Timing Synchronization for 802.11a WLANs under Multipath Channels

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Abstract – In the Wireless Local Area Network (WLAN) standard adopted by the IEEE 802.11a standardization group, each data packet starts with a preamble, which consists of ten short training symbols followed by two long training symbols. In this paper, we present an efficient and straightforward coarse timing synchronization scheme based on the short training symbols. Simulation results illustrate that the performance of the proposed scheme is comparable and even superior to that of the conventional timing synchronization method under multipath fading channels. We further show that the combination of the proposed auto-correlation coarse timing synchronization scheme and a frequency domain path delay moving average method leads to a desirable result of symbol timing estimation.

I. Introduction

The Orthogonal Frequency Division Multiplex (OFDM) is an effective modulation technique used in high bit-rate wireless communications. It has been adopted by the wireless LAN standard IEEE 802.11a and high performance LAN type 2 (HIPERLAN/2). However, OFDM systems are extremely sensitive to receiver synchronization imperfections [1], which can cause the degradation of system performance.

Several approaches [2-4] have been proposed on the basis of using preamble symbols or using cyclic prefix and pilot subcarriers. For burst mode transmission such as wireless LAN, the method of using preamble symbols [3] is preferred for fast time and frequency synchronization. When the preambles are repeated training symbols, the timing metric [3, 4] is computed through auto-correlation, which correlates the received samples and their delayed copies (delay and correlation). When the preambles are known to the receiver, the timing metric [5] is obtained by cross-correlation, which correlates the received samples with the locally generated samples. The correlation peak of the timing metric declares a symbol time.

The IEEE 802.11a standard specifies a preamble at the start of every frame. The preamble consists of ten short training symbols and two long training symbols, as explained in [6]. The timing synchronisation process is normally split into an initial coarse synchronization phase and a later fine synchronization phase. The conventional symbol timing synchronization schemes in IEEE 802.11a WLAN systems use short training symbols to estimate a coarse symbol time via auto-correlation and then use long symbols to find a fine symbol time via cross-correlation. Those schemes can achieve fast synchronization between the transmitter and the receiver. The auto-correlation approach presented in [3] is widely used to find a coarse symbol time. However, if the repeated training sequence has a cyclic prefix or the number of repeated symbols is more than two, the timing metric will have a plateau, which causes large variance of the timing estimate. In general, correlation based methods do not find the optimum timing position in the presence of multipath fading channels. This is particularly true for the case where the first arriving replica of the signal (corresponding to the first path) is not the strongest [7-9].

In this paper, we present a double auto-correlation coarse timing synchronization scheme based on the IEEE 802.11a preamble structure. The scheme calculates two auto-correlation based timing metrics and compares the difference between them so as to determine the symbol time. The new approach is more efficient and straightforward than the conventional timing synchronization schemes in that firstly, cross-correlation fine timing synchronization used in the conventional schemes is not needed and secondly, performance of our proposed scheme is comparable and even superior to that of conventional fine timing synchronization under multipath fading channels. To further improve the accuracy of the symbol time acquired from the proposed approach, a frequency domain fine timing synchronization method is applied. The paper is organized as follows. Section II briefly describes the 802.11a baseband signal model and preamble structure. Section III presents the timing synchronization algorithms. Section IV evaluates the proposed algorithm by means of computer simulation. The paper is concluded in section V.

II. An IEEE 802.11A signal model

In IEEE 802.11a, OFDM modulation method is used for data transmission. Consider an OFDM system employing N subcarriers for the transmission of parallel data streams of width N_u, where $N - N_u$ subcarriers (virtual carriers)

at the perimeter of the spectrum are regarded as the guard band.

At the transmitter, the data stream is mapped into N complex symbols in the frequency domain, including null data symbols for virtual subcarriers. These N complex symbols are modulated on to the N subcarriers by using Inverse Fast Fourier Transform (IFFT) to get a time domain complex OFDM symbol, which is represented as

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

where $X(k)$ denotes the data symbol in subcarrier k, $x(n)$ is the nth sample of the OFDM symbol. The last N_g samples of the IFFT outputs are copied and added to form the cyclic prefix at the beginning of each OFDM symbol. In IEEE802.11a, N is equal to 64, N_u is equal to 52 (including 4 pilot subcarriers) and N_g is equal to 16. The time domain symbols are interpolated, D/A converted, mixed with a carrier and transmitted.

At the receiver, the received signals are down-converted, filtered, A/D converted and decimated to reconstruct the baseband signals. The received signals transmitted through multipath channels are expressed by

$$r(n) = \sum_{i=0}^{N_h-1} x(n-\theta-i) h(i) e^{\frac{j2\pi\varepsilon n}{N}} + w(n) \qquad (2)$$

where $h(i)$ is the sampled complex channel impulse response (CIR), N_h is the length of channel impulse response, θ is the time offset, ε is the carrier frequency offset and $w(n)$ is complex white Gaussian noise. With the CIR confined to the cyclic prefix length, after removing the cyclic prefix the received signals are demodulated via FFT, the demodulated signal in subcarrier k is given by

$$Y(k) = X(k) H(k) e^{-\frac{j2\pi k\theta}{N}} \qquad (3)$$

where $H(k)$ is the channel transfer function, $e^{-\frac{j2\pi k\theta}{N}}$ is the phase rotation introduced by time offset θ. If the time offset is not in the range of cyclic prefix, it will cause inter-symbol interference and inter-channel interference. The effect of carrier frequency offset is not discussed here. As coherent modulation is used for transmission, the channel impact has to be estimated and compensated.

In IEEE 802.11a, the training sequence is transmitted at the beginning of each frame to help the receiver accomplish synchronization and channel estimation. The training sequence consists of ten short training symbols, each of which is 0.8 μs, followed by two long training symbols, each of which is 3.2 μs plus a 1.6-μs prefix which precedes the long training symbol. The short training symbols are responsible for signal and packet detection, Automatic Gain Control (AGC) level setting, coarse timing synchronization and coarse carrier frequency offset correction. The long training symbols are used for fine carrier frequency offset and channel estimation. The data payload is composed of a variable number of OFDM symbols. Figure 1 illustrates the frame structure specified in IEEE802.11a.

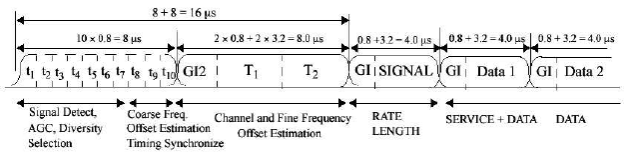

Figure 1 The structure of an OFDM packet[6]

III. TIMING SYNCHRONIZATION ALGORITHMS

A. Coarse timing synchronization

With the repeated short training symbols specified in IEEE802.11a, we can readily use the auto-correlation method introduced in [3] to do initial coarse timing synchronization. In order to overcome the uncertainty in the timing metric plateau and increase accuracy, we calculate two normalized auto-correlation timing metrics. The first metric $M_1(\theta)$ is the normalized correlation between the received signal and itself with a delay of one short symbol N_s, where $N_s = 16$. It creates a plateau of the length of nine short symbols. The second metric $M_2(\theta)$ is the normalized correlation between the received signal and itself with a delay of two short symbols. It creates a plateau of the length of eight short symbols.

$$M_1(\theta) = \frac{\sum_{m=0}^{N_s-1} r(\theta+m) \times r^*(\theta+m+N_s)}{\sum_{m=0}^{N_s-1} \left| r(\theta+m) \right|^2} \qquad (4)$$

$$M_2(\theta) = \frac{\sum_{m=0}^{N_s-1} r(\theta+m) \times r^*(\theta+m+2N_s)}{\sum_{m=0}^{N_s-1} \left| r(\theta+m) \right|^2} \qquad (5)$$

By subtracting $M_2(\theta)$ from $M_1(\theta)$, we obtain a triangular shaped timing metric, as shown in Figure 2. By searching the maximum value of the difference $M_1(\theta) - M_2(\theta)$, we can detect a peak that indicates the

start of the 9th short symbol. Hence, the coarse timing estimate is achieved, which is

$$\hat{\theta} = \arg\max_{\theta}(M_1(\theta) - M_2(\theta)) \tag{6}$$

The auto-correlation in equations (4) and (5) can be calculated iteratively. As shown in Section IV, the output from the estimator (6) is highly localised. The performance of this estimator can be further improved by, for example, averaging over the time stamps of more short symbols acquired by calculating more normalized auto-correlation metrics with different time delay.

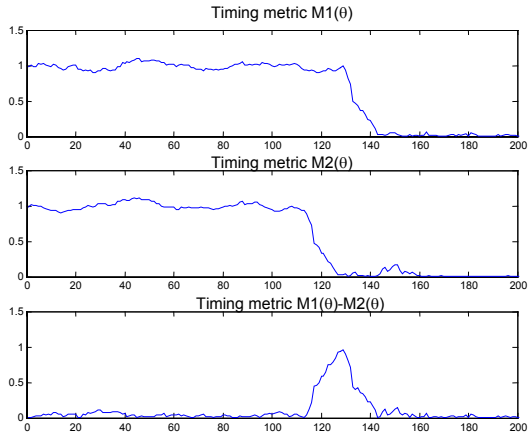

Figure 2 Timing metric of double auto-correlation

The symbol timing estimate $\hat{\theta}$ could be earlier or later than the true time. If $\hat{\theta}$ is earlier than the true time, part of the cyclic prefix of the current symbol is taken as data, thus causing no interference. If $\hat{\theta}$ is later than the true time, part of the cyclic prefix of next symbol is taken as data, which introduces inter-symbol interference (ISI). Therefore, the synchronization process is proceeded with fine symbol time estimation to further improve accuracy.

B. Fine timing synchronization

After performing the coarse timing synchronization scheme described in the previous sub-section, the estimated symbol time is moved earlier by a few samples (eg. 4 samples) to avoid ISI. The conventional method applied to fine timing synchronization is to cross-correlate the received sample with the known long training symbols. As the performance of the auto-correlation approach presented in sub-section A is comparable or even superior (in case of high SNR) to that of the cross-correlation method under multipath channels, as shown in Section IV, the conventional cross-correlation fine timing estimator is not needed under the proposed scheme.

In order to estimate the path delay, we adopt a method that has been studied for continuous transmission [7, 9], for example digital video broadcasting. The received samples are first transformed to frequency domain by FFT, and then the known long training symbols are used to estimate the channel frequency response. The Least Square (LS) estimation of the channel frequency response using long training symbols is written as

$$H(k) = \begin{cases} \dfrac{Y(k)}{X_{LP}(k)}, & k \in N_u \\ 0, & k \notin N_u \end{cases} \tag{7}$$

where $H(k)$ is the channel frequency response at the subcarrier k and $X_{LP}(k)$ is the kth sample of the long training symbol. The channel impulse response can be derived via IFFT, which is

$$h(i) = \sum_{k=0}^{N-1} H(k) e^{\frac{j2\pi ki}{N}}, \qquad i = 0,1,\ldots,N-1 \tag{8}$$

The first path of the channel can be found from channel impulse response by comparing it with a predefined threshold Γ [7, 9]. Hence, the path delay is

$$\hat{\theta}_e = \arg\min_i \{h(i) > \Gamma\} \tag{9}$$

However, it is generally difficult to analytically choose the threshold Γ because it can be affected by channel condition, SNR and IFFT leakage.

An alternative approach to finding the path delay is to employ optimum timing synchronization, discussed in [1] in the presence of multipath channels. The optimum time is defined as the starting time of a window, with a width equal to cyclic prefix, which contains the maximum power of the estimated channel impulse response. Resorting to this idea, by obtaining the moving average of the energy of consecutive impulse response over a window, the optimum time is regarded as the starting time of the window that contains the maximum energy.

$$\hat{\theta}_e = \arg\max_i \{\sum_{j=0}^{N_w-1} |h(i+j)|^2\} \tag{10}$$

where N_w is the size of the window. If the channel length is shorter than that of the cyclic prefix, the selection of window size is dependent on channel length. If the channel length is larger than that of the cyclic prefix, the window size is chosen to equal the length of cyclic prefix, as shown in [1]. Clearly, a majority of replica samples from the previous symbol that cause ISI can be removed by deleting the cyclic prefix. Therefore, the combination of the proposed auto-correlation coarse timing synchronization scheme and the above-mentioned moving average method leads to a desirable result in finding the symbol time, which is further shown in the next section.

IV. SIMULATION AND DISCUSSION

The performance of the proposed algorithm is evaluated by computer simulation. A selection of the parameters of the OFDM system according to IEEE 802.11a is listed in Table 1. The tapped-delay-line channel model is used in the simulation. The power delay profile is HIPERLAN/2

non-line-of-sight (NLOS) channel A as presented in [10], which has 390ns maximum excess delay spread. The normalized Doppler frequency is $f_d T_s = 0.04$.

Sampling rate, F_s	40MHz
Up sampling rate, L_s	2
Channel bandwidth	16.25MHz
Number of FFT points, N	64
FFT symbol period, T_{FFT}	3.2us
Cyclic prefix duration, T_{GI}	0.8us
Number of data subcarriers, N_{SD}	48
Number of pilot subcarriers, N_{SP}	4
Subcarrier spacing, Δf	312.5KHz
Maximum excess delay spread, τ_x	390ns

Table 1 Parameters for the OFDM system under simulation

To demonstrate the features of the proposed scheme as opposed to the conventional timing synchronization scheme [5], using matlab simulation, the probability distribution function (PDF) of the symbol timing offset estimates under both schemes is shown in Figure 3.

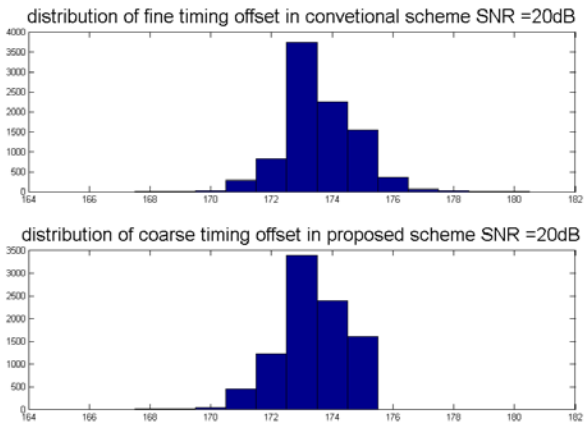

Figure 3 PDF of symbol timing offset in multipath channels

The top plot shows the PDF of cross-correlation fine timing estimator under the conventional scheme, while the bottom plot illustrates the PDF of the proposed auto-correlation coarse estimator (6). The results in this figure were obtained using 10000 simulation runs with SNR = 20dB. The x-axis represents the estimated timing offset $\hat{\theta}$. From these histograms, it is observed that the probability of correct timing offset estimation between the proposed algorithm and the conventional method is very close in multipath fading channels. This observation coincides with the study of cross-correlation synchronization method in multipath fading channels [8] and justifies that the conventional fine timing synchronization phase is not needed under the proposed scheme.

Furthermore, Figure 4 shows the variance of symbol timing offset estimate obtained for different methods. Case I is the variance of timing offset estimate using the cross-correlation method under the conventional scheme [5]. Case II is the variance of timing offset estimate using the proposed auto-correlation approach. Case III is the variance of timing offset estimate based on the method of finding the path delay with first path search (9). Since the predefined threshold is in general difficult to determine analytically, it is chosen through simulation. Case IV is the variance of timing offset estimate using the path delay method with a moving window (10). The size of the window depends on the length of channel impulse response. In this case, we choose 4 in simulation. In a general situation, the maximum window size can be set to the length of cyclic prefix [1].

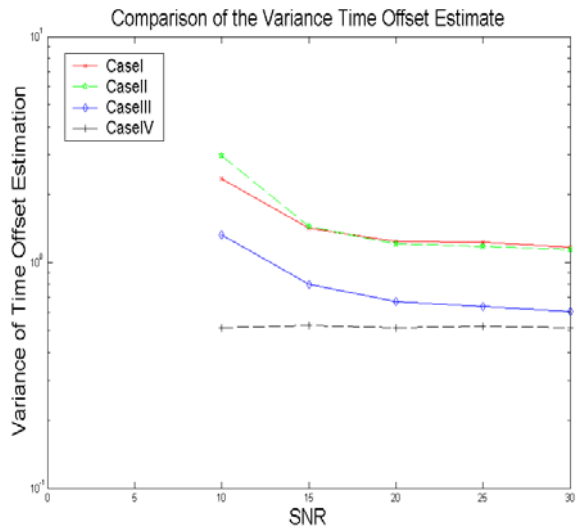

Figure 4 Variance of symbol timing offset estimate

It can be seen from Figure 4 that Case I and Case II have a variance floor and show an equivalent performance, in particular, Case II outperforms Case I when SNR is high. The occurrence of the variance floor is due to the fact that both correlation based algorithms seek to synchronize to the strongest path but not the first path, whereas in a multipath fading environment, the amplitude of each path is variable. As a result, for a high SNR, the channel condition determines the variance of timing offset estimate. For the conventional cross-correlation method, only received samples are affected by the channel condition, whereas in the proposed auto-correlation algorithm, both received samples and their delayed copies are altered by the same amount of the channel effect. When SNR is high, the result of auto-correlation tends to be better than that of cross-correlation.

Regarding Case III, the path delay estimator (9) tries to synchronize with the first path, that is why it has a smaller estimation variance than correlation based methods. However, its performance can be affected by the threshold selection and frequency leakage. As to Case IV, the path delay estimator (10) tends to synchronize to a region that contains the maximum energy of channel

impulse response, which results in a minimum variance of timing offset estimate and shows the robustness of the moving average method. This further demonstrates that using the proposed auto-correlation coarse timing synchronization algorithm followed by the moving average method gives rise to good performance in finding the symbol time.

V. CONCLUSION

In the paper, we have presented a double auto-correlation coarse timing synchronization scheme using short training symbols. The scheme is efficient and straightforward compared to the conventional timing synchronization scheme in that the cross-correlation fine timing synchronization phase in the conventional scheme is not needed under the proposed scheme. Besides, the performance of the new approach is comparable or even superior to that of the conventional scheme under multipath fading channels. It is further demonstrated that combining the proposed auto-correlation coarse timing synchronization algorithm with the path delay moving average method leads to a desirable result of symbol timing estimation.

VI. REFERENCES

[1] R. V. Nee and R. Prasad, *OFDM for Wireless Multimedia Communications*: Artech House, 2000.

[2] J.-J. v. d. Beek, M. Sandell, and P. O. Börjesson, "ML estimation of time and frequency offsets in OFDM systems," *IEEE Trans. Signal Processing*, vol. 45, pp. 1800-1805, 1997.

[3] T. M. Schmidl and D. C. Cox, "Robust frequency and timing synchronization for OFDM," *IEEE Trans. Commun.*, vol. 45, pp. 1613-1621, 1997.

[4] Minn, M. Zeng, and V. K. Bhargava, "On Timing Offset Estimation for OFDM Systems," *IEEE Commun. Letters*, vol. 4, pp. 242-244, 2000.

[5] I. Greenberg, "Retraining WLAN Receivers for OFDM Operation," *Communication Systems Design*, vol. 8, 2002.

[6] IEEE, "IEEE802.11a-1999 part 11: Wireless LAN Medium Access Control (MAC) and Physical Layer(PHY) specifications," IEEE 1999.

[7] N. Chen, M. Tanaka, and R. Heaton, "OFDM Timing Synchronization Under Multi-path Channels," presented at IEEE VTC Spring 2003, Soul Korea, 2003.

[8] K. W. Yip, T. S. Ng, and Y. C. Wu, "Impacts of Multipath Fading on the Timing Synchronization of IEEE802.11a Wireless LANs," presented at IEEE ICC'02, 2002.

[9] J.-J. Kim, Y.-J. Ryu, H.-S. Oh, and D.-S. Han, "Frame selection algorithm with adaptive FFT input for OFDM systems," presented at IEEE ICC'02, 2002.

[10] J. Medbo and P. Schramm, "Channel Modles for HIPERLAN/2 in Different Indoor Scenarios," ETSI BRAN 3ER1085B, 1998.