Orthogonal frequency division multiplexing (OFDM) is the most preferred access technique in Wireless LAN applications that have low implementation cost besides acting as an antidote to highly dispersive and time varying transmission media. In this paper, we focus on the front-end algorithms necessary for the synchronization of the OFDM receiver. The paper details the effects of relevant non-ideal transmission conditions and the design of the synchronization algorithms for the OFDM receivers. The algorithms include frame synchronization; carrier and clock frequency offset estimation and correction. The performances of the algorithms are analyzed and a qualitative estimate of the resulting complexity is given.

1 INTRODUCTION

Wireless local area networks (WLAN's) are gaining popularity since they provide the users with the connectivity of a wired LAN, coupled with the freedom of a wireless link. With WLAN's being developed with data rates comparable to those of wired LAN's, more interests are devoted in developing these systems. The IEEE 802.11 and the HYPERLAN working group have developed WLAN standards so that inexpensive, interoperable equipment can be designed by different manufacturers and used to build different wireless infrastructures [1]. The IEEE 802.11a and the HYPERLAN2 are the two high speed WLAN standards operating in the 5GHz ISM band with OFDM as the modulation scheme and delivering data rates ranging from 6Mbps to 54Mbps. The IEEE working group recently proposed the integration of 802.11a and 802.11b standards in the 2.4 GHz band in the form of 802.11g standard.

It is observed in [2] that OFDM systems offer possibilities for alleviating many of the problems encountered in the wireless environment. It has the advantage of randomizing the burst error caused by the Raleigh fading, so that instead of several adjacent symbols being completely destroyed; many symbols are only slightly distorted. It also leads to significant reduction of ISI. The sub-carrier tones are made orthogonal as in Fig. 1, for efficient bandwidth utilization. In such a system, the coarse equalization is simpler than any other modulation scheme. A simplified schematic of an OFDM transmitter and receiver system is as shown in Fig.2. Inclusion of the cyclic prefix (appending the last G of the N samples corresponding to an OFDM symbol to the beginning of that symbol) helps in combating the effect of multi-path fading while achieving high data-rates.

OFDM SIGNAL MODEL

We consider an OFDM system using an IFFT of size N for modulation. Each OFDM symbol is composed of K<N data symbols $A_{l,k}$, where $l$ denotes the OFDM symbol time index and $k$ denotes the sub-carrier frequency index. The transmitted complex base-band signal can then be described at time $t$ by,

$$t_t(t) = \left(\frac{1}{\sqrt{T_u}}\right) \sum_{l=-\infty}^{\infty} \sum_{k=-K/2}^{K/2-1} A_{l,k} \exp(j2\pi k(t-T_g-lT_s)/T_u) u(t - l.T_s).$$

(1)

where,

$$u(t) = \begin{cases} 1, & 0 \leq t < T_u \\ 0, & \text{otherwise} \end{cases}$$

(2)

Each data symbol is shaped by a rectangular pulse of length $T_u$ and modulated onto a sub-carrier with base-band frequency, $k/T_u$ To avoid inter symbol interference (ISI), the OFDM symbol is preceded by a guard interval of length $T_g$. The resulting symbols are of length

$$T_s = T_u + T_g$$

i.e.

$$N_s = N + Ng$$

(3)

The received sampled signal after passing through a channel with channel impulse response, $h(t,T)$ along with the noise $\eta$ can be described as,

$$x(t_n) = \sum_i h(nT_i). t_s(nT - \tau_i) + \eta(nT)$$

(4)

After removing the guard interval for further receiver processing, the $i$th OFDM symbol is represented by,

$$x_{l,n} = x((n + N_g + 1.N_s).T), \forall n \in (0, N-1).$$

(5)

Demodulation of the sub-carriers via FFT yields the received data symbols as,
\[ X_{i,k} = \sum x_{i,n} \cdot \exp (-j \cdot 2\pi \cdot n \cdot k / N). \]  

(6)

2. PARAMETERS OF THE 802.11 a/g STANDARD

The different 802.11 a/g modes and parameters important to the receiver are : B = 20 MHz ; \( \Delta f \) (as in Fig. 1) = 0.3125 MHz ; \( \tau_{\text{FFT}} \) = 3.2 \( \mu s \) ; \( \tau_{\text{PREAMBLE}} \) = 16 \( \mu s \) ; \( \tau_p \) = 0.8 \( \mu s \) ; \( N \) = 64 (48 data carriers + 4 Pilots ). The sub-carrier indices 0, (27,31) and (-32,-27) do not carry any information ; \( N_g = 16 \). The standard uses Punctured Convolutional Code with code rates \( \frac{1}{2} \) for data rates 6, 12 and 24 Mbps, \( \frac{2}{3} \) for data rate 48 Mbps and \( \frac{3}{4} \) for data rates 9, 18, 36, 54 Mbps; The scheme uses BPSK for 6 and 9 Mbps, QPSK for 12 and 18 Mbps, 16-QAM for 24 and 36 Mbps and 64-QAM for 48 and 54 Mbps. for modulation; The standard [1] further defines interleaving across sub-carriers, the data frame structure, the parity and pad bits and also the dedicated synchronization pilot symbols \( P_{i,k} \) embedded into the OFDM data stream at the sub carriers of indices 7,21,-7,-21 as well as the preamble constituents ,viz. Short Training Sequences and the Long Training sequences .

3. PROPOSED ALGORITHMS FOR RECEIVED SIGNAL

Several papers dealt with different algorithms for combating the undesired effects in an OFDM system. Instead of presenting a detailed analysis of isolated algorithms, this paper focuses on the complete system design process considering the IEEE 802.11 a/g standard. Here we propose a novel method for symbol timing and efficient algorithms for frequency and phase estimation and correction, as described below.

A. Packet detection

Let \( x(i) \) be the current sample under consideration. We define the estimate of packet detection \( Re[e_i] \) where,

\[
e_i = \frac{\sum x^*(i-n) x(i-N-n)}{\sum x^*(i-n) x(i-n)} \quad (n=0) \quad (7)
\]

\( N \), being the period of the 10 short training sequences (STS) each of duration 16 samples. In the ideal situation \( Re[e_i] \) is equal to 1. We fixed a packet detection threshold based on simulation results using HyperLan channel models [3]. If \( Re[e_i] \) is greater than the threshold then we assume that a packet is detected.

B. Frame Synchronization

Since the mechanism used for packet detection is not able to provide the required accuracy needed, a different effect must be exploited to further refine the initial estimate. We define a new estimate \( e'_i \) as,

\[
e'_i = \sum_{n=0}^{N-1} \sum_{k=0}^{M-1} x^*(i-k.n) STS(N-n) \quad (8)
\]

where \( M \) is the total number of cross-correlators. Using the peaks of \( e'_i \), we synchronize the packet. We define \( i_1 \) and \( i_2 \) as the 1st and 2nd peak locations of \( e'_i \) respectively. Therefore,

\[
i_1 = \arg \max \left| e'_{i-k} \right| \quad \forall k \in (4,35) \quad (9)
\]

\[
i_2 = \arg \max \left| e'_{i-k} \right| \quad \forall k \in (20,35). \quad (9)
\]

A correct packet is detected if \( 14 < (i_1-i_2) < 18 \) and \( e'_{i_1} > e'_{i_2} \). The corresponding Packet thus starts at \( (i_2 - 32) \). The details of the frame synchronization with \( M = 3 \) is explained in Fig (3) below.

![Frame Synchronization for an 802.11 a/g packet](image)

Peaks are searched within this window.

C. Carrier and clock frequency-offset estimation and correction

This is a pre-FFT algorithm based on data-aided maximum-likelihood estimator that operates on the received time-domain signal. The training information required is at least two consecutive repeated symbols. For a complex base-band transmitted signal \( s_n \), the pass-band signal \( y_n \) is,

\[
y_n = s_n \cdot \exp (j \cdot 2\pi \cdot f_n \cdot n \cdot T_s) \quad (10)
\]

where \( f_n \) is the transmitter carrier frequency. After the receiver downconverts the signal with a carrier frequency \( f_n \) the received complex base-band signal \( r_n \) is,
\[ r_n = s_n \exp(j.2\pi f_\delta \cdot n.T_s) \exp(-j.2\pi f_\delta \cdot n.T_s) \]
\[ = s_n \exp(j.2\pi (f_{tx} - f_{rx}) \cdot n.T_s) \]
\[ = s_n \exp(j.2\pi f_\delta \cdot n.T_s) \]
where \[ f_\delta \] is the local oscillator frequency mismatch between the transmitter and the receiver carrier frequencies. Thus the received sample can be expressed as,
\[ x_{l,n} = \exp(j.2\pi.k.\pi.N_s.n/T) \cdot (\sum_k \sum a_{l,k} \cdot \Psi(n'.T' - \tau_l) + \eta_{l,n}) \]
\[ \Psi(n'.T' - \tau_l) = \exp(j.2\pi.N_s.n/T) \cdot \eta_{l,n} \]
\[ \forall n \in (0, L-1) \]
\[ \sum_k P_{l,k} \cdot H_k + ICI + \eta_{l,k} \]
\[ \forall i \in (-K/2, K/2 - 1) \& i \neq k. \]

D. Residual phase estimation and correction.
This is a post-FFT data-aided approach based on the frequency-domain data. The 4 pilot sub-carriers that are embedded into an OFDM symbol at sub-carrier index \( k \in \{ -21, -7, +7, +21 \} \) can be used to estimate the residual phase offset.

E. Sampling clock offset estimation and compensation

The oscillators used to generate the Digital to Analog Converter (DAC) and the Analog to Digital Converter (ADC) sampling instants at the transmitter and the receiver often will not have exactly the same period. Thus the sampling instants slowly shift relative to each other. A slow shift of the symbol timing point rotates the sub-carriers along with performance degradation due to inter-carrier interference (ICI) generated by the slightly incorrect sampling instants. It causes loss of the orthogonality of the sub-carriers. Another important effect is either re-sampling of the same data or missing of one sample completely. The normalized sampling error is defined as
\[ t_s = (T' - T)/T \]
where, \( T \) and \( T' \) are the transmitter and receiver sampling periods respectively. Then the overall effect after FFT, on the received sub-carriers \( X_{l,k} \) in the absence of ISI (correct timing) can be shown to be as in \[ 5 \],
\[ X_{l,k} = \exp(j.2\pi.k.t_s.(l.N_s + N_p)/N) \cdot \text{sinc}(\pi.k.t_s).A_{l,k}.H_k + ICI + \eta_{l,k} \]
where,
\[ ICI_k = \sum_i \exp(j.2\pi.i.t_s.(l.N_s + N_p)/N) \cdot \text{sinc}(\pi.(i.t_s + i.k)/N) \cdot A_{l,i}.H_i \]
\[ \forall i \in (-K/2, K/2 - 1) \& i \neq k. \]

WLAN systems typically have relatively small number of sub-carriers and quite small \( t_s \), and hence \( k. t_s \ll 1 \), so sinc (\( \pi \cdot k \cdot t_s \)) \approx 1 and can therefore be neglected. Likewise \( \exp(j.2\pi.k.t_s.(l.N_s + N_p)/N) \) being time-invariant cannot be distinguished from \( H_k \). Thus the most significant problem is caused by the term \( \exp(j.2\pi. k.t_s.(l.N_s + N_p)/N) \) and the ICI which cannot be avoided in practice. Sampling clock-offset estimation is generally done using the pilots. The pilot sub-carriers are divided into two sets; C1 corresponds to the pilots on the negative sub-carriers \( k \in \{-21, -7\} \), and C2 correspond to pilots on the positive sub-carriers \( k \in \{7, 21\} \). The sampling frequency offset is estimated using the knowledge of the linear relationship between the phase rotation caused by the offset and the pilot sub-carrier index. The received pilot sub-carriers, in a simplified form is,
\[ P_{l,k} = H(k).P_{l,k}.\exp(j.2\pi.k.t_s.(l.N_s + N_p)/N) \]
where \( P_{l,k} \) is a pilot of the \( l \)th symbol. For evenly distributed pilot carriers, the estimate of \( t_s \) is given by,
\begin{equation}
t'_{\delta} = (1/2\pi) . ( (L.N_s+N_g)/N ) \left( \varphi_{2,n} - \varphi_{1,n} \right) / (\min_{k\in C2} + \max_{k\in C2}). \tag{26}
\end{equation}

where,
\begin{align*}
\varphi_{1,1} &= \arg \left[ \sum_{k\in C1} P_{1,k} . P'_{1,k} * \right] \forall k \in C1 \\
\varphi_{2,1} &= \arg \left[ \sum_{k\in C2} P_{1,k} . P'_{1,k} * \right] \forall k \in C2 \tag{27}
\end{align*}

The effect caused by the sampling frequency offset can be corrected as in [6] by adjusting the sampling frequency of the receiver DAC. The rotation of sub-carriers are compensated by,
\begin{equation}
X'_{1,k} = X_{1,k} . \exp ( -j . 2\pi . k . t'_{\delta} (L.N_s+N_g)/N), \forall k\in(-K/2, K/2-1) \tag{28}
\end{equation}

4. SIMULATION STUDIES and DISCUSSION

To simulate the received random packets at the receiver, the packets are passed through various channel impulse responses (a typical example is shown in Fig.4), some amount of phase noise and frequency offset are also added to the samples. Then AWGN of random length is appended to the transmitted packet and also with the contents of the packet.

Some simulation studies carried on the packet detection algorithm for HyperLan channel models [3], indicate that a threshold of 0.4 yields a probability of 0.8, over 2 STS i.e. 32 samples. The constraints of the fixed point word length, the size of the buffer to store the previous values, the small size of the training sequence in WLAN and the time required to perform other algorithms restricts the value of M i.e. the number of cross-correlators in the frame synchronization algorithm to be less than equal to 4.

As for the frequency offset estimation is concerned, during STS of the preamble of 802.11 a/g packet, the value of L = 144, and the value of D = 16. This can be used for the initial coarse frequency estimation. Maximum coarse frequency offset that can be estimated is \( f_{\delta \text{max}} = 1/(2.16.50\text{ns}) = 625 \text{ KHz}. \) For fine Frequency offset estimation the 2 long training sequences (LTS) can be used where, L = 64, D = 64. The Maximum estimated fine frequency offset can therefore be \( f_{\delta \text{max}} = 1/(2.64.50\text{ns}) = 156.250 \text{ KHz}. \) The efficiency of the frequency and phase compensation algorithm are shown in Fig.5. Fig. 6 shows the degradation due to sampling clock offset and the performance of the compensation algorithm. The Maximum sampling frequency error that can be supported as per the 802.11 a/g standard features is +/- 50 PPM (parts per million). i.e. one sample may be in error in every 1e6/50 = 20000 samples. The host is supposed to view the 802.11g card as a LAN card; therefore the largest octet length that the MAC may send to the PHY is 1500 octets equivalent to 40480 samples. Hence a maximum error of about 2 samples can be tolerated and can be taken care by rob/stuff block.

5 CONCLUSION

In this paper we have considered the systematic design of receiver synchronization algorithms for OFDM-based WLAN systems taking into account the 802.11a/g standards and the WLAN transmission scenario.

The receiver algorithms have high performance in AWGN and Rician channels. The performance degrades in the case of Rayleigh faded channels. The complexity of the receiver is normally dominated by the FFT and channel compensation rather than by the synchronization algorithms. The performance of the algorithm could be further improved by allowing significantly more complexity.

6. REFERENCES


Figure 1. Concept of Orthogonal frequencies.

Figure 2. Block schematic of the N-sub carrier OFDM system.

Figure 3. HyperLan channel model showing power delay profile & rms. delay spread, common to typical office and indoor environments.

Figure 4. HyperLan channel model showing power delay profile & rms. delay spread, common to typical office and indoor environments.

Figure 5. Constellation plots showing the effects of frequency and phase offset and the performance of the corresponding algorithm for QPSK modulation scheme at 20 dB SNR.

Figure 6. Constellation plots showing the effects of the sampling clock offset and the performance of the compensation algorithm for BPSK modulation at 15 dB SNR.