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A Technique for Dominant Path Delay Estimation in an OFDM System and Its Application to Frame Synchronization in OFDM Mode of WMAN

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Orthogonal frequency division multiplexing (OFDM) is a parallel transmission scheme for transmitting data at very high rates over time dispersive radio channels. In an OFDM system, frame synchronization and frequency offset estimation are extremely important for maintaining orthogonality among the subcarriers. Recently, several techniques have been proposed for frame synchronization in OFDM system. In multipath environment, the transmitted signal arrives at the receiver through direct and multiple delayed paths. In some cases, it is possible that power of the signal arriving through delayed path may be larger than that of the direct path (earliest path if there is no direct path). In those cases, estimate of the frame boundary may be shifted by a quantity equal to the delay of the dominant path. In such cases, there will be intersymbol interference (ISI) in the demodulated symbols unless the frame boundary estimate is preadvanced such that it dwells in the ISI-free portion of cyclic prefix or at the symbol boundary. In this paper, we propose a method for estimating the shift in the frame boundary estimate using a preamble having two identical halves. We assume that frame boundary and frequency offset estimation have been performed prior to the estimation of the shift. We also examine the quality of the frequency offset estimate when the frame boundary estimate is shifted from the desired value. The proposed method is applied to downlink synchronization in OFDM mode of WMAN (IEEE 802.16-2004). We use simulations to illustrate the usefulness of the method and also to support our assertions.

1. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) is a multicarrier modulation scheme for transmitting data at very high rates over multipath radio channels. Recently, OFDM has been adopted as a modulation technique in *wireless metropolitan area network* (WMAN) standard [1]. In OFDM system, timing and frequency synchronization are crucial for the retrieval of information (see [2]). Several techniques have been recently proposed for OFDM frame and frequency synchronization.

In multipath environment, the transmitted signal arrives at the receiver through multiple paths and the corresponding signals are delayed with respect to direct path or earliest path if there is no direct path. In some cases, power of the signal arriving through a delayed path is larger than that of the signal from the earliest path (we refer to the corresponding delayed path as the dominant path). In those cases, estimate of the frame boundary may be shifted by a quantity equal to the delay of the dominant path (the delay is measured with respect to the arrival time of the earliest path). In [3], Yang et al. proposed a method that uses pilot subcarriers throughout the frame for estimating the shift in the frame boundary. However, their approach requires huge memory to store around 10 estimates of channel impulse responses and also the received samples corresponding to 10 OFDM symbols, thereby introducing a delay of 10 OFDM symbols in the demodulation process. To correct the error in the boundary estimate due to channel dispersion, the authors in [4] suggest a method which is computationally expensive due to matrix operations involved. For finding the channel impulse response, their method takes around LM complex multiplications and additions where M is the length of the repeated training symbol segment and L is the length of the cyclic prefix (CP). Moreover, both methods [3, 4] find the channel impulse response using the received OFDM symbols starting from the preadvanced boundary, where preadvancement is chosen by an arbitrary amount, and find the first significant channel tap by testing the tap power in a window of L' using a threshold. This preadvancement may degrade the estimation performance in the cases where the preadvanced frame boundary falls in the interference portion of CP.

In this paper, we propose a method for estimating the shift in the frame boundary using a preamble having two identical halves. This is the same as the first symbol of the preamble considered in [5] and the second symbol of the preamble specified for downlink mode of WMAN-OFDM. Our method requires $2M$ complex multiplications for finding the channel impulse response and a memory to store samples of two received OFDM symbols. We assume that frame boundary and frequency offset estimation have been performed prior to the shift estimation. We examine the quality of the frequency estimate when the frame boundary estimate is shifted from the desired value. Though the method is based on the preamble cited above, it is not restricted to the way the frame boundary and frequency offset are estimated. The proposed method integrates very well with our synchronization technique [6], and for this reason, we give the steps of the algorithm of [6] for the sake of continuity. The method is applied to downlink synchronization in OFDM mode of WMAN (IEEE 802.16-2004). We use simulations to illustrate the usefulness of the method and also to support our assertions.

The rest of the paper is organized as follows. Section 2 gives briefly the background of the frame synchronization technique [6]. Quality of frequency offset estimate as given by the algorithm of [6], when the frame boundary estimate is shifted from the desired symbol boundary, is examined in Section 3 and simulation results to support our assertions in this regard are also given in this section. The proposed method of estimating the shift in the frame boundary and simulation results to illustrate its effectiveness in the frame synchronization are presented in Section 4. Performance of the method when applied to downlink synchronization in WMAN-OFDM mode is demonstrated through simulations in Section 5. Finally, Section 6 concludes the paper.

2. BACKGROUND

The samples of the base-band equivalent OFDM signal are given by

$$x(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X(k) e^{j2\pi k(n-L)/N}, \tag{1}$$

where $0 \le n \le N + L - 1$, N is the total number of carriers, $X(k)$ is the kth subsymbol, $j = \sqrt{-1}$, and L is the length of cyclic prefix (CP), which is assumed to be longer than the length of channel impulse response. The signal is transmitted through a frequency selective multipath channel. Let $h(n)$ denote the base-band equivalent discrete-time channel impulse response of length υ. A carrier frequency offset of ϵ (normalized with subcarrier spacing) causes a phase rotation of $2\pi\epsilon n/N$. Assuming a perfect sampling clock, the received samples of the OFDM symbol are given by

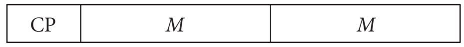

Figure 1: Preamble, preceded by CP, considered for the proposed method.

$$r(n) = e^{j[(2\pi\epsilon n/N)+\theta_0]} \sum_{l=0}^{\upsilon-1} h(l)x(n-l) + \eta(n), \tag{2}$$

where θ_0 is an initial arbitrary carrier phase and $\eta(n)$ is a zero mean circularly symmetric complex white Gaussian noise with variance σ_η^2. In this paper, we consider packet-based OFDM communication system, where a preamble is placed at the beginning of the packet. We consider an OFDM symbol with two identical halves, as shown in Figure 1, as the preamble. The two halves of this symbol are made identical (in time domain) by loading even carriers with a pseudonoise (PN) sequence. This is the same as the second symbol of the preamble specified in [1] and also the first symbol of the preamble considered in [5].

The samples of the transmitted preamble (excluding CP) are given by

$$a(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{M-1} A(2k) e^{j(2\pi/M)kn}, \quad 0 \le n \le N-1, \tag{3}$$

where $M = N/2$ with N denoting the symbol length, $A(2k)$, for $0 < k \le M - 1$, is the PN sequence that is loaded on even subcarriers, and $A(2k+1) = 0$ for $0 \le k \le M - 1$ [1]. The samples $a(n)$, for $n = 0, 1, \ldots, M - 1$, are assumed to be known to the receiver. We now introduce briefly the frame boundary and frequency estimation technique of [6].

The timing metric, as given in [6], is given by

$$M(d) = \frac{|P(d)|^2}{R^2(d)}, \tag{4}$$

where $P(d)$ and $R(d)$ are

$$P(d) = \sum_{i=0}^{M-1} [r(d+i)a(i)]^* [r(d+i+M)a(i)],$$
$$R(d) = \sum_{i=0}^{M-1} |r(d+i+M)|^2. \tag{5}$$

The superscript "$*$" denotes complex conjugation, $r(n)$ are the samples of the base-band equivalent received signal, and d is a sample index corresponding to the left edge of a window of $2M$ samples. $R(d)$ gives an estimate of the energy in M samples of the received signal. For a specified SNR, mean and variance of $M(d)$ are analytically evaluated for AWGN channel, and a threshold is selected based on them. We then estimate the symbol boundary as follows.

(i) Compute the timing metric $M(d)$ from a block of N received samples, shifting the block by one sample index each time, and find the sample index d_{th} where $M(d)$ crosses the threshold.

(ii) Evaluate $M(d)$ in the interval $d_{\text{th}} < d \leq (d_{\text{th}} + L - 1)$.

(iii) Find the sample index where $M(d)$ is largest in the interval $d_{\text{th}} \leq d \leq (d_{\text{th}} + L - 1)$. This sample index is taken as the estimate of the preamble boundary.

(iv) If the metric $M(d)$ does not cross the threshold at all, declare the detection as a miss detection.

After the symbol boundary estimate is found, we perform frequency offset estimation. Let the actual frequency offset be $\overline{\epsilon} = m + \delta$ with $m \in \mathcal{Z}$ and $|\delta| < 1$. Then, the estimate of

$$\epsilon = (m + \delta - \overline{m}) = \overline{\epsilon} - \overline{m} \tag{6}$$

is given by

$$\widehat{\epsilon} = \frac{\widehat{\phi}}{\pi}, \tag{7}$$

where

$$\widehat{\phi} = \text{angle}\left(P(d_{\text{opt}})\right) \tag{8}$$

with d_{opt} denoting the sample index corresponding to the estimate of the preamble boundary. Here, ϵ is the fractional part and $\overline{m}$ is the integer part such that their sum is the actual offset $m + \delta$. $\overline{m}$ is an even integer closest to $\overline{\epsilon}$ since the repeated halves of the preamble are the result of loading the even subcarriers with nonzero value and the odd subcarriers with zero value. The estimate of $\overline{m}$ is obtained from the bin shift as outlined below.

Let $r(d_{\text{opt}} + n)$, $n = 0, 1, \ldots, N - 1$, be the received preamble symbol. This sequence is first compensated with the fractional offset estimate $\widehat{\epsilon}$ giving

$$c(n) = e^{-j2\pi\widehat{\epsilon}n/N} r(d_{\text{opt}} + n), \quad n = 0, 1, \ldots, N - 1. \tag{9}$$

The bin shift estimation is then performed as follows.

(i) Obtain the product sequence $c(n)a^*(n)$ for $n = 0, 1, \ldots, M - 1$.

(ii) Evaluate the M-point DFT of the product sequence obtained in step (i).

(iii) Find the bin l_1 corresponding to the largest magnitude DFT coefficient. Then, $2l_1$ is the estimate of $\overline{m}$.

3. QUALITY OF FREQUENCY OFFSET ESTIMATE WHEN THE FRAME BOUNDARY ESTIMATE SHIFTS DUE TO DOMINANT PATH

Since we carry out estimation of the shift in the frame boundary estimate after compensating the received preamble with the frequency offset estimate, consisting of both fractional and integer parts, one needs to know if the quality of this estimate, as given by the algorithm of [6], is affected when the frame boundary estimate shifts due to dominant path.

Let τ be the difference between the estimate of the preamble boundary and the actual value which we denote by d_s, that is, $d_{\text{opt}} = d_s + \tau$. Then, samples of the received signal starting from d_{opt} are given by

$$r(d_{\text{opt}} + n) = e^{j\theta(n)} s(n + \tau) + \eta(d_{\text{opt}} + n), \quad 0 \leq n \leq N - \tau - 1, \tag{10}$$

where $s(n) = \sum_{p=0}^{\upsilon-1} h(p) a((n - p) \bmod M)$ and $\theta(n) = 2\pi\overline{\epsilon}(n + \tau)/N + \theta_0$ for $0 \leq n \leq N - \tau - 1$.

Recall that the fractional frequency offset is estimated from the angle of $P(d_{\text{opt}})$, which is given by

$$P(d_{\text{opt}}) = \sum_{i=0}^{M-1} r^*(d_{\text{opt}} + i) r(d_{\text{opt}} + i + M) \left| a(i) \right|^2. \tag{11}$$

Substituting (10) in (11) and using the relation (6), we get

$$\begin{aligned} P(d_{\text{opt}}) = {} & e^{j\pi\epsilon} \sum_{i=0}^{M-\tau-1} \left| s(\tau + i) \right|^2 \left| a(i) \right|^2 \\ & + \sum_{i=0}^{M-\tau-1} e^{-j\theta(i)} s^*(\tau + i) \eta(d_{\text{opt}} + M + i) \left| a(i) \right|^2 \\ & + \sum_{i=0}^{M-\tau-1} e^{j[\theta(i)+\pi\epsilon]} s(\tau + i) \eta^*(d_{\text{opt}} + i) \left| a(i) \right|^2 \\ & + \sum_{i=0}^{M-\tau-1} \eta^*(d_{\text{opt}} + i) \eta(d_{\text{opt}} + M + i) \left| a(i) \right|^2 \\ & + \sum_{i=M-\tau}^{M-1} r^*(d_{\text{opt}} + i) r(d_{\text{opt}} + M + i) \left| a(i) \right|^2. \end{aligned} \tag{12}$$

Note that the first four terms will be present even when τ=0. For $\tau \neq 0$, the magnitude of the last term is very small compared to that of the first since $r(d_{\text{opt}} + n)$ and $r(d_{\text{opt}} + M + n)$ for $n \geq M - \tau$ correspond to short segments of the received preamble and data symbols, respectively, which are uncorrelated. Though the upper limit in the sum for the first four terms differs from $M - 1$, its effect on the estimate of fractional part ϵ is negligible because of the following.

(i) In practice, τ is very small compared to M.

(ii) The magnitudes of the second, third, and fourth terms are very small compared to that of the first because signal and noise are assumed to be uncorrelated and noise is assumed to be white.

We, therefore, conclude that the quality of the estimate of fractional part ϵ will be nearly the same as that when there is no shift in the frame boundary estimate. Simulation results given below support this statement.

The integer part is estimated in [6] by finding the M-point DFT of the sequence $c(n)a^*(n)$ (see (9) for $c(n)$)

$$R_{\text{AC}}(2l) = \sum_{i=0}^{M-1} c(i) a^*(i) e^{-j2\pi li/M} \tag{13}$$

and the index $(2l_1)$, where l_1 corresponds to the DFT coefficient with largest magnitude, gives the estimate of integer part $\overline{m}$. From (9) and (10), and assuming that variance of the error $(\epsilon - \hat{\epsilon})$ is very small, (13) can be expressed as

$$R_{AC}(2l) \approx \sum_{i=0}^{M-1} e^{j[(2\pi/N)\overline{m}i+\theta_1]} \times \left\{ \sum_{p=0}^{\upsilon-1} h(p)a((i+\tau-p) \bmod M)a^*(i)e^{-j2\pi li/M} \right\} + \sum_{i=0}^{M-1} \eta'(d_{\text{opt}}+i)a^*(i)e^{-j(2\pi/N)li}, \quad (14)$$

where $\theta_1 = (2\pi/N)\overline{\epsilon}\tau + \theta_0$ and $\eta'(d_{\text{opt}}+i) = e^{-j2\pi\hat{\epsilon}}\eta(d_{\text{opt}}+i)$. Interchanging the summations, (14) can be rewritten as

$$R_{AC}(2l) \approx e^{j\theta_1}h(\tau)\sum_{i=0}^{M-1} |a(i)|^2 e^{j(2\pi/M)(\overline{m}/2-l)i} + e^{j\theta_1}\sum_{p=0,\, p\neq\tau}^{\upsilon-1} h(p)\left\{ \sum_{i=0}^{M-1} a((i+\tau-p) \bmod M) \times a^*(i)e^{j(2\pi/M)(\overline{m}/2-l)i} \right\} + \sum_{i=0}^{M-1} \eta'(d_{\text{opt}}+i)a^*(i)e^{-j2\pi/Nli}. \quad (15)$$

We note, once again, that the three terms will be present even when τ=0, and the last term is independent of τ. For $2l = \overline{m}$, the magnitude of the first term is much larger than that of the second for the following reasons.

(i) The magnitude of the coefficient $h(\tau)$, corresponding to the dominant path, is assumed to be significantly larger than the magnitudes of other coefficients.

(ii) The quantities corresponding to the sum in the first term and the inner sum in the second term represent cyclic autocorrelation of the sequence $a(i)$ for zero and $(\tau - p)$ lags, respectively, for $2l = \overline{m}$. Since $a(i)$, $0 \leq i \leq M-1$, is the time domain sequence obtained from a frequency domain loading by a PN sequence (see (3)), its zero-lag autocorrelation magnitude is much larger than that of a nonzero lag. The magnitude of lag-one autocorrelation, which is the next largest value, is found to be 12 dB lower.

In view of the above, the estimate of integer part, $\overline{m}$, will be nearly the same as the value we get when there is no shift in the frame boundary. Simulations given below support this assertion.

Simulations

To examine the quality of frequency offset estimate in the presence of a shift in the frame boundary estimate in frequency selective channels, we have performed simulations. The preamble is generated with 200 used carriers, 56 null carriers—28 on the left and 27 on the right, and a dc carrier. The even (used) carriers are loaded with a PN sequence given in [1] for OFDM mode. A frequency offset of 10.5 times the subcarrier spacing and a cyclic prefix of length 32 samples are assumed in the simulations. Stanford University Interim (SUI) channel modeling [7] is used to simulate a frequency selective channel. The impulse response of the channel is normalized to unit norm. Variance of a zero mean complex white Gaussian noise that is added to the signal component, which is the transmitted preamble in AWGN case and is the convolution of transmitted preamble and channel impulse response in the case of SUI channels, is adjusted according to the required SNR. A SNR of 9.4 dB is assumed in the simulations as it is the recommended SNR for the preamble [1]. The received signal generated as above is preceded by noise and followed by data symbols. We considered 250 different realizations of SUI channels where the second tap of the channel impulse response, corresponding to $\tau = 5$, is largest in amplitude. The fractional and integer frequency offsets are estimated using the algorithm of [6] for all these realizations. From the 250 results, the mean and standard deviations of the fractional frequency offset estimate, and the number of times the integer frequency offset estimate is equal to the true value, are found (see Table 1). We note from the results that the mean of the fractional frequency offset estimate is very close to the true value and the standard deviation is very small. The integer frequency offset estimate is 10 in 97% of the realizations.

Table 1: Integer frequency offset estimate $\hat{\overline{m}}$, mean and standard deviations of fractional frequency offset estimate $\hat{\epsilon}$ (actual frequency offset = 10.5 and SNR = 9.4 dB).

Channel type	Mean	Standard deviation	Integer frequency offset estimate
SUI-1	0.5007	0.0104	10 in 243 realizations
SUI-2	0.5009	0.0107	10 in 243 realizations
SUI-3	0.5007	0.0103	10 in 243 realizations

4. ESTIMATION OF THE SHIFT IN THE FRAME BOUNDARY

After estimating the integer frequency offset, the same M samples $c(n)$ (see (9)) are further compensated with this estimate as given below

$$b(n) = e^{-j2\pi(\hat{\overline{m}})n/N}c(n), \quad n = 0, 1, \ldots, M-1. \quad (16)$$

From (9) and (10), and assuming that $\hat{\overline{m}} = \overline{m}$ and variance of error $(\hat{\epsilon} - \epsilon)$ is very small, (16) can be expressed as

$$b(n) \approx e^{-j\theta_1}\sum_{l=0}^{\upsilon-1} h(l)a((n+\tau-l) \bmod M) + \eta''(d_{\text{opt}}+n), \quad (17)$$

where $\eta''(d_{\text{opt}} + n) = e^{-j(2\pi/N)(\widehat{\widehat{m}}+\widehat{\epsilon})n}\eta(d_{\text{opt}} + n)$ and $\theta_1 = (2\pi/N)\overline{\epsilon}\tau + \theta_0$. Taking M-point DFT of $b(n)$, we have

$$B(k) = \frac{1}{\sqrt{M}} \sum_{n=0}^{M-1} b(n)e^{-j(2\pi/M)kn} \tag{18}$$

which, in view of (17), can be written as

$$\begin{aligned} B(k) = {} & \frac{e^{-j\theta_1}}{\sqrt{M}} \sum_{n=0}^{M-1} \sum_{l=0}^{\upsilon-1} \{h(l)a((n+\tau-l) \bmod M)e^{-j(2\pi/M)kn}\} \\ & + \frac{1}{\sqrt{M}} \sum_{n=0}^{M-1} \eta''(d_{\text{opt}}+n)e^{-j(2\pi/M)kn}. \end{aligned} \tag{19}$$

Note here that the spacing of DFT bins $B(k)$ is twice that of $A(k)$.

Substituting (3) into (19) and simplifying, we get

$$B(k) = \sqrt{\frac{M}{2}} e^{-j\theta_1} A(2k)H(k)e^{j(2\pi/M)k\tau} + W(k), \tag{20}$$

where $H(k) = (1/\sqrt{M}) \sum_{l=0}^{\upsilon-1} h(l)e^{-j(2\pi/M)lk}$ and $W(k) = (1/\sqrt{M}) \sum_{n=0}^{M-1} \eta''(d_{\text{opt}} + n)e^{-j(2\pi/M)nk}$. The estimate of $H(k)$ is then given by

$$\widehat{H}(k) = \frac{B(k)}{A(2k)} = \sqrt{\frac{M}{2}} e^{-j\theta_1} H(k)e^{j(2\pi/M)k\tau} + \frac{W(k)}{A(2k)}. \tag{21}$$

Taking M-point IDFT of (21) gives

$$\widehat{h}(n) = \sqrt{\frac{M}{2}} e^{-j\theta_1} h(n+\tau) + w(n), \tag{22}$$

where

$$w(n) = \left(\frac{1}{\sqrt{M}}\right) \sum_{k=0}^{M-1} \left(\frac{W(k)}{A(2k)}\right) e^{j(2\pi/M)kn} \tag{23}$$

is a zero mean complex Gaussian noise with variance $(\sigma_\eta^2/M) \sum_{k=0}^{M-1} (1/|A(2k)|^2)$, and

$$\widehat{h}(n) = \left(\frac{1}{\sqrt{M}}\right) \sum_{k=0}^{M-1} \widehat{H}(k)e^{j(2\pi/M)kn}. \tag{24}$$

We make use of the FFT block available at the receiver twice, once for computing $B(k)$ and next for computing $\widehat{h}(n)$. The method requires $2M$ complex multiplications for the estimation of channel impulse response. The use of the FFT block twice introduces delay of approximately two OFDM symbols. Consequently, the method requires a memory to store received samples corresponding to two OFDM symbols before the demodulation of the received OFDM data symbols.

Using the estimated channel impulse response, we test the tap power in a window of L' to find the shift in the estimate of the impulse response $h(n)$, corresponding to the shift in the frame boundary estimate τ. Since the preamble we use has two identical halves, picking up M samples from the estimated boundary will not fall in the interference portion of CP. As $\tau \le L$, τ is estimated as

$$\widehat{\tau} = M - \arg\max_d \{\mathcal{E}_h(d) : M - L \le d \le M - 1\}, \tag{25}$$

where $\mathcal{E}_h(d)$ is the energy in the estimated channel impulse response in a window of length L', given by

$$\mathcal{E}_h(d) = \sum_{l=0}^{L'-1} |\widehat{h}((d+l) \bmod M)|^2. \tag{26}$$

An appropriate value for L' is the channel delay spread υ since the dominant path corresponds at most to the last coefficient of the channel response. Since we do not have a priori knowledge of the channel length, one can choose $L' = L$. If $L' > \upsilon$, $\mathcal{E}_h(d)$ will have a plateau of length $L' - \upsilon$. If we preadvance the frame boundary estimate with $\widehat{\tau}$ obtained from (25), it falls in the interval covering ISI-free portion of CP or at the preamble boundary. With the corrected frame boundary estimate, there will be no ISI in the demodulated OFDM symbols at the output of FFT.

Simulations

To illustrate the usefulness of the proposed dominant path delay estimation, we conducted the following simulations. We generated the preamble and noise as described in Section 3. We estimated the frame boundary, d_{opt}, and frequency offset from the signal modeled as in (2) following the technique of [6], and then estimated the dominant path delay τ as described above. We then preadvanced d_{opt} by $\widehat{\tau}$. Starting from the corrected boundary estimate, we selected N samples from the noise-free and frequency offset-free signal modeled as

$$r(n) = \sum_{l=0}^{\upsilon-1} h(l)x(n-l) \tag{27}$$

and computed its DFT. If the DFT coefficients $R(k)$ are zero for odd values of k, we declare the detection of the frame boundary as correct. Otherwise, we declare it as a false detection. We repeated this step for the uncorrected boundary estimate case too. This measure is equivalent to checking if the frame boundary estimate, after correction by $\widehat{\tau}$, falls in the interval covering ISI-free portion of CP or at the frame boundary.

We considered 1000 realizations of AWGN, SUI-1, SUI-2, and SUI-3 channels, choosing a different realization of noise and channel in each case. The results are shown in Table 2. We note from the results that preadvancement of the preamble boundary estimate by $\widehat{\tau}$ enhances the detection performance, in particular for SUI-2 and SUI-3 channels.

To further demonstrate the practical utility of the dominant path delay estimation method described here, we performed simulations to find bit error rate (BER) with and without dominant path delay estimate compensation. The

Table 2: Detection performance with and without preadvancement of the frame boundary estimate (number of realizations = 1000, SNR = 9.4 dB).

Channel	without preadvancement		with preadvancement	
	False detections	Correct detections	False detections	Correct detections
AWGN	0	1000	0	1000
SUI-1	3	996	0	999
SUI-2	40	952	6	986
SUI-3	198	752	6	948

simulation setup is the same as the one described in Section 3 except that we considered 1000 realizations of SUI-1, SUI-2, and SUI-3 channels whose second tap has the largest magnitude among all the three taps. We appended OFDM data symbols to the preamble, where these symbols are generated by loading data subcarriers with BPSK or QPSK subsymbols as in [1], to form an OFDM packet. We did not use any forward error correction (FEC) technique in the simulations. For each realization of the channel, the generated OFDM packet is convolved with channel impulse response and noise is added to the convolved output to give the received packet.

We performed the frame and frequency synchronization on each received packet as described in Section 2. We obtained the estimate of the channel frequency response at the even carriers from (21) and estimated the frequency response at the odd carriers by linear interpolation. We then performed frequency domain equalization of the received data symbols. After demapping the equalized subsymbols, we computed BER. We repeated this with dominant path delay estimate compensation. The average BER computed from the results of 1000 realizations is shown in Figures 2 and 3.

From Figures 2 and 3, we observe that without preadvancement of the preamble boundary estimate by the dominant path delay estimate the BER remains almost constant beyond 20 dB SNR, while with preadvancement it falls with SNR. These results clearly bring out that preadvancement of the preamble boundary estimate with the dominant path delay estimate is necessary.

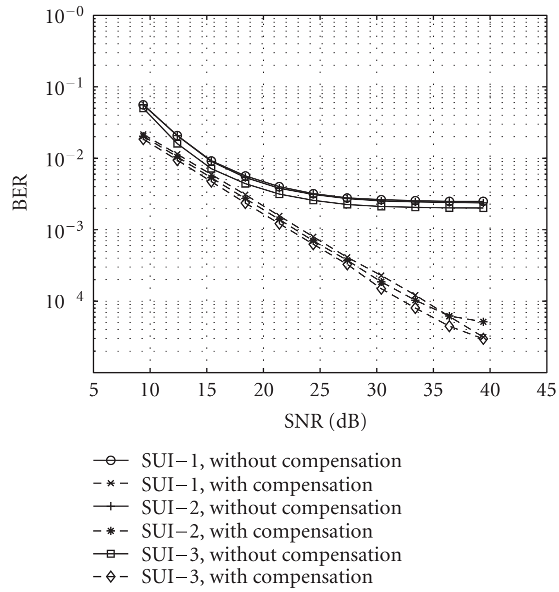

Figure 2: BER versus received SNR, with and without dominant path delay estimate compensation, in SUI channels with BPSK mapping.

5. APPLICATION TO DOWNLINK SYNCHRONIZATION IN WMAN-OFDM

The WMAN-OFDM physical layer is based on the OFDM modulation with 256 subcarriers. For this mode, the preamble consists of two OFDM symbols. Each of these symbols is preceded by a cyclic prefix (CP), whose length is the same as that for data symbols. In the first OFDM symbol, only the subcarriers whose indices are multiple of 4 are loaded. As a result, the time domain waveform (IFFT output) of the first symbol consists of 4 repetitions of 64-sample fragment. In the second OFDM symbol, only the even subcarriers are loaded which results in a time domain waveform consisting of 2 repetitions of 128-sample fragment. The corresponding preamble structure is shown in Figure 4. In the downlink synchronization, we have to estimate the symbol boundary, frequency offset, and the CP value using the preamble given in Figure 4. To evaluate the CP value, we estimate the left edge of one of the 64-sample segments and the left edge of the second symbol following the approach given in [6], and evaluate the CP from

$$\hat{L} = Q[(d_2 - d_1 - 1) \bmod 64], \quad (28)$$

where d_1 and d_2 are sample indices corresponding to the estimates of the left edges of one of the first three segments of the first symbol and the second symbol, respectively. d_1 and d_2 are obtained using the practical detection strategy described in [6]. The function $Q(x)$ denotes quantization of x to a value nearest to 8, 16, 32, and 64 (or 0), corresponding to CP lengths of 8, 16, 32, and 64, respectively. Then, we proceed with the estimation of shift in the left edge of the second symbol and preadvance it by the estimate $\hat{\tau}$.

We conducted the simulations using the preamble shown in Figure 4 and the same simulation setup as in the previous section. The results are given in Table 3. We note from the results that preadvancement enhances the correct detection of the frame boundary, particularly in SUI-2 and SUI-3 channels. The results show that the CP estimate is correct even in

TABLE 3: Detection of the frame boundary with and without preadvancement by $\hat{\tau}$, and the number of times the CP is estimated correctly (number of trials = 1000, SNR = 9.4 dB).

Channel	Before preadvancement			After preadvancement	
	False	Correct	CP	False	Correct
AWGN	0	1000	1000	0	1000
SUI-1	3	997	1000	0	1000
SUI-2	32	961	987	2	991
SUI-3	174	791	937	5	960

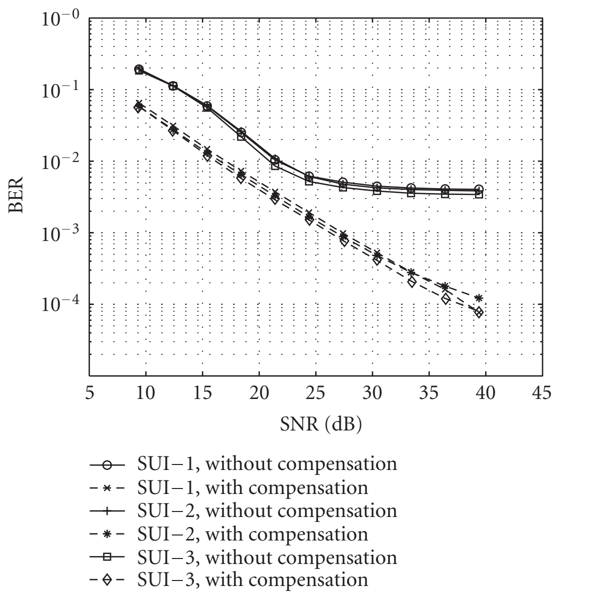

FIGURE 3: BER versus received SNR, with and without dominant path delay estimate compensation, in SUI channels with QPSK mapping.

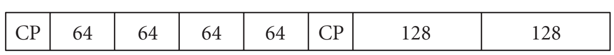

FIGURE 4: Downlink preamble structure in WMAN-OFDM mode.

those cases where the left edge of the second symbol is not detected correctly. This is because shifts in the left edges of the detected segment and the second symbol do not affect the CP estimate if the difference between the shifts is less than 4 samples. We have found that estimating the shifts and correcting the edge estimates prior to evaluating the CP did not make any difference in the CP estimates.

The small difference in the number of correct detections between Table 2 and Table 3 may be because (i) noise realizations in the two cases are different and (ii) search intervals over which we find the maximum value of the metric are different (the interval is 32 for the results of Table 2 while it is 64 for those of Table 3). The number of misdetections in the case of SUI-3 channel is fewer in WMAN-OFDM case because computation of the timing metric for the detection of the symbol boundary starts earlier in this case.

6. CONCLUSIONS

In a multipath radio environment, the transmitted signal arrives at the receiver through multiple delayed paths. In some cases, the power of the signal arriving through a delayed path may be larger than that of the earliest path. In those cases, the estimate of the frame boundary may be shifted by a quantity equal to the delay of the dominant path. We presented a method to estimate the shift in the frame boundary caused by the dominant path. We also examined the quality of the frequency offset estimate, given by [6], in the presence of a shift in the frame boundary. We illustrated the usefulness of the correction of the frame boundary estimate through simulations. The simulation results show that without preadvancement of the frame boundary estimate by the dominant path delay estimate, the BER saturates beyond 20 dB SNR of the received signal with BPSK/QPSK mapping while with preadvancement it falls with SNR. We have also discussed the application of the proposed method to downlink synchronization in WMAN-OFDM and presented some simulation results to demonstrate its utility.

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