

A NEW SYMBOL TIMING RECOVERY ALGORITHM FOR OFDM SYSTEMS[†]

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Abstract—A new, patent-pending, symbol timing recovery algorithm for OFDM based systems is described for which supplementary pilot signals are not required. The proposed algorithm is composed of a coarse timing acquisition stage which does not require prior carrier synchronization and a fine symbol time tracking stage which requires (fine) carrier frequency synchronization so that the residual frequency offset is a multiple of the subcarrier spacing. However, the results of the coarse timing acquisition provides information required for the fine carrier frequency synchronization thus making the algorithm self-contained. Simulation results show that the proposed symbol timing recovery algorithm gives satisfactory performance for E_b/N_0 as low as 0dB.

I. INTRODUCTION

UNLIKE traditional single carrier transmission techniques, OFDM (Orthogonal Frequency Division Multiplexing) modulation employs multiple orthogonal subcarriers for the modulation of a parallel data stream [1]. Compared with single carrier transmission systems, OFDM modulation offers improved performance under frequency selective multipath and impulsive noise channels [2]. Also, OFDM modulation has the advantage of the flexibility in the use of the assigned spectrum and is suitable for constructing Single Frequency Networks (SFN.) Due to these advantages, the OFDM modulation scheme was adopted as the modulation scheme for a DAB (Digital Audio Broadcasting) system [3] and ADSL (Asymmetry Digital Subscriber Loop) [4] and is also proposed as the terrestrial HDTV transport in Europe [5].

Like any other digital communications system, accurate synchronization of various signal epochs in the OFDM system is crucial for proper receiver operation. One of these synchronization tasks is to locate and

track the symbol boundaries between consecutively received OFDM symbols and remove the guard interval inserted to mitigate intersymbol interference (ISI.) Symbol timing offsets in OFDM receivers do not only induce ISI but also cause linear phase rotations of the signal constellation of the receiver Fast Fourier Transform (FFT) outputs [6] which degrades system performance.

In this paper, we propose patent-pending, coarse and fine symbol timing recovery algorithms for OFDM systems. The coarse timing acquisition algorithm, unlike many previously proposed algorithms [6-11], does not require the transmission of a supplementary pilot signal nor requires carrier synchronization as a prerequisite. The fine time tracking algorithm, which is also based on an NDA (Non Data-Aided) approach, does not require a pilot signal but requires fine carrier synchronization so that the residual carrier frequency offset is a multiple of the subcarrier spacing.

The organization of this paper is as follows. In Section II, the OFDM system under consideration is briefly described. In Section III, the proposed symbol timing recovery algorithm is presented and in Section IV, simulation results are provided to evaluate the performance of the proposed algorithms. Finally, conclusions are drawn in Section V.

II. OFDM SYSTEM

We 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 used as the guard band [1]. Fig. 1 shows the block diagram of the OFDM system under consideration.

At the transmitter, N complex data symbols (including null data symbols for virtual carriers,) assumed to have a QAM constellation, are modulated onto the

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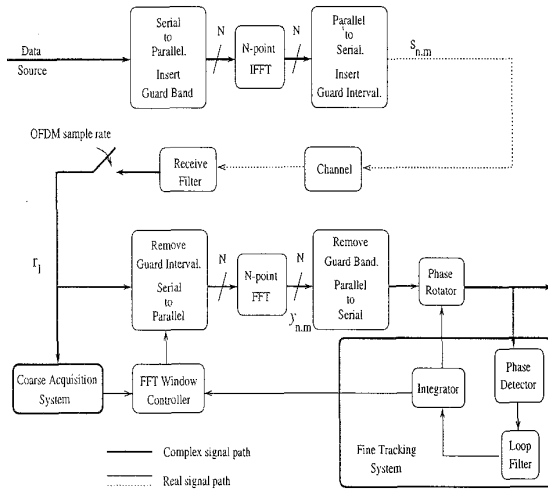


Fig. 1. OFDM system

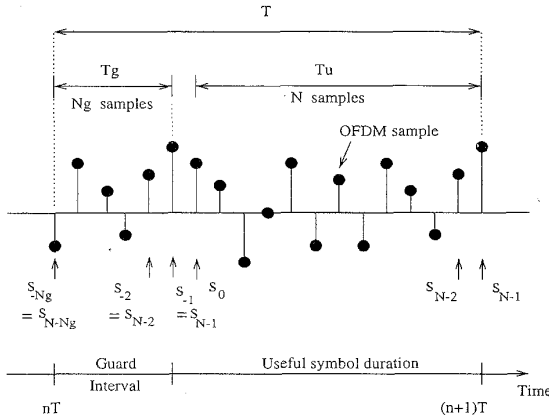


Fig. 2. OFDM symbol structure

N subcarriers by using Inverse Fast Fourier Transform (IFFT.) The last N_g samples of the IFFT outputs are then copied and added to form the guard interval at the beginning of each OFDM symbol. Therefore the serial OFDM samples $S_{n,m}$ composing the n th OFDM symbol (Fig. 2) are given as follows:

$$S_{n,m} = \frac{1}{\sqrt{N}} \sum_{k=-\frac{N}{2}}^{\frac{N}{2}-1} a_{n,k} e^{j\frac{2\pi}{N}mk}, -N_g \leq m \leq N-1, \quad (1)$$

$$= S_{n,m+N}, \quad -N_g \leq m \leq -1, \quad (2)$$

where $a_{n,k}$ is the data symbol modulating the k th subcarrier during the n th OFDM symbol duration. Under the assumption that $a_{n,k}$ are independent and identically distributed (i.i.d.) zero-mean random variables

and the number of subcarriers is sufficiently large, $S_{n,m}$ may safely be approximated as zero-mean complex Gaussian random variables with the following correlation:

$$E\{S_{n,i}S_{m,j}^*\} = \begin{cases} \sigma_a^2, & n = m, \quad i = j, \\ \sigma_a^2, & n = m, \quad |i - j| = N, \\ 0, & \text{otherwise.} \end{cases} \quad (3)$$

The OFDM samples $S_{n,m}$ are then pulse-shaped by a transmit filter with frequency response $B(f)$ and are transmitted over a multipath channel.

At the receiver, the output of the receiver matched filter with frequency response $B^*(f)$ is sampled at the OFDM sample rate by an Analog-to-Digital Converter (ADC.) After transmission over a multipath channel with M rays (paths,) the ADC outputs r_l with a carrier frequency offset of f_0 Hz are given as

$$r_l = \sum_{p=0}^{M-1} S_l - \tau_p \rho_p e^{j(\frac{2\pi}{N}lf_0 + \theta_p)} + n_l, \quad -\infty < l < \infty, \quad (4)$$

where S_l is the signal component, ρ_p , θ_p and τ_p are, respectively, the amplitude attenuation, the phase rotation and the delay in OFDM samples corresponding to the p th path. Also, n_l is a zero-mean complex Gaussian random variable with variance $\frac{N_0}{2}|B(f)|^2$ representing the contribution of the AWGN with a double-sided power spectral density of $\frac{N_0}{2}$. For the special case of an AWGN channel (no multipath,) (4) can be simplified as follows:

$$r_l = S_l e^{j\frac{2\pi}{N}lf_0} + n_l, \quad -\infty < l < \infty. \quad (5)$$

After ADC, the guard interval is removed and the remaining N useful samples are demodulated by using FFT. From the basic properties of FFT and (4), the FFT output $y_{n,m}$ corresponding to the m th subcarrier for the n th OFDM symbol is given as follows for the case when the symbol timing offset is small enough for the ISI term to be safely neglected [12-13]:

$$y_{n,m} = a_{n,m-k} H_m e^{j\frac{2\pi}{N}\zeta m} + W(f_\epsilon) + N_{n,m}, \quad -N/2 \leq m < N/2, \quad (6)$$

where H_m is the frequency response of the multipath channel at the m th subcarrier frequency, ζ is the symbol timing offset in OFDM samples, $W(f_\epsilon)$ is the InterChannel Interference (ICI) term [13] due to a carrier frequency offset relative to the f_ϵ defined by

$$f_0 = k\Delta f + f_\epsilon, \quad |f_\epsilon| < \frac{\Delta f}{2}, \quad (7)$$

where Δf is the subcarrier spacing and k is an integer.

From (6), we observe that the symbol timing offset ζ causes a linear rotation in the phase of the FFT outputs by an amount of $e^{j\frac{2\pi}{N}\zeta m}$ so that subcarriers corresponding to large values of m may seriously be affected even by small symbol timing offsets. Therefore, the design of an efficient and accurate symbol timing recovery algorithm is crucial in obtaining a satisfactory performance in OFDM systems.

III. PROPOSED SYMBOL TIMING RECOVERY ALGORITHM

In the proposed algorithm, the symbol timing recovery procedure is broken down into two stages - a coarse timing acquisition/fine carrier frequency synchronization stage and a fine symbol time tracking stage (Fig. 3.)

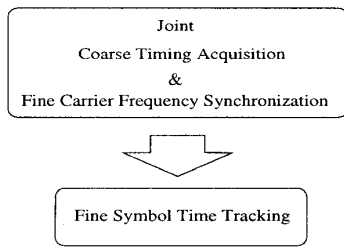


Fig. 3. Procedure of proposed symbol timing recovery

The coarse timing acquisition stage, which is shown in Fig. 3, acquires symbol timing based on a time-domain (pre-FFT) NDA approach. The proposed coarse acquisition algorithm has the advantage of low hardware complexity compared to other time-domain processing algorithms [8,9] and its operation is independent of carrier frequency offsets. It is shown in the simulation results that the expected residual timing offset after completion of the coarse timing acquisition stage is approximately 1% of the OFDM symbol duration. Subsequent fine tracking stage will track out this residual timing error.

A fine carrier frequency synchronization stage (compensating for f_e in (7)) is required for proper operation of the fine time tracking algorithm due to the large ICI term $W(f_e)$ in (6) for large values of $|f_e|$. In order for the proposed fine time tracking algorithm to operate for E_b/N_0 down to 0dB, f_e should be limited to within 4% of the subcarrier spacing [13]. In Subsection B, we briefly describe how the results obtained from the

coarse timing acquisition stage may be used for fine carrier frequency synchronization using the algorithm proposed in [14]. This combination of coarse timing acquisition and fine carrier tracking algorithms has the advantages of jointly achieving fine carrier frequency synchronization with coarse timing acquisition and has lower hardware complexity than algorithms proposed in [8,9]. For fine carrier frequency synchronization, other algorithms insensitive to the residual timing offsets after coarse timing acquisition are also applicable such as that proposed in [15].

Unlike the coarse timing acquisition stage, the fine symbol time tracking stage is based on frequency-domain (post-FFT) processing of the received signal. The fine symbol time tracking stage compensates for the residual timing offset after the initial coarse timing acquisition and tracks the timing drift of FFT window position due to sampling clock frequency offsets. The proposed fine tracking algorithm is also based on an NDA approach and its operation is independent of the remaining carrier frequency offset of a multiple of the subcarrier spacing remaining after fine carrier recovery.

A. Coarse Timing Acquisition Algorithm

In the coarse timing acquisition stage, a guard interval of length N_g samples is used to extract the coarse symbol timing information. A block diagram of the coarse timing acquisition algorithm is shown in Fig. 4.

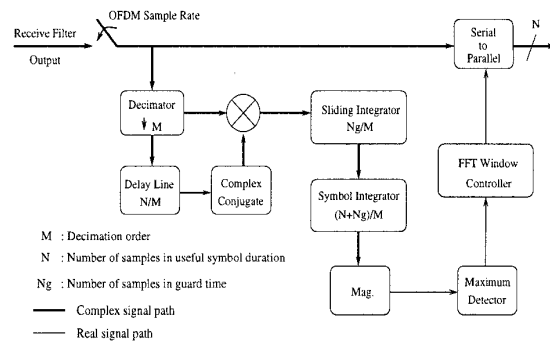


Fig. 4. Block diagram of the Coarse timing acquisition algorithm.

The ADC output r_l is decimated by a factor of M ($M < N_g$ and a divisor of $N + N_g$) and fed into a delay-line of length N/M . By selecting an appropriate $M > 1$, the implementation complexity of the coarse timing acquisition system may be reduced with-

out seriously degrading the performance.

The complex conjugate of the delay-line output $R_{n-N/M}$ is multiplied with the decimator output R_n as shown in Fig. 5, where A represents the region in which R_n and $R_{n-N/M}$ satisfy (2). From (2), (3) and

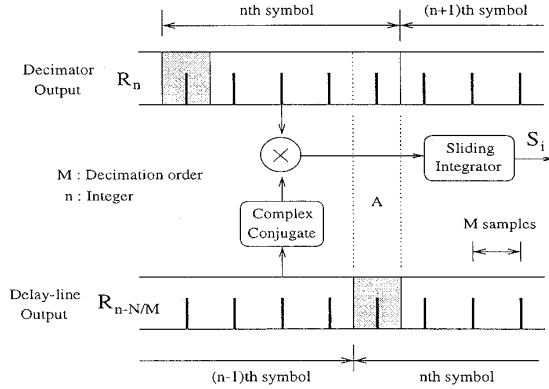


Fig. 5. Relationship between R_n , $R_{n-N/M}$ and S_i .

(7), we note that the complex multiplier output is a cyclostationary process with a period equal to one OFDM symbol duration. Therefore, for the case of an AWGN channel without multipath, the mean of the complex multiplier output within one period is given as follows:

$$E \{ R_n R_{n-N/M}^* \} = \begin{cases} \sigma_a^2 e^{2\pi f_\epsilon}, & n \in A, \\ 0, & \text{otherwise.} \end{cases} \quad (8)$$

The complex multiplier outputs are then summed by the sliding integrator to accumulate the energy of the periodic signal components due to the guard interval. Thus, the output S_i of the sliding integrator with a window size of N_g/M samples can be written as

$$S_i = \sum_{n=i-N_g/M+1}^i R_n R_{n-N/M}^*, \quad (9)$$

where N_g/M is the number of the decimator output samples in region A . Clearly, S_i is also cyclostationary with a period equal to one OFDM symbol duration whose mean provides the coarse symbol timing information.

We may accumulate (average) the sliding integrator outputs (symbol integrator in Fig. 4) in order to increase the reliability of the coarse timing information. The symbol integrator output I_i may be written as

$$I_i = \sum_{k=0}^{L-1} S_{i-\frac{N_g+N}{M}k}, \quad (10)$$

where L is the number of accumulations performed. By the cyclostationarity of S_i , the symbol integrator outputs corresponding to the guard interval steadily increase while those corresponding to random data do not.

Reliable coarse symbol timing is then acquired by finding the position corresponding to the maximum magnitude of the symbol integrator outputs within a symbol duration. The estimated coarse OFDM symbol timing $\hat{T}_c$ is therefore given as follows:

$$\hat{T}_c = \arg \max_i |I_i|. \quad (11)$$

B. Fine Carrier Synchronization Algorithm

From (8), we note that f_ϵ may be estimated by observing the complex multiplier outputs at index $i \in A$. Since complex multiplier inputs are separated by N OFDM samples, the effect of the remaining carrier frequency offset of a multiple of the subcarrier spacing is transparent to this fine carrier recovery algorithm.

Therefore, the phase of the sliding integrator output S_i at index $i = \hat{T}_c$ provides information on the carrier frequency offset f_ϵ as follows:

$$f_\epsilon = \frac{1}{2\pi} \text{Arg} \{ S_i \} \Big|_{i=\hat{T}_c}, \quad (12)$$

where $\text{Arg}\{x\}$ denotes the phase of a complex number x .

After fine carrier recovery using the algorithm introduced here, the carrier frequency offset of a multiple of the subcarrier spacing still remains, but this remaining carrier frequency offset does not affect the operation of the subsequent fine symbol time tracking algorithm.

C. Fine Symbol Time Tracking Algorithm

Simulation results given in Section IV show that after coarse timing acquisition is achieved, the residual symbol timing offset is confined to within several OFDM samples even at very low signal-to-noise ratios. Under these conditions, and after fine carrier frequency synchronization, (6) is valid and the ICI term $W(f_\epsilon)$ is negligible so that (6) may be simplified as follows:

$$y_{n,m} = a_{n,m-k} H_m e^{j\frac{2\pi}{N}\zeta m} + N_{n,m}, \quad ; -\frac{N}{2} \leq m < \frac{N}{2}. \quad (13)$$

Fig 6. illustrates the effects of a residual symbol timing offset ζ and a carrier frequency offset of a multiple of the subcarrier spacing on the receiver FFT output.

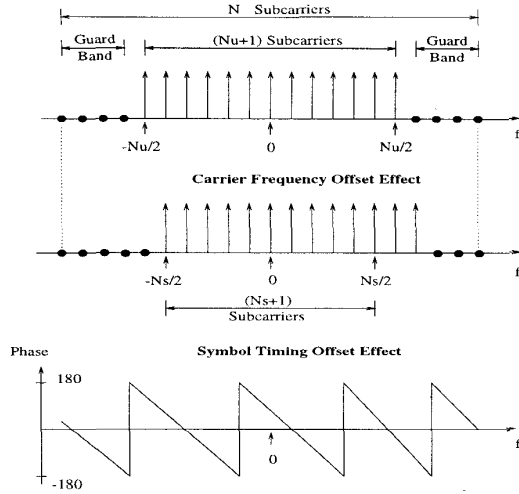


Fig. 6. Receiver FFT outputs showing the effects of residual timing and carrier frequency offsets of a multiple of the subcarrier spacing.

Note that the residual symbol timing offset ζ causes the constellation of receiver FFT outputs to linearly rotate by a factor of $e^{j\frac{2\pi}{N}\zeta m}$. Thus, we observe that the relative phase rotations between the adjacent subcarriers provide information on the residual timing offset. However, we need to extract the phase rotations due only to the residual timing offset, since the data modulation, channel distortion and background noise also affect the phase of the receiver FFT outputs.

For this purpose, the proposed fine symbol time tracking algorithm makes use of the fourth moment of the receiver FFT outputs. Fig. 7 shows the phase detector configuration of the proposed fine tracking system where we have assumed QAM modulation for the subcarriers.

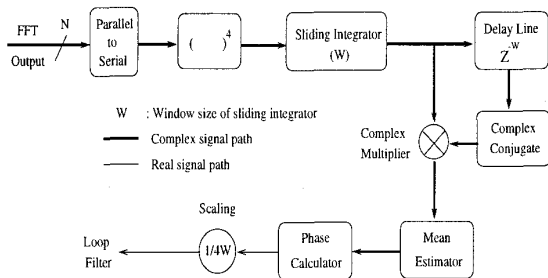


Fig. 7. Block diagram of the phase detector configuration for fine symbol time tracking algorithm.

First, the receiver FFT outputs $y_{n,m}$ are serially con-

verted and raised to the fourth power. For QAM modulation, we note that in spite of the random data modulation, $y_{n,m}^4$ bear information on the phase rotation due to the symbol timing offset, while $y_{n,m}$ do not.

The outputs of the fourth power device, $y_{n,m}^4$ are then summed by a sliding integrator with a window size of W in order to suppress the noise due to the random data modulation and background noise. The sliding integrator output $R_{n,j}$ for the N th symbol is therefore written as

$$R_{n,j} = \frac{1}{W} \sum_{k=-W/2}^{W/2-1} y_{n,j+k}^4, \quad |j| < \frac{N_s}{2} - \frac{W}{2}, \quad (14)$$

where N_s is defined as

$$N_s = N_u - 2k \quad (15)$$

with $k\Delta f$ being the remaining carrier frequency offset of a multiple of the subcarrier spacing after fine carrier recovery. Under the reasonable assumption that $2k$ is less than the number of virtual carriers, N_s is given as follows:

$$N_s = 2N_u - N. \quad (16)$$

Thus, the carrier frequency offset of a multiple of the subcarrier spacing effectively reduces the number of subcarriers available for fine symbol time tracking.

The relative phase rotations between the non-overlapping outputs of the sliding integrator are then calculated and averaged by the mean estimator. The complex mean phase rotation $P_n(\zeta)$ for a symbol timing offset of ζ samples is given as

$$P_n(\zeta) = \frac{1}{K} \sum_{j=0}^{K-1} R_{n,Wj} R_{n,W(j+1)}^*, \quad (17)$$

where K is the total number of non-overlapping sliding integrator outputs within $N_s + 1$ subcarriers.

By using $N_s + 1$ available subcarriers, the phase rotations due to channel distortion and background noise are averaged out so that their effect on the mean estimator output is minimized. In addition, by employing an NDA method, the proposed fine tracking algorithm is independent of residual carrier frequency offset of a multiple of the subcarrier spacing.

Finally, the resulting phase of the mean estimator output provides the fine symbol timing estimate $\hat{\zeta}$ to compensate for the residual timing offset as follows:

$$\hat{\zeta} = \frac{N}{8\pi W} \text{Arg}\{P_n(\zeta)\}. \quad (18)$$

IV. SIMULATION RESULTS

We have performed waveform level simulations to evaluate the performance of the proposed coarse acquisition and fine symbol time tracking algorithms. Simulations were performed under an AWGN channel without multipath and multipath channels having a multipath profile with 21 rays. Table 1 gives the assumed multipath channel profile - amplitude attenuations ρ_p , phase rotations θ_p and the delays τ_p of each ray. In ad-

p	ρ_p	$\tau_p[\mu s]$	$\theta_p[rad]$
0	3.396428	0.0	0.0
1	0.057662	1.003019	4.855121
2	0.176809	5.422091	3.419109
3	0.407163	0.518650	5.864470
4	0.303585	2.751772	2.215894
5	0.258782	0.602895	3.758058
6	0.061831	1.016585	5.430202
7	0.150340	0.143556	3.952093
8	0.051534	0.153832	1.093586
9	0.185074	3.324866	5.775198
10	0.400967	1.935570	0.154459
11	0.295723	0.429948	5.928383
12	0.350825	3.228872	3.053023
13	0.262909	0.838831	0.628578
14	0.225894	0.073883	2.128544
15	0.170996	0.203952	1.099463
16	0.149723	0.194207	3.462951
17	0.240140	0.924450	3.664773
18	0.116587	1.381320	2.833799
19	0.221155	0.640512	3.334290
20	0.259730	1.368671	0.393889

TABLE I
ASSUMED MULTIPATH PROFILE.

dition to the channel with delay profile given by Table 1 (multipath channel I,) we consider a second channel with delay profile also given by Table 1 but with path 0 (LOS path) absent (multipath channel II.) Fig. 8 shows the frequency responses of the multipath channels under consideration. For the multipath channel case, we measure the symbol timing offset with respect to the center of mass of the channel energy delay profile. Table 2 lists the center of mass for the two multipath channel models with respect to the first arriving path.

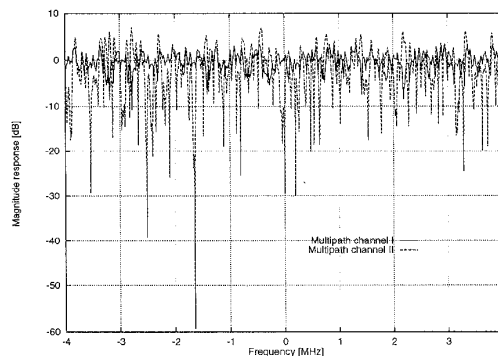


Fig. 8. Frequency responses of multipath channels I and II.

Channel	AWGN	Multipath I	Multipath II
[OFDM sample]	0	3.0	14.0

TABLE II
CENTER OF THE MASS OF THE CHANNEL ENERGY DELAY PROFILES

To validate the proposed algorithms, we have taken the system parameters from the European draft specification for terrestrial digital TV [15]. The OFDM system employs 2,048 subcarriers (including 343 virtual carriers) in a 7.61Mhz bandwidth for a useful symbol duration of $224\mu s$ and a guard interval of $7\mu s$ (1/32 useful symbol duration.)

In this section, simulation results for the fine carrier frequency synchronization loop are not presented and we simply assume that fine carrier recovery is jointly achieved with coarse timing acquisition so that ICI is negligible after the coarse timing acquisition stage.

A. Coarse Symbol Timing Acquisition

The simulation results for the coarse timing acquisition algorithm are shown in Figs. 9-10. Fig. 9 shows the outputs of the symbol integrator with a decimation order of $M = 8$ under multipath channel II. The symbol integrator outputs due to the guard interval steadily increase and clearly displays a periodic peak with period equal to the OFDM symbol duration for a carrier frequency offset of 70 subcarrier spacings and an E_b/N_0 of 0dB.

Fig 10. shows the estimated timing error for the coarse acquisition stage with sampling frequency offset of 50ppm of the OFDM sample rate and QPSK and 64-QAM modulations. Simulation results indicate that

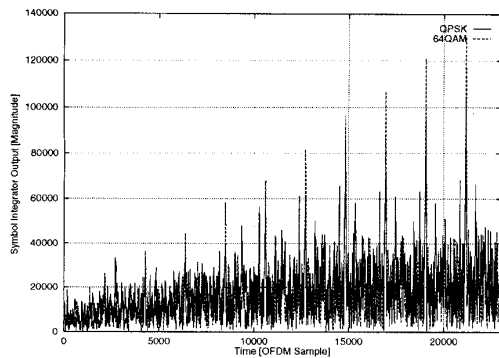


Fig. 9. Sliding integrator output: QPSK, 64-QAM, multipath channel II, $E_b/N_0=0\text{dB}$, carrier frequency offset=70 subcarrier spacings, FFT size=2048, decimation order=8, guard time=1/32 useful symbol duration.

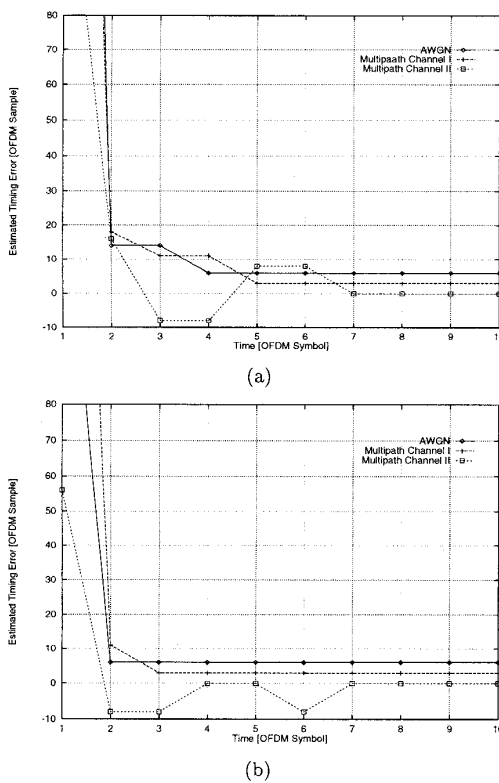


Fig. 10. Estimated error in the coarse acquisition stage: (a) QPSK and (b) 64-QAM. AWGN channel, multipath channels I and II, $E_b/N_0=0\text{dB}$, carrier frequency offset=70 subcarrier spacings, FFT size=2048, decimation order=8, guard time=1/32 useful symbol duration.

in spite of employing the decimator with a decimation order of 8, the estimated error decreases to within approximately ± 10 OFDM samples well within 10 OFDM symbol durations. From these figures, we note that the proposed coarse acquisition algorithm gives a symbol timing estimate corresponding to the center of mass of the channel energy delay profile. The small offset observed after convergence is due to the decimation and is not a bias in the algorithm.

B. Fine Symbol Time Tracking

Fig. 11 shows the characteristic S-curve of the described phase detector with QPSK and 64-QAM data modulation. For all cases, the phase detector characteristics are approximately linear to within ± 40 OFDM samples which is much larger than the expected range of the residual symbol timing offset after initial coarse symbol timing acquisition.

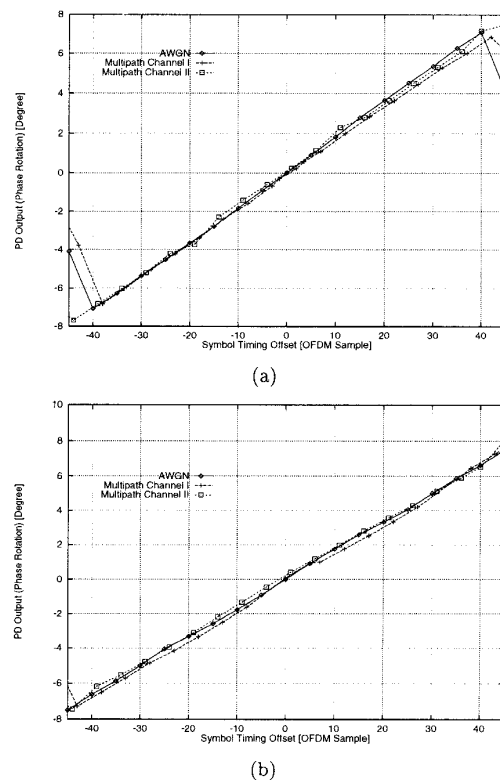


Fig. 11. S-curve of the proposed phase detector: (a) QPSK and (b) 64-QAM. AWGN channel, multipath channels I and II.

From Fig. 10-11, we note that the characteristics of the symbol time tracking algorithm is compatible with

that of the coarse timing acquisition algorithm, i.e., the stable operating point of the symbol time tracking loop coincides with the center of mass of the channel energy delay profile given in Table 2.

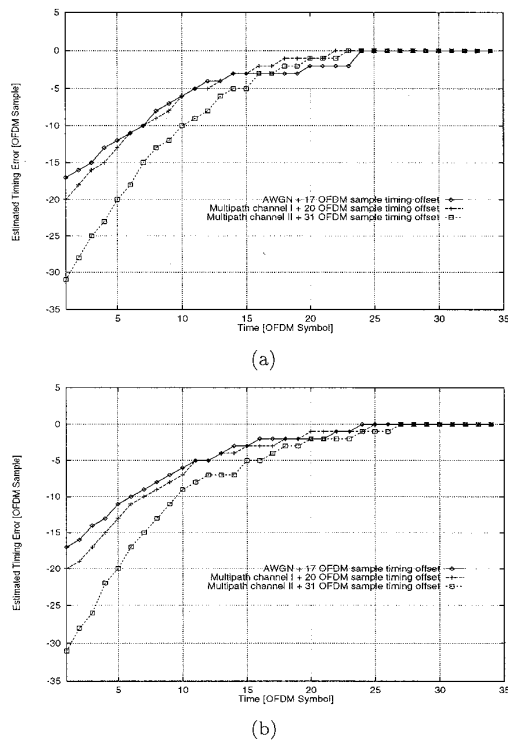


Fig. 12. Transient behavior of the fine tracking stage: (a) QPSK and (b) 64-QAM. AWGN channel and multipath channel I and II, $E_b/N_0=0\text{dB}$, $W=5$, carrier frequency offset=0.05 subcarrier spacing, initial symbol timing offset=17, 20 and 31 OFDM samples, FFT size=2048, number of the useful subcarriers=1705.

Fig. 12 shows the transient responses of a second-order fine tracking loop with the initial residual symbol timing offsets of 17, 20 and 31 OFDM samples. It can be seen from Fig. 12 that accurate symbol timing can be successfully achieved in less than 30 OFDM symbol durations even for E_b/N_0 as low as 0dB.

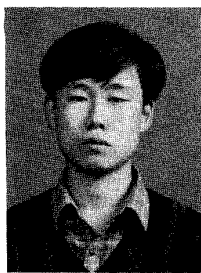
V. CONCLUSIONS

We proposed a novel, patent-pending non data-aided symbol timing recovery algorithm which is insensitive to carrier frequency offsets. A method of jointly achieving coarse symbol timing acquisition and fine carrier frequency synchronization was also proposed. Simu-

lation results show that the described algorithm can recover symbol timing well within 50 OFDM symbol duration at E_b/N_0 as low as 0dB. With proper use of gear-shifting techniques, shorter acquisition times and smaller steady-state timing jitter may be achieved.

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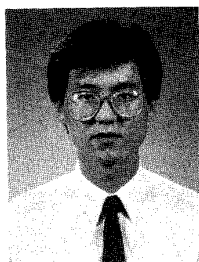
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