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Performance of Pilot-Symbol-Aided 16QAM with N-branch Postdetection Diversity Reception over Multipath Fading Channels

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Abstract – The paper studies the bit-error-rate (BER) performance of a pilot-symbol-aided (PSA) 16QAM system with an *N*-branch postdetection selection combining diversity reception over multipath fading channels. Both pilot symbols and data symbols are used to compensate for the fading distortion. Computer simulation results have shown that the use of diversity reception can significantly improve the BER performance of the PSA-16QAM system. The improvement is more significant at high signal-to-noise ratio (SNR) and at high signal-to-interference ratio (SIR).

I. Introduction

To cope with the bandwidth requirement for wireless personal multimedia applications, spectrally efficient signals (e.g. 16QAM) are expected to be used [1]. When spectrally efficient signals are transmitted over mobile radio channels, multipath fading causes severe signal distortion and hence degradation in system performance. It is known that pilot-symbol-aided (PSA) techniques can be used to compensate for multipath fading distortion [1-8]. In conventional PSA techniques [2-4], the fading estimation processes make use of only the pilot symbols but ignore the fading information in data symbols. A simple and novel PSA technique that makes use of both pilot and data symbols has been proposed and results have shown that substantial improvements on bit-error rate (BER) performance can be achieved [5]. Recently, the use of diversity reception in PSA systems received much attention and some work has been done in this aspect [6-8]. This paper extends the previous work in [5] and studies the effectiveness of incorporating an *N*-branch postdetection selection combining diversity reception technique in a PSA-16QAM system.

II. System Model

The baseband equivalent model of the data-transmission system with diversity reception is shown in Fig. 1. The information to be transmitted is carried by binary digits $\{u_n\}$.

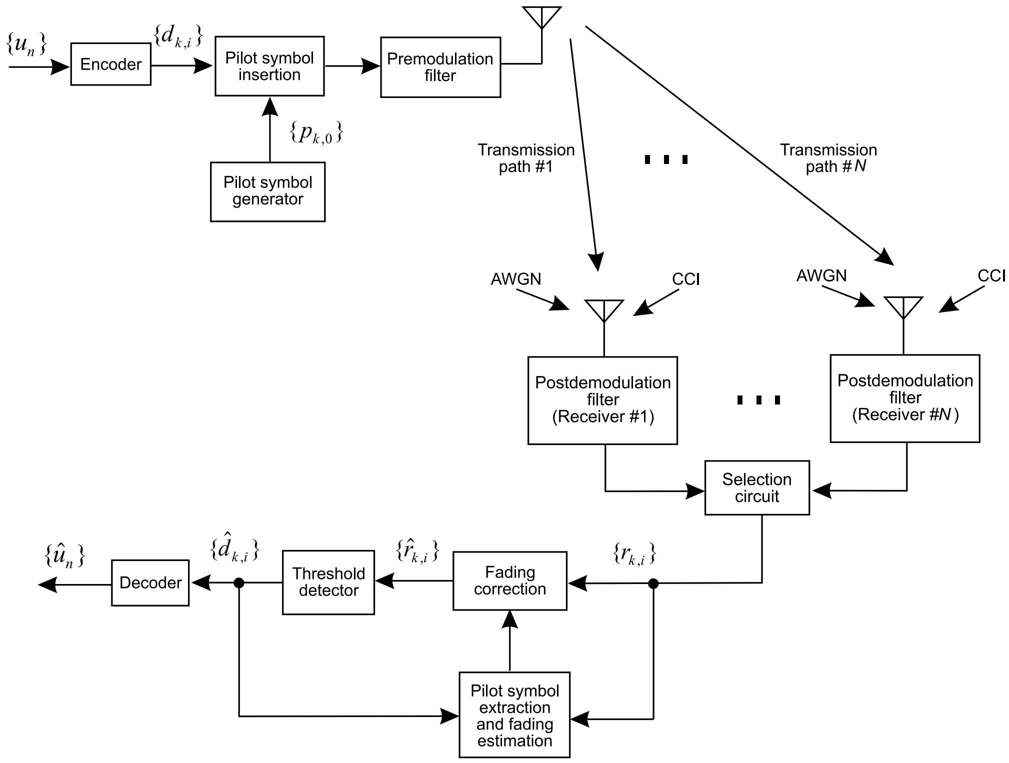

Fig. 1. System model.

When the encoder has received the binary data $\{u_n\}$ at time $t = nT$ seconds (where n is a positive integer and T is the symbol duration), it maps these signals into appropriate data symbol $\{d_{k,i}\}$ according to the 16QAM signal constellation. In the transmitter of Fig. 1, for every (L-1)-data symbols, a pilot symbol from a known pseudorandom-symbol sequence $\{p_{k,0}\}$ is inserted to form a frame of L-symbol long, as shown in Fig. 2. A pseudorandom-symbol sequence of pilot symbols is used to avoid transmitting tones. To minimize the performance degradation due to additive white Gaussian noise (AWGN), the signals in $\{p_{k,0}\}$ are chosen from those signal vectors with the largest energy level in the signal constellation [5].

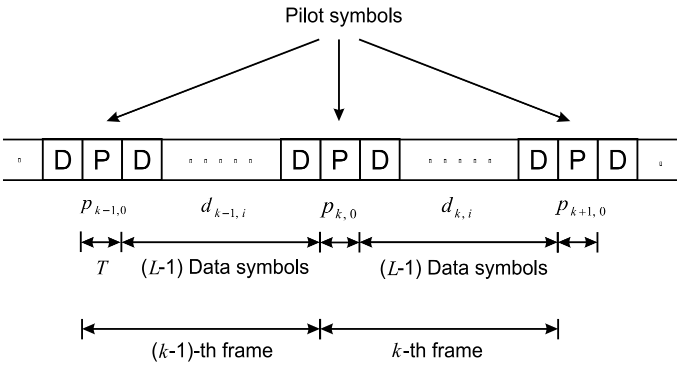

Fig. 2. Frame structure.

At time $t = nT$ seconds, the symbol (either a data symbol or a pilot symbol) to be transmitted is used to form an impulse $q_n \delta(t - nT)$, which is fed to a premodulation filter with an impulse response $a(t)$. The q_n is complex-valued and is either a data symbol $d_{k,i}$ or a pilot symbol $p_{k,0}$, and $\delta(t)$ is the Dirac-delta function. At the premodulation filter output, the signal becomes $\sum_n q_n a(t - nT)$. This signal is then used to linearly modulate a carrier signal to produce the transmit signal. Each of the transmission paths (transmission paths #1, #2, …, and #N) in Fig. 1 introduces a Rayleigh-fading distortion to the corresponding input signal. It is assumed that all the received signals through transmission paths #1, #2, ..., and #N have the same power level and arrived at the receivers at the same time. The distortions in these transmission paths are assumed to be statistically uncorrelated and independent of each other. The N receivers in Fig. 1 account for the N-branch diversity reception system and each path is called a diversity branch. Stationary noise and CCI are assumed to be added at the inputs of the N receivers.

Signals from the N receivers are then filtered by the associated postdemodulation filters which are taken to have the same impulse response $a(t)$ as the premodulation filter in the transmitter. To eliminate the effect of ISI due to the premodulation and postdemodulation filters, the resultant transfer function of the two filters in cascade has a sinusoidal roll-off of 100%. The selection circuit will then select a data frame from one of the N receivers for signal detection. The proposed selection criterion is described in the next section.

III. Selection Method of Diversity Reception

It is assumed here that frame synchronization of the received signals from the N receivers has been achieved. At the output of the postdemodulation filter in receiver #J, where $J = 1$, 2, ..., or N, the sample signal at the i-th position of the k-th received frame can be written as

$$r_{J,k,i} = q_{k,i}\, y_{J,k,i} + c_{J,k,i} + w_{J,k,i} \tag{1}$$

where $q_{k,i}$ is either a pilot symbol or a data symbol, $y_{J,k,i}$ is the fading distortion, $c_{J,k,i}$ is the sample of CCI, and $w_{J,k,i}$ is the sample of noise. For $i = 0$, the signal is

$$r_{J,k,0} = p_{k,0}\, y_{J,k,0} + c_{J,k,0} + w_{J,k,0} \tag{2}$$

where $p_{k,0}$ is the pilot symbol in the k-th frame. For $i = 1, 2, \ldots, L-1$, the signal is

$$r_{J,k,i} = d_{k,i}\, y_{J,k,i} + c_{J,k,i} + w_{J,k,i} \tag{3}$$

where $d_{k,i}$ is the data symbol at the i-th position of the k-th frame. Since the pilot symbol $p_{k,0}$ is known at the receivers, $y_{J,k,0}$ for receiver #J can be obtained, from (2), as

$$y_{J,k,0} = \frac{r_{J,k,0}}{p_{k,0}} - \frac{c_{J,k,0}}{p_{k,0}} - \frac{w_{J,k,0}}{p_{k,0}} \tag{4}$$

In low noise and low CCI environments, $y_{J,k,0}$ and $y_{J,k+1,0}$ can be estimated, respectively, as

$$\hat{y}_{J,k,0} = \frac{r_{J,k,0}}{p_{k,0}} \tag{5a}$$

$$\hat{y}_{J,k+1,0} = \frac{r_{J,k+1,0}}{p_{k+1,0}} \tag{5b}$$

and are used to select the best data-frame from the N receivers. The proposed selection method depends on the estimated fading distortions of the two pilot symbols (both in-phase, quadrature components, $\left|\mathrm{Re}\left(\hat{y}_{J,k,0} + \hat{y}_{J,k+1,0}\right)\right|$ and $\left|\mathrm{Im}\left(\hat{y}_{J,k,0} + \hat{y}_{J,k+1,0}\right)\right|$) in the k-th and (k+1)-th frame. Thus, for signals in the k-th frames, the selection circuit computes the values of

$$H_J = \left|\, \mathrm{Re}\left(\hat{y}_{J,k,0} + \hat{y}_{J,k+1,0}\right)\right| + \left|\, \mathrm{Im}\left(\hat{y}_{J,k,0} + \hat{y}_{J,k+1,0}\right)\right| \tag{6}$$

for $J = 1$, 2, …, and N, using the two pilot symbols in the associated k-th and (k+1)-th frames. The circuit selects the k-th data frame associated with the largest value of H_J from these N receivers. If more than one receiver have the same value of H_J, the selection circuit selects any one of them. Once a receiver is selected for signal detection, the fading compensation process of the k-th frame from that receiver begins.

IV. Multipath Fading Compensation

Assume the signal from receiver #M has been selected for detection. At the postdemodulation filter output of receiver #M, the baseband signal is

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$$r_M(t) = \sum_n \left[q_n a(t-nT) y_M(t) \right] * a(t) + c_M(t) + w_M(t) \qquad (7)$$

where $y_M(t)$ is the combined fading distortion introduced in transmission path #*M*, $c_M(t)$ and $w_M(t)$ are the filtered CCI and the filtered noise waveforms with a single-sided power spectrum of N_0, at receiver #*M*, respectively. When the baseband signal $r(t)$ is sampled in synchronism at time instants $\{nT\}$ and $a(t) * a(t) = 1$ at time $t = 0$, signal at time $t = nT$ is given by

$$r_{M,n} = q_n y_{M,n} + c_{M,n} + w_{M,n} \qquad (8)$$

where $y_{M,n} = y_M(nT)$, $c_{M,n} = c_M(nT)$, and $w_{M,n} = w_M(nT)$.

To simplify the notations, the sample signal at the *i*-th position of the *k*-th received frame (from receiver #*M*) can be written as

$$r_{k,i} = q_{k,i} y_{k,i} + c_{k,i} + w_{k,i} \qquad (9a)$$

where $q_{k,i}$ is either a pilot symbol or a data symbol, $y_{k,i}$ and $w_{k,i}$ are the change, the CCI and the noise sample at the *i*-th symbol of the *k*-th frame, respectively. For *i* = 0, the sample signal is

$$r_{k,0} = p_{k,0} y_{k,0} + c_{k,i} + w_{k,0} \qquad (9b)$$

where $p_{k,0}$ is the pilot symbol in the *k*-th frame. For *i* = 1, 2, ..., *L*-1, the sample signal is

$$r_{k,i} = d_{k,i} y_{k,i} + c_{k,i} + w_{k,i} \qquad (9c)$$

where $d_{k,i}$ is a data symbol at the *i*-th symbol of the *k*-th frame. When $d_{k,i} y_{k,i} >> c_{k,i} + w_{k,i}$, $y_{k,0}$ and $y_{k+1,0}$ can be estimated as

$$\hat{y}_{k,0} = \frac{r_{k,0}}{p_{k,0}} \qquad (10a)$$

$$\hat{y}_{k+1,0} = \frac{r_{k+1,0}}{p_{k+1,0}} \qquad (10b)$$

The compensation process is suitable for frame lengths of 4, 8, 16, ..., or 2^m, where *m* is an integer, and consists of two stages. The first stage works on the data symbols in the even-number positions (i.e., *i* = 2, 4, ..., *L*-2), while the second stage works on the data symbols in the odd-number positions (i.e., *i* = 1, 3, ..., *L*-1) of a frame.

Compensation in Even-Number Positions

The changes on the pilot symbols due to fading distortion in the *k*-th frame, $\hat{y}_{k,0}$, and in the (*k*+1)-th frame, $\hat{y}_{k+1,0}$, can be obtained using (10). The estimate of $y_{k,L/2}$ is obtained as

$$\tilde{y}_{k,L/2} = \frac{\hat{y}_{k,0} + \hat{y}_{k+1,0}}{2} \qquad (11a)$$

This signal is then used to correct the fading effects in $r_{k,L/2}$ to give an estimate of the data signal

$$\hat{r}_{k,L/2} = \frac{r_{k,L/2}}{\tilde{y}_{k,L/2}} \qquad (11b)$$

which is threshold detected to produce the data symbol $\hat{d}_{k,L/2}$.

A more accurate fading distortion in $r_{k,L/2}$ is then given by $r_{k,L/2} / \hat{d}_{k,L/2}$. If *L* = 4, the compensation process for the data symbols in the even-number positions is completed, and the compensation process for the data symbols in the odd-number positions begins. However, if $L = 8$, the compensation process for the data symbols in the even-number positions continues as follows. Since the detected data symbol $\hat{d}_{k,L/2}$ is a possible signal vector on the constellation and the estimate of $y_{k,L/2}$ using $r_{k,L/2} / \hat{d}_{k,L/2}$ is closer to $y_{k,L/4}$ and $y_{k,3L/4}$ than $\hat{y}_{k,0}$ and $\hat{y}_{k+1,0}$, respectively, in terms of time. In a slowly faded channel, better estimates of $y_{k,L/4}$ and $y_{k,3L/4}$ can then be obtained as, respectively,

$$\tilde{y}_{k,L/4} = \frac{1}{2}\left(\hat{y}_{k,0} + \frac{r_{k,L/2}}{\hat{d}_{k,L/2}} \right) \qquad (12a)$$

$$\tilde{y}_{k,3L/4} = \frac{1}{2}\left(\frac{r_{k,L/2}}{\hat{d}_{k,L/2}} + \hat{y}_{k+1,0} \right) \qquad (12b)$$

These signals $\tilde{y}_{k,L/4}$ and $\tilde{y}_{k,3L/4}$ are used to correct $r_{k,L/4}$ and $r_{k,3L/4}$, respectively, to obtain the signals

$$\hat{r}_{k,L/4} = \frac{r_{k,L/4}}{\tilde{y}_{k,L/4}} \qquad (12c)$$

$$\hat{r}_{k,3L/4} = \frac{r_{k,3L/4}}{\tilde{y}_{k,3L/4}} \qquad (12d)$$

which are used to obtain the detected data symbols $\hat{d}_{k,L/4}$ and $\hat{d}_{k,3L/4}$. For *L* = 16, 32, ..., 2^n, similar equations can be derived. The compensation process continues until all the data symbols in the even-number positions are done. Then the compensation process for the data symbols in the odd-number positions begins.

Compensation in Odd-Number Positions

For compensation of the data symbols in the odd-number positions, $\tilde{y}_{k,i}$ and $\hat{r}_{k,i}$ are obtained as

$$\tilde{y}_{k,1} = \frac{1}{2}\left(\hat{y}_{k,0} + \frac{r_{k,2}}{\hat{d}_{k,2}} \right) \qquad (13a)$$

$$\tilde{y}_{k,L-1} = \frac{1}{2}\left(\frac{r_{k,L-2}}{\hat{d}_{k,L-2}} + \hat{y}_{k+1,0} \right) \qquad (13b)$$

$$\tilde{y}_{k,j} = \frac{1}{2}\left(\frac{r_{k,j-1}}{\hat{d}_{k,j-1}} + \frac{r_{k,j+1}}{\hat{d}_{k,j+1}} \right) \quad \text{for } j = 3, 5, ..., L-3 \qquad (13c)$$

$$\hat{r}_{k,i} = \frac{r_{k,i}}{\tilde{y}_{k,i}} \quad \text{for } i = 1, 3, ..., L-1 \qquad (13d)$$

All the corrected signal samples $\{\hat{r}_{k,i}\}$ are fed to the threshold detector to produce $\{\hat{d}_{k,i}\}$ which are finally decoded into the binary data $\{\hat{u}_i\}$. Every time, when a data frame is selected by the selection circuit from one of the receivers, the whole compensation process repeats.

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The major advantages of the proposed selection diversity reception technique are simple implementation and only simple signal processing is required at the receiver. These features are particularly essential to many real-time and delay-sensitive applications, such as voice communications, in the mobile environments.

V. BER Performance of PSA-16QAM

Basic Assumptions

Intensive computer-simulation tests have been carried out to investigate the effect of the proposed N-branch diversity reception technique, with $N = 1$, 2, 3, and 4, on the BER performance of PSA-16QAM over multipath-fading channels. Throughout the tests, a normalized Doppler spread, $f_D T$, of 0.005 (which corresponds to a carrier frequency of 900 MHz at a relative velocity between the transmitter and receiver of 48 km/hr), a transmission rate of 32 kbps and a frame length of $L = 8$ are used. All results presented are already taken into account the signal energy and system throughput degradations due to transmitting the pilot symbols [4]. The average signal-to-noise ratio (SNR) is defined as

$$\text{SNR} = 10 \log \left(E_b / N_0\right) \quad \text{dB} \tag{14}$$

where E_b is the average transmitted bit energy (after taken into account the energy loss due to transmitting the pilot symbols) and N_0 is the single-sided power spectral density of AWGN. The average signal-to-interference ratio (SIR) is defined as

$$\text{SIR} = 10 \log \left(S / I\right) \quad \text{dB} \tag{15}$$

where S is the average signal power and I is the average power of CCI. In the frequency-selective fading channels, the power ratio of the main-path signal to delayed-path signal is defined as

$$\text{CDR} = 10 \log \left(C / D\right) \quad \text{dB} \tag{16}$$

where C and D are the main-path signal power and delayed-path signal power, respectively. The normalized time-delay of the delayed-path signal, relative to the main-path signal is defined as τ / T, where τ is the time-delay and T is the symbol duration.

BER Performance

In the frequency-nonselective Rayleigh-fading channel (i.e. $\text{CDR} = \infty$ dB) and absence of CCI (i.e. $\text{SIR} = \infty$ dB), the BER performance of the PSA-16QAM against SNR is shown in Fig. 3. When SNR = 20 dB, the use of the diversity reception technique improves the BERs by factors of 10.89 (from 9.8×10^{-3} to 9.0×10^{-4}), 98 (from 9.8×10^{-3} to 1.0×10^{-4}) and 408.33 (from 9.8×10^{-3} to 2.4×10^{-5}), for $N = 2$, 3 and 4, respectively, relative to that of the non-diversity reception system ($N = 1$ with a BER of 9.8×10^{-3}). The improvement is more significant for higher SNR. When SNR = 30 dB, the diversity improvement for $N = 2$ is increased to 83.33 (from 1.0×10^{-3} to 1.2×10^{-5}).

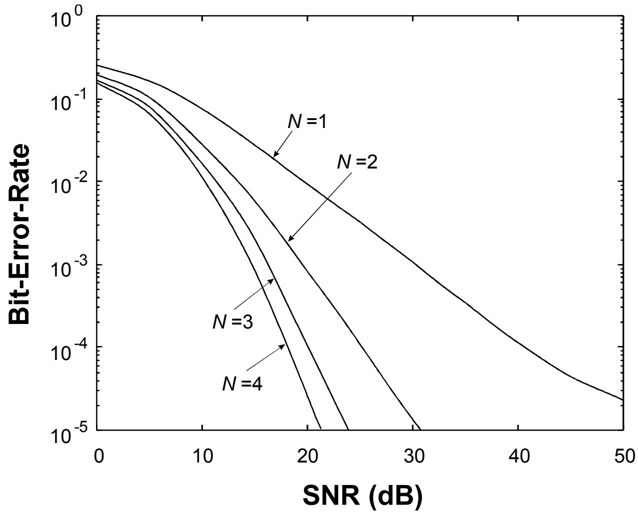

Fig. 3. BER performance against SNR.

In the absence of noise, the BER performance of PSA-16QAM against SIR is shown in Fig. 4. When SIR = 20 dB, the use of the diversity reception technique improves the BERs by factors of 3.7 (from 3.0×10^{-2} to 8.1×10^{-3}), 7.5 (from 3.0×10^{-2} to 4.0×10^{-3}) and 15 (from 3.0×10^{-2} to 2.0×10^{-3}) for $N = 2$, 3, and 4, respectively, relative to that obtained by the non-diversity reception system. The improvement is more significant for higher SIR.

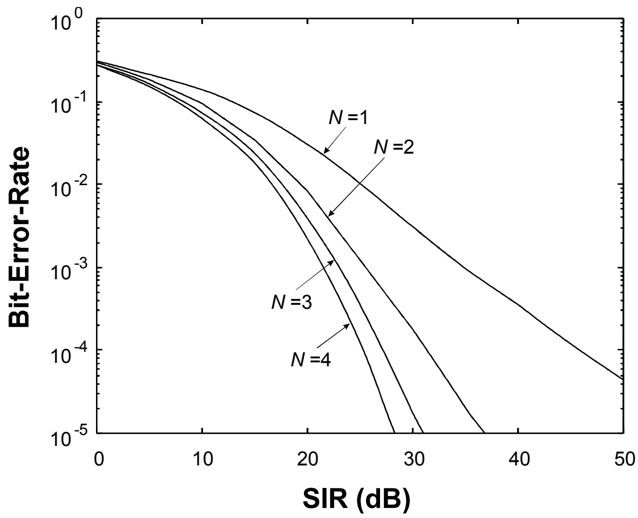

Fig. 4. BER performance against SIR.

In the absence of both noise and CCI, the BER floor performance of PSA-16QAM against CDR in the frequency-selective Rayleigh-fading channels, with normalized delays of $\tau / T = 0.125$ and 0.5, is shown in Fig. 5. It can be seen that the diversity reception technique is more effective for shorter normalized delay. When $\tau / T = 0.125$

and CDR = 0 dB, the 2-branch diversity reception system reduces the BER floor by a factor of 15.48 (from 4.8×10^{-3} to 3.1×10^{-4}), relative to that of the non-diversity reception system. The reduction is less significant for $\tau/T = 0.5$ and CDR = 0 dB, thus the BER floors are sensitive to the normalized delay.

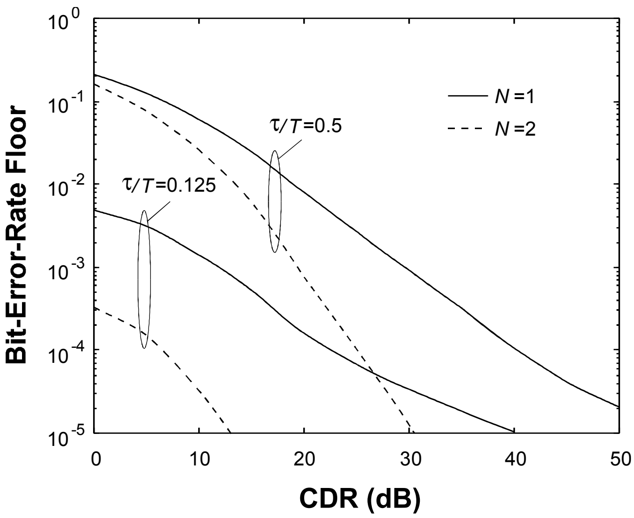

Fig. 5. BER floor performance against CDR.

The BER floor performance of PSA-16QAM against normalized delay, τ/T, in the absence of noise and CCI, when CDR = 3 dB and 10 dB, is shown in Figs. 6a and 6b, respectively. It is evident again that the diversity reception technique is more effective for shorter normalized delay and higher CDR. When $\tau/T = 0.125$ and CDR = 3 dB, the diversity reception system improves the BER floors by factors of 17.7 (from 3.9×10^{-3} to 2.2×10^{-4}) and 150 (from 3.9×10^{-3} to 2.6×10^{-5}), for $N = 2$ and 3, respectively. When CDR = 10 dB and $N = 2$, the improvement becomes 40 (from 1.4×10^{-3} to 3.5×10^{-5}). As the normalized delay increases, the diversity improvement reduces.

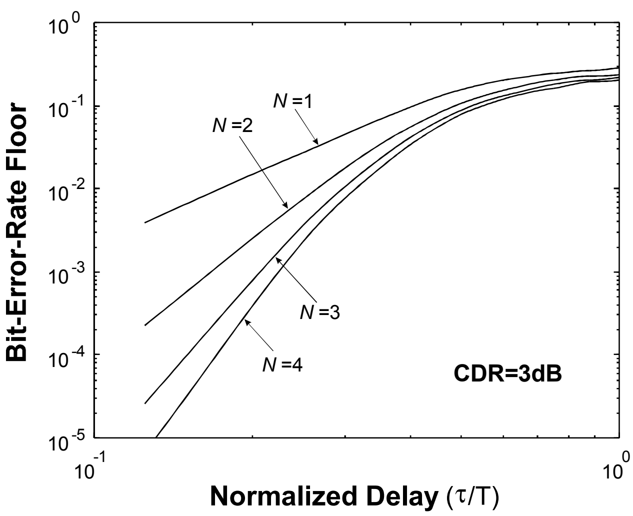

Fig. 6a. BER floor performance against τ/T when CDR = 3 dB.

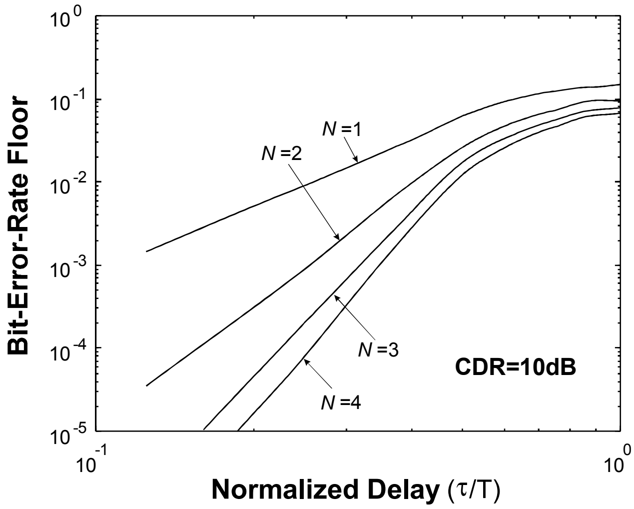

Fig. 6b. BER floor performance against τ/T when CDR = 10 dB.

VI. Summary and Conclusions

The paper studies the effectiveness of incorporating an N-branch postdetection selection combining diversity reception technique, with N = 1, 2, 3 and 4, on the BER performance of a PSA-16QAM system over multipath fading channels. The diversity improvement is found more significant for higher SNR, higher SIR, lower BER, shorter normalized delay and higher CDR. The major advantage of the proposed diversity reception technique is the simple signal-processing requirement, which are essential for wireless personal multimedia applications.

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