IMPROVED TIME SYNCHRONIZATION IN PRESENCE OF IMPERFECT CHANNEL STATE INFORMATION

Cong Luong Nguyen[1], *Anissa Mokraoui*[1], *Pierre Duhamel*[2], *Nguyen Linh-Trung*[3]

[1]L2TI, Institut Galilée, Université Paris 13, Sorbonne Paris Cité
99 avenue J.-B Clément, F93430 Villetaneuse, France
[2]LSS/CNRS, SUPELEC, 3 rue Joliot Curie, F91192 Gif sur Yvette, France
[3]VNU University of Engineering and Technology, 144 Xuan Thuy, Cau Giay, Hanoi, Vietnam
{luong.nguyen, anissa.mokraoui}@univ-paris13.fr, pierre.duhamel@lss.supelec.fr, linhtrung@vnu.edu.vn

ABSTRACT

This paper addresses the time synchronization problem in IEEE 802.11a OFDM wireless systems. To enhance the coarse time synchronization mechanism, recent methods exploit not only traditional training sequences as specified by the standard but also additional knowledge available when the Carrier Sense Multiple Access with Collision Avoidance mechanism (CSMA/CA) is triggered. In this case, additional information can be used as training sequences based on the protocol knowledge training sequence known by the receiver. This step is followed by a time synchronization and channel estimation which results in the smallest Channel Estimate Errors (CEE) according to the selected criterion (e.g. LS, MAP). However it was found that the synchronization failure probability heavily depends on the channel estimate quality. Therefore to improve the performance of this class of algorithms, we propose an optimal time synchronization metric that minimizes the average of the transmission error over all CEE. Simulation results show a strongly improved performance in terms of synchronization failure probability in comparison with the existing algorithms.

Index Terms— IEEE 802.11a, OFDM, CSMA/CA, Time Synchronization, Channel Estimation Errors.

1. INTRODUCTION

IEEE 802.11a standard exploits Orthogonal Frequency Division Multiplexing (OFDM) as an effective modulation technique that supports a high-speed data transmission [1]. However, it is very sensitive to the Inter-Symbol Interference (ISI) and Inter-Carrier Interference (ICI). Accurate time and frequency synchronization is therefore required before demodulating a physical packet at the receiver.

A training sequence composed of two symbols is proposed in [2] where the first symbol consists of two identical halves. The Auto-Correlation Function (ACF) based on these two halves is applied on the received signal. The maximum absolute value of this function allows the receiver to estimate the symbol timing. To reduce the training symbols, authors in [3] proposed to use only one symbol that is copied from the first symbol of the data symbols. The time offset estimate is achieved from the ACF between the two repetition symbols.

In [4], a Maximum Likelihood (ML) algorithm relied on the CP of OFDM symbols is exploited for the symbol timing estimation. Instead of working directly with the received complex samples, these samples are quantized into one of four new complex samples. The new samples contain real and imaginary parts of values ± 1. After being quantized, the new complex samples still contain time offset information. Using the quantized samples, the time offset is then estimated by maximizing the ML function.

Based on the IEEE 802.11a preamble structure, time synchronization can be performed in two steps: Coarse Time Synchronization (CTS) followed by Fine Time Synchronization (FTS), as proposed in [5]. In the CTS, based on the Short Training Field (STF) as specified by the standard, the author computes two auto-correlation metrics and afterward compares the difference between them. The symbol timing estimate is deduced from the index at which the difference between both metrics is maximum. To estimate the remaining time offset after the CTS step, the FTS step is based on a joint time synchronization and channel estimate according to the Least Square (LS) criterion. In [7], apart from the STF, the CTS step exploits the SIGNAL field of the frame thanks to the fact that its unknown parts are predictable at the receiver when the Carrier Sense Multiple Access with Collision Avoidance (CSMA/CA) mechanism is triggered and the Request To Send (RTS) and Clear To Send (CTS) control frames are exchanged. Moreover, the FTS step is improved when the joint time synchronization and channel estimate is performed according to the Minimum Mean Square Error (MMSE) criterion. The same approach is also found in [8], the Maximum A Posteriori (MAP) criterion is applied instead of the MMSE one.

The analysis of the proposed solutions in [5], [7] and [8], shows that the channel estimation method strongly affects the time synchronization accuracy. Indeed when the channel estimation error is smaller, the performance of the time synchronization improves. Based on this observation, rather than trying to find the best channel estimate method, we propose an optimal time synchronization metric that minimizes the average of transmission error over all possible Channel Estimation Errors (CEE).

The paper is organized as follows. Section 2 briefly introduces the IEEE 802.11a wireless communication system. Section 3 explains the recent time synchronization algorithms which exploit the SIGNAL field [7], [8] and [10]. Section 4 describes the proposed time synchronization algorithm when channel estimate inaccuracies are taken into consideration. Section 5 presents some simulation results to confirm the advantage of the proposed algorithm. Finally, Section 6 concludes the paper.

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2. WIRELESS COMMUNICATION SYSTEM

The IEEE 802.11a physical packet is composed of three main fields: a PREAMBLE training field, a SIGNAL field and a DATA field (see Fig. 1). The PREAMBLE field contains (i) ten Short Training Field (STF) repetitions usually used for Automation Gain Control (AGC), diversity selection, signal detection and coarse frequency synchronization; and (ii) two Long Training Field (LTF) repetitions exploited for channel estimation and fine frequency synchronization at the receiver. The physical packet modulation is depicted on

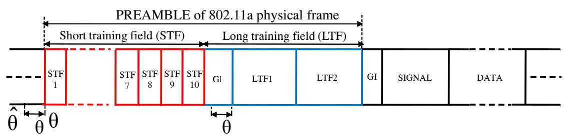

Fig. 1. IEEE 802.11a physical packet

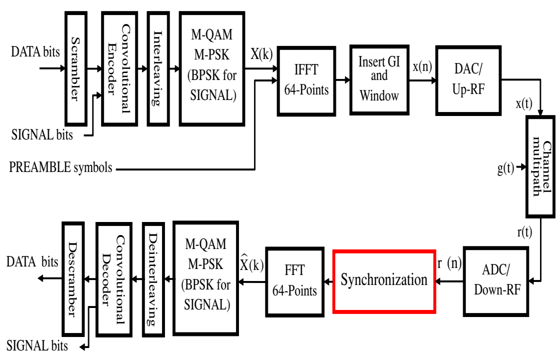

Fig. 2. Wireless communication system using OFDM

Fig. 2. N symbols $X(k)$ (with $0 \leq k \leq N-1$ and N the number of IFFT points) are applied to an Inverse Fast Fourier Transform (IFFT) to get time-domain samples $x(n)$. These samples are then shaped by a windowing function to form an OFDM symbol. Note that each OFDM symbol is preceded by a Cyclic Prefix (CP) to reduce the ISI. Sequence $x(n)$ is transmitted via a multipath fading channel, here considered to be a Finite Impulse Response (FIR) filter of length L.

The received discrete baseband signal $r_\Delta(n)$ is given by

$$r_\Delta(n) = \sum_{i=0}^{L-1} h(i)x(n-i-\theta)e^{j\frac{2\pi\epsilon(n-\theta)}{N}} + g(n), \quad (1)$$

where $h(i)$ denotes the slowly time-varying discrete complex Channel Impulse Response (CIR) with $\sum_{i=0}^{L-1} \mathrm{E}\{|h(i)|^2\} = 1$ (E is the expectation operator). $g(n)$ is the complex additive white Gaussian noise. $\epsilon = \Delta F_c T$ is normalized frequency offset where ΔF_c is the frequency offset between the transmitter and the receiver, T is the OFDM symbol duration and θ is the symbol timing. In order to demodulate the received signal $r_\Delta(n)$, the parameters θ and ϵ must be estimated by the receiver. In this paper, we only consider the time synchronization algorithms to estimate the symbol timing θ.

3. TIME SYNCHRONIZATION USING THE SIGNAL FIELD

This section reviews previous time synchronization algorithms as developed in [7], [8] and [10]. These algorithms are performed in two main steps (see Fig. 3): CTS followed by FTS using joint time synchronization and MAP channel estimation.

3.1. First step: Coarse time synchronization

The coarse time synchronization consists of two steps: (i) Symbol timing estimation; and (ii) remaining time estimation that can be considered as a "fine time synchronization" step in this section.

3.1.1. Symbol timing estimation

The symbol timing of equation (1) is estimated by

$$\hat{\theta} = \arg\max_{\theta} |Z(\theta)|, \quad (2)$$

where $Z(\theta)$ is the Cross-Correlation Function (CCF) given by

$$Z(\theta) = \sum_{n=0}^{L_{\mathrm{STF}}-1} c^*(n) r_\Delta(n+\theta), \quad (3)$$

with $c(n)$ the known sequence corresponding to ten STF repetitions (see Fig.1) and L_{STF} the number of samples in $c(n)$.

The received signal $r_\Delta(n)$ is the transmitted signal $x(n)$ distorted by the characteristics of the channel. To improve the accuracy of the symbol timing estimate, the authors of [10] proposed to replace $r_\Delta(n)$ in (3) by its estimate $\hat{x}(n)$ which should be closer to the transmitted signal $x(n)$. For this, the authors proposed a MAP channel estimation from the reception of RTS control frame when the CSMA/CA mechanism is active. Assume that the channel is slowly time-varying and remains stationary during at least the transmission time between the RTS control frame and the data frame. Denote $H_{MAP}(k)$ the MAP channel estimate in frequency-domain, $\hat{X}(k)$ the transmitted symbol estimate in frequency-domain based on a Zero-Forcing (ZF) equalizer and given as follows:

$$\hat{X}(k) = R_\Delta(k)/\hat{H}_{MAP}(k), \quad (4)$$

where $R_\Delta(k)$ is the received data signal in frequency-domain. The estimate of the transmitted DATA signal in the time-domain is then deduced as follows:

$$\hat{x}(n) = \frac{1}{N}\sum_{i=0}^{N-1} \hat{X}(k) e^{j2\pi k \frac{n}{N}}, \quad (5)$$

which is used by (3) to improve the estimation accuracy of the time offset.

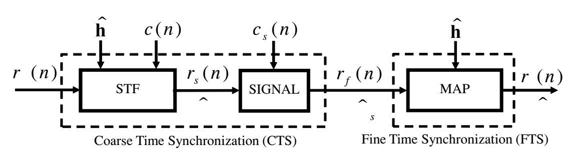

Fig. 3. Previous time synchronization algorithms ([7], [8], [10])

3.1.2. Remaining time offset estimation

The signal after the symbol timing estimation step is given by

$$r_s(n) = \sum_{i=0}^{L-1} h(i)x(n-i-\Delta\theta_s)e^{j\frac{2\pi\epsilon(n-\Delta\theta_s)}{N}} + g(n), \quad (6)$$

where $\Delta\theta_s = \theta - \hat{\theta}$ is the remaining time offset. To estimate $\Delta\theta_s$, in [7] and [8] the IEEE 802.11a SIGNAL field is exploited as an additional reference sequence at the receiver when the CSMA/CA mechanism is triggered. After predicting its unknown parts, the SIGNAL field is thus known at the receiver (this is fully compatible with the existing standard). The remaining time offset is then estimated as follows:

$$\widehat{\Delta\theta_s} = \arg \max_{\Delta\theta_s^{(k)} \in \Theta} |Z(\Delta\theta_s^{(k)})|, \tag{7}$$

where $\Theta = \{\Delta\theta_s^{(k)} | k = -K, \ldots, K\}$ with K a predefined integer value and $Z(\Delta\theta_s^{(k)})$ is the CCF, given by

$$Z(\Delta\theta_s^{(k)}) = \sum_{n=0}^{L_{\text{SIG}}-1} c_s^*(n) r_s(n + \Delta\theta_s^{(k)}), \tag{8}$$

where $c_s(n)$ is the known sequence corresponding to the SIGNAL field and L_{SIG} is the length of the SIGNAL field added to the CP length. Therefore the received signal with the remaining time offset ($\Delta\theta = \Delta\theta_s - \widehat{\Delta\theta_s}$) is given by

$$r_s'(n) = \sum_{i=0}^{L-1} h(i) x(n - i - \Delta\theta) e^{-j\frac{2\pi\epsilon(n-\Delta\theta)}{N}} + g(n). \tag{9}$$

This paper focuses only on time synchronization problem and assumes a perfect knowledge of the frequency offset (i.e. ϵ is known). The frequency offset compensation is thus performed as follows: $r_s'(n) e^{-j\frac{2\pi\epsilon n}{N}}$. The received signal after frequency offset compensation is expressed by

$$r_f(n) = \sum_{i=0}^{L-1} h(i) x(n - i - \Delta\theta) e^{-j\frac{2\pi\epsilon\Delta\theta}{N}} + g(n). \tag{10}$$

3.2. Second step: Joint time synchronization and MAP channel estimation

Equation (10) is rewritten in a matrix form as follows:

$$\mathbf{r}_f = \mathbf{G}\mathbf{h}_{\Delta\theta} + \mathbf{g}, \tag{11}$$

where $\mathbf{r}_f$ of size $N \times 1$ is the received vector corresponding to the LTF, here, N is also the length of one LTF repetition. $\mathbf{G} = \mathbf{F}^H\mathbf{X}\mathbf{F}$ where $\mathbf{F}$ is the $N \times N$ FFT matrix and $\mathbf{X}$ is the $N \times N$ diagonal matrix whose diagonal elements are the known LTF symbols. Notation $(\cdot)^H$ denotes the conjugate transpose operator. $\mathbf{g}$ is the noise vector of size $N \times 1$ and $\mathbf{h}_{\Delta\theta}$ is the CIR vector of size $N \times 1$ following a Gaussian distribution with a mean vector μ_h.

Let $\mathbf{\Lambda} = \{-\Delta\theta_M, .., \Delta\theta_M\}$ be the set of $2M + 1$ possible time offset values. For each $\Delta\theta_m \in \mathbf{\Lambda}$, the time-domain CIR (i.e. $\mathbf{h}_{\Delta\theta_m}$) is estimated according to the MAP criterion expressed as

$$\widehat{\mathbf{h}}_{\Delta\theta_m} = (\mathbf{G}^H\mathbf{G} + \sigma_g^2\mathbf{R_h^{-1}})^{-1}(\mathbf{G}^H\mathbf{r}_f + \sigma_g^2\mathbf{R_h^{-1}}\mu_h) \tag{12}$$

where σ_g^2 is the noise variance. $\mathbf{R_H} = \mathbf{F}\mathbf{R_h}\mathbf{F}^H$ is the frequency-domain correlation matrix of the true channel; here, $\mathbf{R_h} = E\{\mathbf{h}\mathbf{h}^H\}$ can be estimated using the LS approximation as given by

$$\mathbf{R_h} \approx E\{\tilde{\mathbf{h}}_{\Delta\theta_m}\tilde{\mathbf{h}}_{\Delta\theta_m}^H\} = \text{diag}(|\tilde{h}_{\Delta\theta_m}(0)|^2, ..., |\tilde{h}_{\Delta\theta_m}(L-1)|^2), \tag{13}$$

where $\tilde{\mathbf{h}}_{\Delta\theta_m}$ is the LS-based CIR estimate. Similarly, we get $\mu_h \approx E\{\tilde{\mathbf{h}}_{\Delta\theta_m}\}$. Among $2M+1$ estimates of $\widehat{\mathbf{h}}_{\Delta\theta_m}$ obtained from (12), the optimal time offset satisfies the following condition:

$$|\hat{h}_{\Delta\theta_m}(0)| > \beta \max_{\Delta\theta_k \in \mathbf{\Lambda}} |\hat{h}_{\Delta\theta_k}(0)|, \tag{14}$$

where β is a given threshold. Therefore, the set $\mathbf{\Lambda}$ becomes $\mathbf{\Gamma}$

$$\mathbf{\Gamma} = \{\omega_0, \ldots, \omega_{M'};\ \ M' \leq 2M\}. \tag{15}$$

The remaining time offset is finally estimated by

$$\widehat{\Delta\theta} = \arg \max_{\omega_{m'}} \{S(\omega_{m'})\}, \tag{16}$$

where $S(\omega_{m'})$ is the energy associated to the estimated CIR $\hat{\mathbf{h}}_{\omega_{m'}}$ and is given by

$$S(\omega_{m'}) = \sum_{n=0}^{L-1} |\hat{h}_{\omega_{m'}}(n)|^2. \tag{17}$$

4. PROPOSED TIME SYNCHRONIZATION ALGORITHM

This section presents the proposed time synchronization algorithm where changes are made to the fine time synchronization step. The algorithm is summarized in Fig. 4 and is described below.

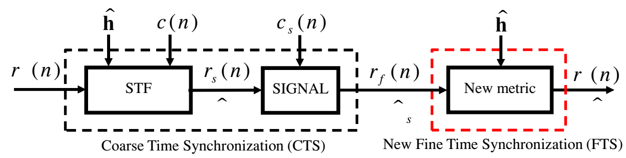

Fig. 4. Proposed time synchronization algorithm

4.1. First step: Coarse time synchronization

The first step is concerned with CTS as described in section 3.1 where equations (2) and (3) are used for the symbol timing estimation. The remaining time offset is then implemented as described in section 3.1.2 where the SIGNAL field is exploited as an additional information.

4.2. Second step: New fine time synchronization

The purpose of this section is to estimate the remaining time offset $\Delta\theta$ in equation (10) which can be expressed in the frequency-domain in matrix form as follows:

$$\mathbf{Y}_f = \mathbf{X}\mathbf{H} + \mathbf{G}, \tag{18}$$

where $\mathbf{Y}_f$ is the received frequency-domain vector of size $N \times 1$ (FFT transform applied to $r_f(n)$). $\mathbf{G}$ is the noise vector of size $N \times 1$. $\mathbf{H}$ of size $N \times 1$ is the FFT transform of $\mathbf{h_{\Delta\theta_m}}$ and thus it contains parameter $\Delta\theta$.

The Probability Density Function (PDF) of $\mathbf{H}$, $\Psi(\mathbf{H})$, is assumed to follow a circular Gaussian distribution with zero mean; that is, $\mathbf{H} \sim C\aleph(0, \mathbf{R_H})$, where $\mathbf{R_H}$ is the covariance matrix of size $N \times N$. This PDF is then given by

$$\Psi(\mathbf{H}) = \frac{1}{\pi^N det(\mathbf{R_H})} \exp{(\mathbf{H}^H\mathbf{R_H}^{-1}\mathbf{H})}. \tag{19}$$

The remaining time offset $\Delta\theta$ to be estimated is considered the one which minimizes the following new metric:

$$\Delta\hat{\theta} = \arg \min_{\Delta\theta \in \mathbf{\Lambda}} \{\widetilde{D}(\Delta\theta)\}, \tag{20}$$

where $\mathbf{\Lambda}$ is given in section 3.2, $\widetilde{D}(\Delta\theta)$ is the average of the transmission error $D(\mathbf{H}) = \|\mathbf{Y}_f - \mathbf{XH}\|^2$ over all channel estimation errors and is given by

$$\widetilde{D}(\Delta\theta) = E_{\mathbf{H}|\hat{\mathbf{H}}}[D(\mathbf{H})] = \int_{\mathbf{H}} D(\mathbf{H})\Psi(\mathbf{H}|\hat{\mathbf{H}})d(\mathbf{H}). \qquad (21)$$

$\Psi(\mathbf{H}|\hat{\mathbf{H}})$ is the PDF of $\mathbf{H}$ (true unknown channel) given $\hat{\mathbf{H}}$; and $\hat{\mathbf{H}}$ is considered in this paper as the LS channel estimate provided by $\hat{\mathbf{H}} = \mathbf{X^{-1}Y_f}$. Note that the form of equation (21) has been inspired from [6]. This equation requires the knowledge of the PDF $\Psi(\mathbf{H}|\hat{\mathbf{H}})$ determined as follows:

$$\Psi(\mathbf{H}|\hat{\mathbf{H}}) = \frac{\Psi(\hat{\mathbf{H}}|\mathbf{H})\Psi(\mathbf{H})}{\Psi(\hat{\mathbf{H}})}, \qquad (22)$$

where $\Psi(\mathbf{H})$ is given by (19). $\Psi(\hat{\mathbf{H}}|\mathbf{H}) \sim C\aleph(\mu_{\hat{\mathbf{H}}|\mathbf{H}}, \mathbf{\Sigma}_{\hat{\mathbf{H}}|\mathbf{H}})$ and $\Psi(\hat{\mathbf{H}}) \sim C\aleph(\mu_{\hat{\mathbf{H}}}, \mathbf{\Sigma}_{\hat{\mathbf{H}}})$. After some mathematical manipulations, we show that $\Psi(\hat{\mathbf{H}}|\mathbf{H}) \sim C\aleph(\mathbf{H}, \mathbf{\Sigma}_\epsilon)$, $\Psi(\hat{\mathbf{H}}) \sim C\aleph(0, \mathbf{R_H} + \mathbf{\Sigma}_\epsilon)$ where $\mathbf{R_H} = E\{\mathbf{HH}^H\}$ and $\mathbf{\Sigma}_\epsilon = E((\mathbf{X^{-1}G})^H(\mathbf{X^{-1}G}))\mathbf{I}$ with $\mathbf{I}$ the $N \times N$ identity matrix. The calculation of $\mathbf{R_H}$ is based on the MAP channel estimate (see equation (4)) as given by $\mathbf{R_H} \approx E\{\hat{\mathbf{H}}_{MAP}\hat{\mathbf{H}}_{MAP}^H\}$.

Substituting $\Psi(\mathbf{H})$, $\Psi(\hat{\mathbf{H}}|\mathbf{H})$ and $\Psi(\hat{\mathbf{H}})$ into (22) yields $\Psi(\mathbf{H}|\hat{\mathbf{H}}) \sim C\aleph(\mu_{\mathbf{H}|\hat{\mathbf{H}}}, \mathbf{\Sigma}_{\mathbf{H}|\hat{\mathbf{H}}})$ with $\mu_{\mathbf{H}|\hat{\mathbf{H}}} = \mathbf{\Sigma}_\Delta\hat{\mathbf{H}}$ and $\mathbf{\Sigma}_{\mathbf{H}|\hat{\mathbf{H}}} = \mathbf{\Sigma}_\Delta\mathbf{\Sigma}_\epsilon$ where $\mathbf{\Sigma}_\Delta = \mathbf{R_H}(\mathbf{R_H} + \mathbf{\Sigma}_\epsilon)^{-1}$. For simplicity, equation (21) is reduced to the following calculation :

$$\widetilde{D}(\Delta\theta) = E_{\mathbf{W}}[D(\mathbf{W})] = \int_{\mathbf{W}} D(\mathbf{W})\Psi(\mathbf{W})d(\mathbf{W}), \qquad (23)$$

where $\mathbf{W} \sim \Psi(\mathbf{W}) = C\aleph(\mu_{\mathbf{H}|\hat{\mathbf{H}}}, \mathbf{\Sigma}_{\mathbf{H}|\hat{\mathbf{H}}})$ and $D(\mathbf{W}) = \|\mathbf{Y}_f - \mathbf{XW}\|^2$. After some mathematical manipulations, we obtain

$$\begin{aligned}\widetilde{D}(\Delta\theta) = & E[\mathbf{Y}_f^H\mathbf{Y}_f] - E[\mathbf{Y}_f^H\mathbf{X}]\mu_{\mathbf{H}|\hat{\mathbf{H}}} - \\ & - \mu_{\mathbf{H}|\hat{\mathbf{H}}}^H E[\mathbf{X}^H\mathbf{Y}_f] + tr(\mathbf{\Sigma}_{\mathbf{H}|\hat{\mathbf{H}}}\mathbf{XX}^H) + \\ & + \mu_{\mathbf{H}|\hat{\mathbf{H}}}^H\mathbf{X}^H\mathbf{X}\mu_{\mathbf{H}|\hat{\mathbf{H}}}, \end{aligned} \qquad (24)$$

where $tr(.)$ indicates the trace operator. The remaining time offset $\Delta\hat{\theta}$ is then deduced from equation (20).

5. SIMULATION RESULTS

This section discusses the performance of the proposed time synchronization algorithm in terms of Probability of Synchronization Failure (PSF). The communication system provided in Fig. 2 has been implemented. The simulation parameters specified by the IEEE 802.11a standard [1] are listed in Table 1 in the presence of COST207-RA channel model. Simulations are performed with a frequency offset of $0.5\Delta F_c$ where the subcarrier spacing $\Delta F_c = 0.3125$ MHz. However, we only consider the time synchronization, the frequency offset is compensated after coarse time synchronization step (i.e. the first step). Fig. 5 compares the PSF of the following time synchronization algorithms:

i) **Algorithm 1 (SIGNAL-MAP)** [8]: This is the algorithm described in section 3. The CTS (the first step) is based on equations (2) and (3) for the symbol timing estimation and then exploits the SIGNAL field as an additional training sequence. The fine time synchronization is then performed according to the joint time synchronization and MAP channel estimation.

ii) **Algorithm 2 (CE-SIGNAL-MAP)** [10]: This algorithm is similar to Algorithm 1 but in the symbol timing estimation (i.e. section 3.1.1), equations (2) and (**??**) are performed.

iii) **Algorithm 3 (CE-SIGNAL-CEE)**: This is the proposed algorithm described in section 4. The PSF of the proposed time synchronization algorithm (shown by solid line) is much lower than those of the previous time synchronization algorithms (shown by dash lines) (see Fig. 5). Indeed for a given SNR=15 dB, the PSF of algorithms 1, 2 and 3 is respectively: PSF(MAP-SIGNAL)= 3.4×10^{-2}, PSF(CE-MAP-SIGNAL)=4.4×10^{-5} and PSF(CE-SIGNAL-CEE)= 5×10^{-6}. This confirms that the performance of the proposed method improves significantly.

Table 1. Simulation parameters

Parameters	*Values*
Bandwidth (B)	20 MHz
Sampling time (T_s)	50 ns
Number of subcarriers (N_c)	52
Number of points FFT/IFFT	64
Subcarrier spacing (ΔF)	0.3125 MHz
Channel model	Rice with COST207-RA
Channel time delay	(0, 200, 400, 600) ns
Power of channel paths (P_c)	(0, -2, -10, -20) dB
Data rate	6 Mbps
Threshold (β)	0.7
L_{STF}	160
L_{SIG}	80
M	30
K	80

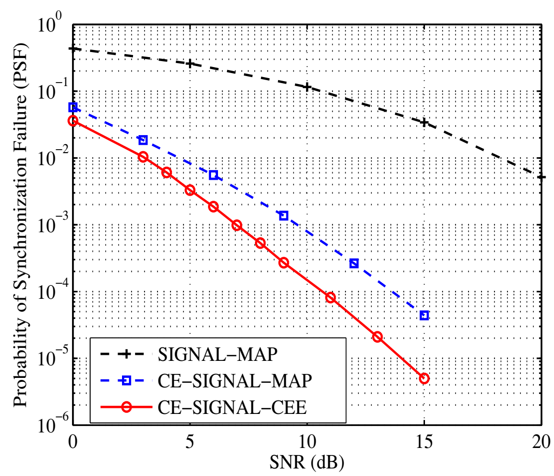

Fig. 5. Probability of synchronization failure performance (with a frequency offset = $0.5\Delta F_c$)

6. CONCLUSION

This paper proposed a time synchronization algorithm for IEEE 802.11a standard wireless communication under imperfect channel state information. A new metric that minimizes the average of transmission error over all CEE has been investigated. Simulation results show that the proposed solution has better performance in terms of probability of synchronization failure as compared to previous time synchronization methods.

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