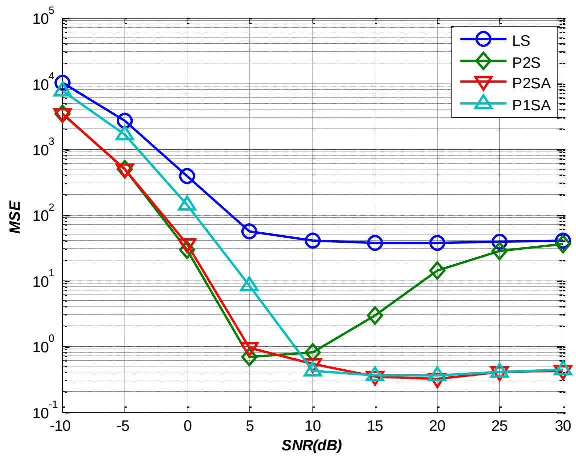

Fig. 3. MSE versus SNR of the timing error for LS method and the proposed methods

Fig. 3 illustrates the MSE of the LS method, proposed timing estimation using two symbols (P2S), proposed timing estimation using one symbol with approximation in (18) (P1SA), and proposed timing estimation using two symbols with approximation (P2SA). We can see that in low SNR's (below $5dB$), P2S performs better. However, due to the ambiguity incurred by the flat area shown in **Fig. 2(a)**, the MSE is high in high SNR. In SNR's above 10dB P1SA almost resembles that one of P2SA.

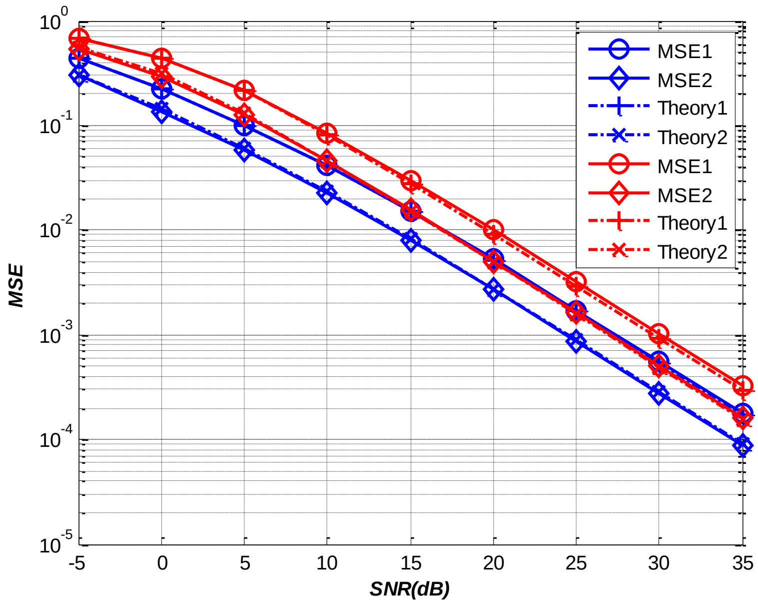

Fig. 4. MSE of channel estimation versus SNR in CM1 (blue) and CM4 (red)

Fig. 4 illustrates the MSE performance of the proposed channel estimator in (11) and (16) in CM1 and CM4. We can see that by averaging over two symbols (MSE2) as in (11), the performance becomes better when using one symbol (MSE1) as in (16). Moreover, we can see that the performance evaluation obtained from analysis (equation (12)) when averaging over two symbols (Theory2) conforms to the one obtained from simulation (MSE2). The same goes for Theory1 obtained from (17) and MSE1. Note that the transmitted signal is assumed of unity power. Therefore, (12) and (17 can be written in terms of SNR by substituting $1/\mathrm{SNR}$ for σ_w^2.

Finally, **Fig. 5** illustrates the BER performance of the combined time and channel estimation in CM1 and CM4. This simulation is done by sending three symbols, 2 symbols are the PLCP preamble [3] separated by two noisy symbols, and data symbol (DPSK modulated) separated by two noisy symbols from the last PLCP symbol. This correspond to TFC number 1[3]. Firstly, time estimation is obtained using MAP channel estimation either from (11) or (16). Then, using this timing and channel estimate, the data symbol is located and the channel is equalized and the BER is evaluated. The BER performance of perfect timing and channel estimation is also shown. Looking at Fig. 5, it is apparent that in SNR's below 10 dB, all the proposed estimators have the same BER, while in SNR's higher than 10 dB P1S and P2S becomes better than P1SA and P2SA.

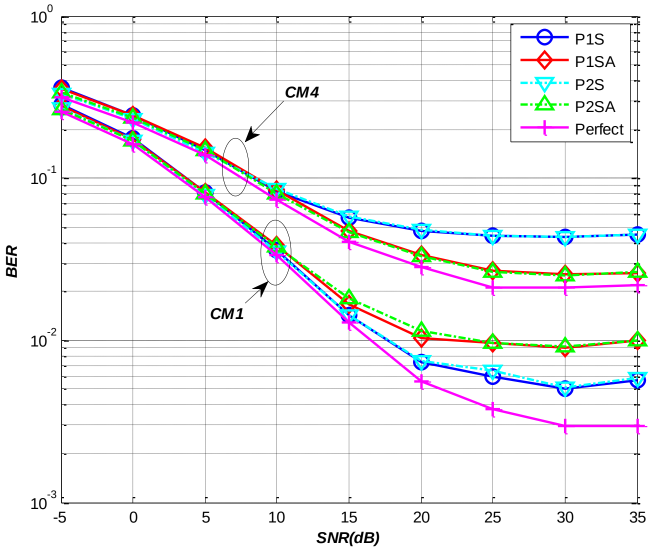

Fig. 5. BER of the proposed methods for channel estimation compared with perfect channel

5. Conclusion

New combined time synchronization and channel estimation method is proposed. Exact timing of the received OFDM symbol is assumed unknown. Therefore, the received signal is considered over one band. Based on the MAP estimated channel impulse response, we obtained timing error estimation. Simulation results showed that the performance of the proposed channel estimation resembles that one of the perfect channel estimation.

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