Phase-combining diversity using adaptive decision-aided branch-weight estimation for reception of faded M-ary DPSK signals

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Phase-Combining Diversity Using Adaptive Decision-Aided Branch-Weight Estimation for Reception of Faded M-ary DPSK Signals

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***Abstract*—A practical adaptive phase-combining (PC) diversity-reception scheme based on approximate maximum-likelihood (ML) decision is proposed for M-ary differentially encoded phase-shift keying (MDPSK) with differential phase detection (DPD). The approximate ML decision chooses the data symbol that minimizes the weighted sum of the squared phase errors of L DPD detector ouptuts, where L is the number of diversity branches. The branch weights are adaptively estimated, based on feeding back past data decisions, from each branch DPD detector output-phase sequence. Adaptive PC diversity utilizes L DPD detector output-phase sequences only and requires no measurement function of the received signal instantaneous powers of the diversity branches. The average bit-error rate (BER) performance in the presence of additive white Gaussian noise (AWGN), Doppler spread, and multipath channel delay spread is evaluated by computer simulations for 4DPSK signal transmission in Rayleigh fading channels.**

***Index Terms*—Adaptive estimation, diversity reception, MDPSK.**

I. Introduction

M-ary differential phase-shift keying (MDPSK) with differential detection has attracted increased attention in the field of mobile radio, where fast tracking, yet accurate carrier recovery, is very difficult because of multipath fading. The causes of decision errors in fading channels are additive white Gaussian noise (AWGN), fading Doppler spread, and multipath channel delay spread [1]. The delay spread places a limit on the achievable maximum transmission bit rate. Diversity reception [1] can be used to combat the effects of fading. Predetection diversity, which coherently combines the received faded signals before detection, may be difficult to implement because of the fast random phase variations of faded signals (this is the same reason for preferring differential detection), so postdetection diversity [2]–[6] is preferable. Recently, postdetection phase-combining (PC) diversity suitable for the differential phase detection (DPD) of MDPSK signals was proposed [7]: the weighted sum of the DPD detector output phases is the decision variable and the branch weights are computed using the measured instantaneous received signal powers. To avoid an incorrect phase sum arising from the 2π periodicity of DPD detector output phases, phase correction is needed before combination [7, (7)]. This limits its application to the two-branch case only. More recently, an extension to the multiple branch case $(L > 2)$ by performing weighted summation of the squared phase errors of DPD detector outputs was proposed [8]. However, the above-mentioned diversity schemes [7], [8] require the measurement of the received signal instantaneous powers of the diversity branches and, thus, need careful adjustment of receiver gains of all branches.

This paper is a followup study of the previous paper [7] in that it extends the number of diversity branches to any arbitrary number and removes the need to measure the instantaneous received signal powers at the receiver. Adopting a Gaussian phase-noise assumption of DPD detector output, a simple adaptive PC diversity scheme, based on approximate maximum-likelihood (ML) decision using L DPD detector output-phase sequences, is developed for the reception of MDPSK signals transmitted over fading channels, where L is the number of diversity branches. The decision variable is the weighted sum of the squared phase errors of L DPD detectors. No power-measurement function is required for computing the branch weights. Branch weights are adaptively estimated from each branch DPD detector output-phase sequence based on feeding back the past decisions. The proposed adaptive PC diversity reception is described in Section II. In Section III, the effects of the proposed scheme on 4DPSK transmission under Rayleigh fading are evaluated by computer simulations, and the resulting bit-error rate (BER) performance in the presence of AWGN, Doppler spread, and delay spread is presented.

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II. Adaptive PC Diversity Using Adaptive Decision-Aided Branch-Weight Estimation

We assume here the very slow multiplicative fading process (i.e., the delay spread is assumed to be very small compared to the symbol duration so that its effect is negligible) to simplify the explanation of the proposed diversity. However, the effect of delay spread is considered in the computer simulations described in Section III.

A. DPD Detector-Output Representation

The transmitted $\log_2 M$-bit symbol is mapped to the phase $\Delta\phi_n \in \{2m\pi/M\}_{m=0}^{M-1}$. The MDPSK signal to be transmitted can be represented in the complex form as

$$s(t) = \sqrt{2(\log_2 M)E_b/T} \sum_{n=-\infty}^{\infty} p(t-nT)\exp j\phi_n \qquad (1)$$

where $\phi_n = \phi_{n-1} + \Delta\phi_n$ is the carrier phase, E_b is the signal energy per bit, T is the symbol duration, and $p(t)$ is the low-pass equivalent impulse response of the transmit filter (we assume a square-root Nyquist filter). The MDPSK signal is transmitted over the Rayleigh fading channel and is received by L antennas (L is an arbitrary positive integer). We assume perfect automatic frequency control such that there is no frequency offset between the transmitter and receiver and perfect sampling timing. The received signal is perturbed by the AWGN with single-sided power-spectrum density N_0 and is bandlimited by the receive matched filter with low-pass equivalent impulse response $q(t) = (1/T)p(-t)$. The receive filter output of the lth branch can be expressed as $r_l(t) = \sqrt{2(\log_2 M)E_b/T}\xi_l(t)\sum_{n=-\infty}^{\infty} h(t-nT)\exp j\phi_n + w_l(t)$, where $h(t) = p(t) \otimes q(t)$ is the overall (transmit plus receive) filter response (here, $\otimes$ is the convolution operation), $\xi_l(t)$ is the fading complex envelope with unity power, and $w_l(t)$ is the filtered AWGN component. After amplitude limiting, the phase of the filtered signal $r_l(t)$ is detected by the phase detector and sampled at $t = nT$: the sampled phase is denoted by $\psi_{l,n}$. Since we are assuming that the overall filter response yields an intersymbol-interference (ISI)-free Nyquist-filter response, i.e., $h(nT) = 1(0)$ for $n = 0$ (otherwise), $\psi_{l,n}$ is the phase of

$$\begin{aligned} r_{l,n} &= r_l(nT) = \sqrt{2(\log_2 M)E_b/T}\xi_{l,n}\exp j\phi_n + w_{l,n} \\ &= |r_{l,n}|\exp j\psi_{l,n} \end{aligned} \qquad (2)$$

where E_b is now the average received signal energy per bit equal on all branches and $w_{l,n} = w_l(nT)$ is the uncorrelated zero-mean complex Gaussian noise samples with variance $2N_0/T$. Denoting the fading-induced random phase noise and AWGN-induced phase noise by $\theta_{l,n}$ and $\eta_{l,n}$, respectively, $\psi_{l,n}$ can be expressed as $\psi_{l,n} = (\phi_n + \theta_{l,n} + \eta_{l,n}) \bmod 2\pi$. The detected phase $\psi_{l,n}$ fluctuates around $\phi_n + \theta_{l,n}$ due to AWGN. The DPD detector produces the phase difference of successive two phases

$$\Delta\psi_{l,n} = (\psi_{l,n} - \psi_{l,n-1}) \bmod 2\pi. \quad (3)$$

At this stage, the unknown fading-induced random phase is removed in the case of very slow fading since $\theta_{l,n} - \theta_{l,n-1} \approx 0$.

B. Approximate ML Decision

From [9, Appendix], $\eta_{l,n}$'s can be approximated as uncorrelated zero-mean Gaussian variables with the same variance $1/2[|\xi_{l,n}|^2 E_s/N_0]^{-1}$, where $E_s = (\log_2 M)E_b$ is the average received signal energy per symbol. Therefore, it can be understood from (3) that $\Delta\psi_{l,n}$'s, $l = 1, 2, \cdots, L$, can be approximated as uncorrelated Gaussian variables with mean $\Delta\phi_n$ and variance $\sigma_{l,n}^2 = [|\xi_{l,n}|^2 E_s/N_0]^{-1}$. Adopting this Gaussian noise assumption of DPD detector output, the joint probability density function (pdf) of $\Delta\psi_{l,n}$'s can be expressed as

$$p(\Delta\psi_{1,n}, \cdots, \Delta\psi_{L,n} \mid \Delta\phi_n) = \prod_{l=1}^{L} \frac{1}{\sqrt{2\pi}\sigma_{l,n}} \exp\left[-\frac{[(\Delta\psi_{l,n} - \Delta\phi_n) \bmod 2\pi]^2}{2\sigma_{l,n}^2}\right]. \quad (4)$$

An approximate ML decision chooses the data symbol that maximizes (4) or equivalently chooses the data symbol that minimizes the weighted sum of squared phase errors of L DPD detector outputs

$$\min_{\text{over } \Delta\phi_n} \sum_{l=1}^{L} \frac{[(\Delta\psi_{l,n} - \Delta\phi_n) \bmod 2\pi]^2}{\sigma_{l,n}^2} \Rightarrow \Delta\tilde{\phi}_n. \quad (5)$$

The received $\log_2 M$-bit symbol is recovered from $\Delta\tilde{\phi}_n$.

C. Adaptive Estimation of $\sigma_{l,n}^2$

The inverse of $\sigma_{l,n}^2$ in (5) can be considered to be the branch weight $g_{l,n}$ associated with the lth branch in diversity combining. The value of $\sigma_{l,n}^2$ under very slow fading is given by $\sigma_{l,n}^2 = [|\xi_{l,n}|^2 E_s/N_0]^{-1}$ from Section II-B, so $g_{l,n} = |\xi_{l,n}|^2 E_s/N_0$. The previously proposed diversity combining scheme [8] directly measures the instantaneous received signal power to obtain $g_{l,n} = |\xi_{l,n}|^2 E_s/N_0$: because of weigh computation based on the power measurement, the careful adjustment of receiver gains among all branches is necessary, otherwise, performance degrades due to mismatch of the branch weights. Here, we want to remove the power-measurement function completely from the receiver and adaptively estimate the value of $\sigma_{l,n}^2$ from the DPD detector output sequence $(\Delta\psi_{l,n-1}, \cdots, \Delta\psi_{l,1})$. In order to make the receiver adaptive in fading channels, the detected symbol sequence $(\Delta\tilde{\phi}_{n-1}, \cdots, \Delta\tilde{\phi}_1)$ is used in a feedback fashion to update the estimate.

The estimate of $\sigma_{l,n}^2$ for the nth decision is denoted as $\tilde{\sigma}_{l,n}^2$. The estimation is based on minimization of the exponentially weighted phase error J_l

$$J_l = \sum_{i=1}^{n-1} \lambda^{i-1} |\Delta\mu_{l,n-i}^2 - \tilde{\sigma}_{l,n}^2|^2 \quad (6)$$

where $\lambda\,(0 < \lambda \le 1)$ is the forgetting factor and

$$\Delta\mu_{l,n-i} = (\Delta\psi_{l,n-i} - \Delta\tilde{\phi}_{n-i}) \bmod 2\pi \quad (7)$$

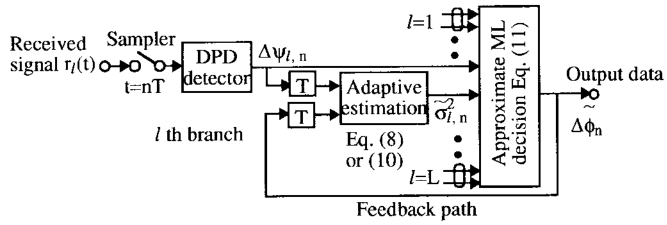


Fig. 1. Diversity receiver structure.

is the phase error at time $n - i$ of the lth branch. Minimization of (6) leads to the following one-tap recursive least square (RLS) estimation [10]:

$$\tilde{\sigma}_{l,n}^2 = (1 - \lambda_n')\Delta\mu_{l,n-1}^2 + \lambda_n'\tilde{\sigma}_{l,n-1}^2 \quad (8)$$

where

$$\lambda_n' = 1 - \frac{1-\lambda}{1-\lambda^n} \quad \text{and} \quad \tilde{\sigma}_{l,0}^2 = \text{small positive value}. \quad (9)$$

As time n elapses, λ_n' rapidly approaches λ if the value of λ is not too close to unity. We may replace λ_n' with λ with a slight sacrifice in convergence speed. In this case, we obtain the one-tap least mean-square (LMS) estimation [10] with step-size λ

$$\tilde{\sigma}_{l,n}^2 = \Delta\mu_{l,n-1}^2 + \lambda\left(\tilde{\sigma}_{l,n-1}^2 - \Delta\mu_{l,n-1}^2\right). \quad (10)$$

Using the estimation algorithm given by (8) or (10), we obtain the following approximate ML decision:

$$\min_{\text{over } \Delta\phi_n} \sum_{l=1}^{L} \frac{[(\Delta\psi_{l,n} - \Delta\phi_n) \bmod 2\pi]^2}{\tilde{\sigma}_{l,n}^2} \Rightarrow \Delta\tilde{\phi}_n. \quad (11)$$

The receiver structure of the proposed adaptive PC diversity reception scheme is illustrated in Fig. 1.

III. Computer Simulations

Theoretical BER analysis is difficult because the exact statistical properties of the weighted sum of phase noises are unknown. Here, by means of computer simulations, we investigate the BER performance achieved by adaptive PC diversity and compare the simulated performance with the theoretical performance of optimal postdetection diversity [2]. Assuming 4DPSK $(M = 4)$, we evaluate the BER performance due to AWGN, Doppler spread, and multipath channel delay spread.

A. Transmission System Model

The modulated desired signal is assumed to be transmitted over Rayleigh fading channels. Frequency selectivity of the channel is determined by the rms delay spread τ_{rms} normalized by the symbol duration T. Since the power delay profile shape is of no importance for channels with $\tau_{\text{rms}}/T < 0.2$–$0.3$ [2], the double-spike delay profile (or two-path model) with equal average power is assumed, each path being subjected to independent Rayleigh fading. When the time difference between the two paths is τ s, the rms delay spread τ_{rms} is given by $\tau_{\text{rms}} = 0.5\tau$ s. The sample timing was assumed to track the first central moment of the double-spike delay profile as used by the theoretical analysis in [2]. The AWGN-perturbed signal sample of the lth branch can be rewritten as

$$r_{l,n} = 2\sqrt{E_b/T}\left[\xi_{l,n}^{(1)} \sum_{m=-\infty}^{\infty} h(mT + \tau_{\text{rms}}) \exp j\phi_{n-m} + \xi_{l,n}^{(2)} \sum_{m=-\infty}^{\infty} h(mT - \tau_{\text{rms}}) \exp j\phi_{n-m}\right] + w_{l,n} \quad (12)$$

where $\xi_{l,n}^{(1)}$ and $\xi_{l,n}^{(2)}$ are the complex envelopes of the first and second paths, respectively, and are independent complex Gaussian variables with a variance of 0.5. When $\tau_{\rm rms} = 0$, the fading is nonfrequency selective and $\xi_{l,n} = \xi_{l,n}^{(1)} + \xi_{l,n}^{(2)}$: we obtain (2) for square-root transmit and receive Nyquist filters.

The transmitted data is a random binary-data sequence. The Gray-code bit-mapping rule of a two-bit symbol into the differential phase $\Delta\phi_n$ is assumed. The four sets of the independent complex Rayleigh fading envelope sample sequences $\{\xi_{l,n}^{(1)}, \xi_{l,n}^{(2)}; n = 1, 2, \cdots\}_{l=1\text{–}4}$ are generated based on the method described in [11]: we assume that 64 multipaths come from all directions (equally spaced arrival angles) with equal amplitude and that the mobile transceiver travels at a constant speed. The BER due to Doppler spread depends on the normalized maximum Doppler frequency $f_D T$, where f_D = mobile speed/carrier wavelength. In the simulation, we vary the value of $f_D T$. The irreducible BER due to delay spread is affected by the overall filter-response shape [2]. We assume square-root raised-cosine Nyquist filters with the rolloff factor of α as the transmit and receive filters, thus, the overall filter response is given by

$$h(t) = \frac{\sin(\pi t/T)}{\pi t/T} \frac{\cos(\alpha \pi t/T)}{1 - (2\alpha t/T)^2}. \tag{13}$$

As $\alpha \to 1$, the BER due to delay spread reduces. This is because the tails of $h(t)$ decay more quickly as α increases, thereby lessening the ISI caused by delay spread. In the simulations, we used $\alpha = 0.5$ and took into account the ISI from two future and two past symbols, i.e., $|m| \leq 2$ in (12).

B. Results

A sufficient number of frames, each consisting of 128 symbols $\{\Delta\phi_n; n = 1\text{–}128\}$, were transmitted so that at least 40 bit errors were counted. First, we compare the convergence properties of RLS and LMS algorithms for the estimation of $\sigma_{l,n}^2$ in a very slow nonfrequency selective Rayleigh fading channel, i.e., $\xi_{l,n} \approx \xi_{l,0}$ for $n = 1, 2, \cdots, 128$. The initial value $\tilde{\sigma}_{l,0}^2$ of the phase-noise variance was set as 0.01. The value of $\tilde{\sigma}_{l,n}^2$ should approach $\tilde{\sigma}_{l,n}^2 = [|\xi_{l,n}|^2 E_s/N_0]^{-1}$ as time n elapses. Here, the convergence time was evaluated by measuring the BER at time position n (4096 frames were transmitted). The result is plotted in Fig. 2 for $\lambda = 0.6$, $L = 2$, the average $E_b/N_0 = 10$ dB, and $f_D T = 0.001$. In detecting the first symbol $\Delta\phi_1$, proper diversity operation cannot be expected since $\tilde{\sigma}_{l,1}^2$ is the same for all branches, and the average BER with respect to $\Delta\phi_1$ may be close to that with no diversity reception (the simulated average BER without diversity was 4.0×10^{-2} for average $E_b/N_0 = 10$ dB and $f_D T = 0.001$). Both RLS and LMS estimation algorithms provide a quite similar rapid convergence property, and the convergence time is around 20 symbols. Since there is no noticeable difference between the two estimation algorithms, hereafter, we only show the results of the LMS algorithm.

The effect of LMS step-size λ on average BER is shown in Fig. 3 for $f_D T = 0.001$ and 0.01 and average $E_b/N_0 = 15$ dB. When $\lambda = 0$, the branch weights are estimated from the phase error associated with the last decision only. As λ increases, the branch weights become more reliable. However, if λ becomes too close to unity, tracking ability against fading is lost. Hence, there exists the optimum λ. However, a broad optimum range of λ is observed from Fig. 3. $\lambda = 0.6$ is nearly optimum for L = 2–4. The measured average BER's with $\lambda = 0.6$ are plotted as a function of average E_b/N_0 for $f_D T = 0.001$ in Fig. 4(a) and $f_D T = 0.01$ in Fig. 4(b). For comparison, the theoretical performance curves of optimal postdetection diversity [2] are plotted as solid curves. Although the BER performance with adaptive PC diversity is inferior to that with optimal diversity, it can significantly improve the performance compared to the no-diversity case ($L = 1$). When $f_D T = 0.001$, the required E_b/N_0 value at BER $= 10^{-3}$ is 26.5 dB for $L = 1$, and this can be reduced to 16.5 dB by two-branch diversity ($L = 2$), i.e., a diversity gain of 10 dB is obtained. The diversity gain increases with L: 17.8 dB gain is obtained with $L = 4$.

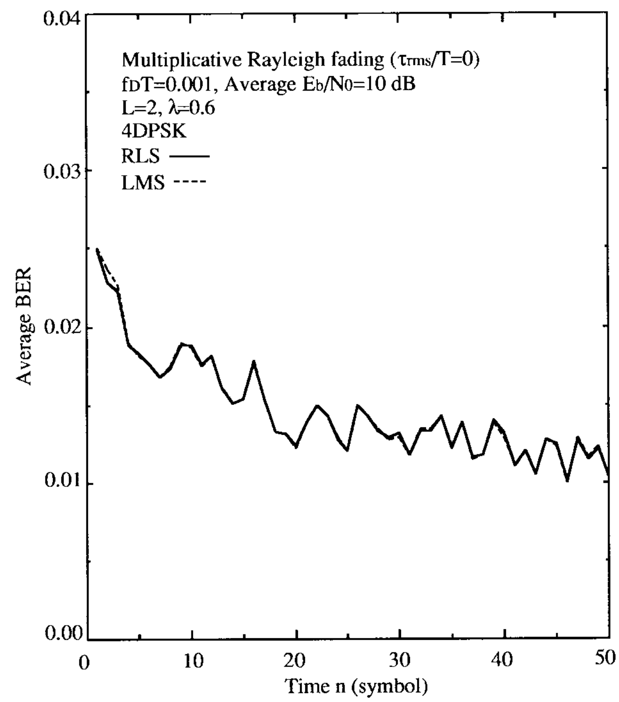


Fig. 2. Comparison of branch-weight convergence with $\lambda = 0.6$, $L = 2$, average $E_b/N_0 = 10$ dB, and $f_D T = 0.001$.

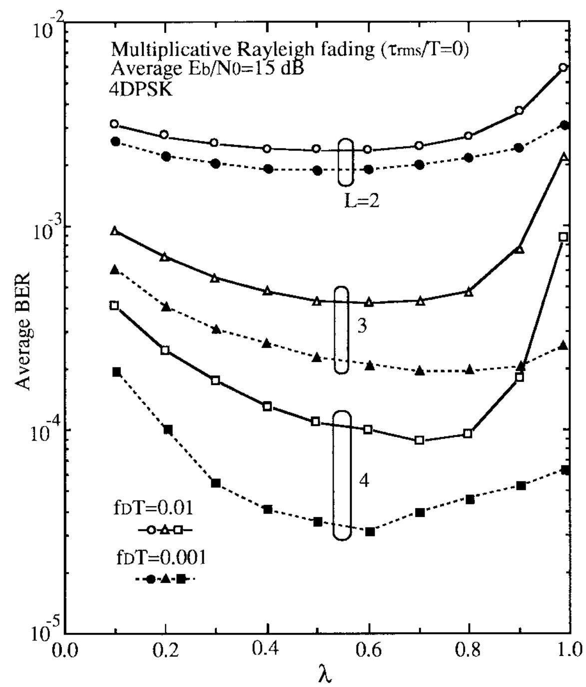
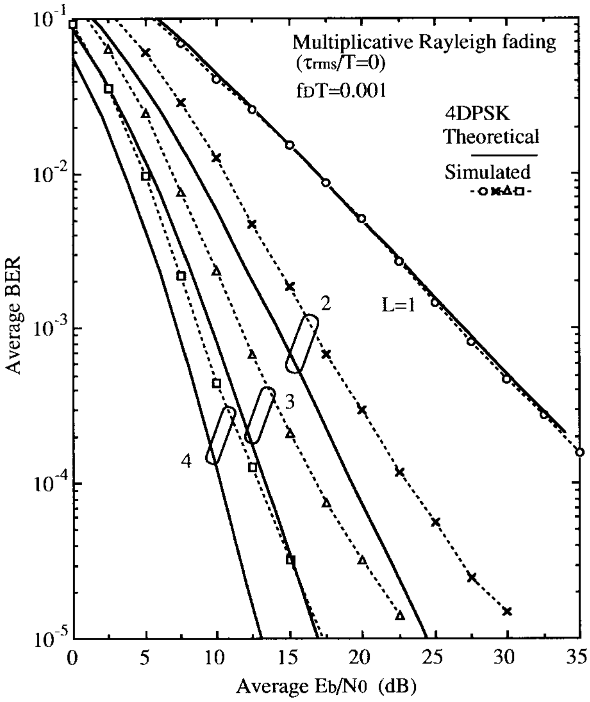
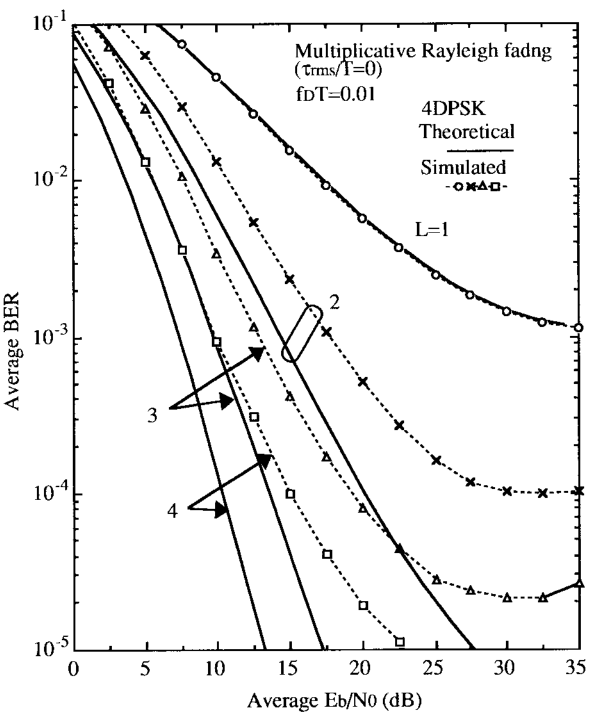


Fig. 3. Effect of LMS step-size λ on average BER due to AWGN with average $E_b/N_0 = 15$ dB and $f_D T = 0.001$ (dotted lines) and 0.01 (solid lines).

The BER performance of adaptive PC diversity is inferior to the theoretically predicted performance of optimal diversity. The



(a)



(b)

Fig. 4. Average BER performance versus average E_b/N_0. (a) $f_DT = 0.001$ and (b) $f_DT = 0.01$.

qualitative reasons for this inferiority are given below. It is understood from (4) that $\sigma_{l,n}^2$ is the ensemble average of squared phase error $\Delta\mu_{l,n}^2 = [(\Delta\psi_{l,n} - \Delta\phi_n) \bmod 2\pi]^2$ and is given by $\sigma_{l,n}^2 = [|\xi_{l,n}|^2 E_s/N_0]^{-1}$, where $\Delta\phi_n$ is the transmitted symbol and $\xi_{l,n}$ is the fading complex envelope at time n. In adaptive PC diversity, $\sigma_{l,n}^2$ is estimated based on the one-tap LMS algorithm with step size λ using the past squared phase-error sequence of $\Delta\mu_{l,i}^2$, $i = 1, 2, \cdots, n-1$, whose ensemble average is given by $\sigma_{l,i}^2 = [|\xi_{l,i}|^2 E_s/N_0]^{-1}$. Notice that $\xi_{l,i} \neq \xi_{l,n}$ and, thus, $\sigma_{l,i}^2 \neq \sigma_{l,n}^2$ in fading environments. A small value of λ, such as $\lambda = 0.6$, is necessary to track the fading variation quickly, however, the insufficiently small number of phase-error samples involved (note that the equivalent number of samples is approximately given by $1/(1-\lambda)$) increases the estimation error. Furthermore, if the same value of λ is used, the performance degrades more as fading becomes faster because the tracking ability against fading tends to be lost. In deriving adaptive PC diversity, the DPD detector output phases are assumed to be Gaussian distributed. However, they are not exactly Gaussian distributed, and this may also have contributed to the performance inferiority.

When average $E_b/N_0 \to \infty$, errors tend to be produced by Doppler spread (or fading-induced random phase variations) and the BER approaches the error floor or irreducible BER. The irredicible BER's with adaptive PC diversity using $\lambda = 0.6$ are much larger than those with optimal diversity [see Fig. 4 (b)], so the use of $\lambda = 0.6$ may not be optimum in large E_b/N_0 regions. Fig. 5 shows the dependence of the irreducible BER on λ for average $E_b/N_0 = 100$ dB and $f_DT = 0.01$. It is seen that λ of around 0.05 is nearly optimum. The qualitative reason for this small optimum value of λ is given below. The fading-induced random phase noise $\theta_{l,n}$ (sometimes called the random FM noise) is large enough to produce decision error only when the fading complex envelope passes close to the origin of the complex plane. This large random phase noise has a time duration much shorter than $1/f_D$ s [12]. In order to track fading-induced random phase noise and thus reduce the BER due to Doppler spread, the observation time interval (approximately given by $1/(1-\lambda)$ symbols) used to estimate the branch weights should be small. The irreducible BER's using $\lambda = 0.05$ are plotted as a function of f_DT for average $E_b/N_0 = 100$ dB in Fig. 6. For comparison, the theoeretical results for optimal postdetection diversity [2] are plotted as solid curves. The simulated BER performances for adaptive PC diversity using $L = 2$–4 lie between the no-diversity case and two-branch optimal diversity $(L = 2)$. However, the use of $L = 2$ reduces the irreducible BER by almost one order of magnitude at $f_DT = 0.01$. Although the irreducible BER can be reduced by increasing L, additional reduction by the use of diversity orders higher than $L = 2$ is small. The possible reason for this is because with $\lambda = 0.05$, the branch weights are estimated mostly from the last detected phase error and the estimation of $\sigma_{l,n}^2$ may not be sufficiently accurate, so proper diversity combining may not be expected. When $L = 2$, however, the tolerable value of f_DT at BER $= 10^{-3}$ is enlarged from 0.01 without diversity $(L = 1)$ to 2.26×10^{-2}. This value corresponds to $f_D = 362$ Hz for 16-k symbol/s transmission.

For high bit-rate transmission, the effect of delay spread becomes predominant at large E_b/N_0 values. The fading is called frequency selective. The measured BER's due to delay spread are plotted in Fig. 7 as a function of normalized rms delay spread $\tau_{\rm rms}/T$ for average $E_b/N_0 = 100$ dB and $f_DT = 0.001$. $\lambda = 0.6$ was used since it was found optimum in the frequency-selective Rayleigh fading. Theoretical performances of optimal diversity [2] are plotted as solid curves. The adaptive PC diversity provides similar performance to optimal diversity, except for small delay spreads. With $L = 1$, the allowable value of $\tau_{\rm rms}/T$ at BER $= 10^{-3}$ is 0.03. The allowable $\tau_{\rm rms}/T$ value increases with L: the use of $L = 2$ (4) can enlarge the allowable $\tau_{\rm rms}/T$ value to 0.09 (0.19), which corresponds to $\tau_{\rm rms} = 1.4$ (3) μs for 64-k symbol/s transmission.

IV. Conclusion

The practical adaptive PC diversity reception of MDPSK signals transmitted over multipath fading channels was described. The sym-

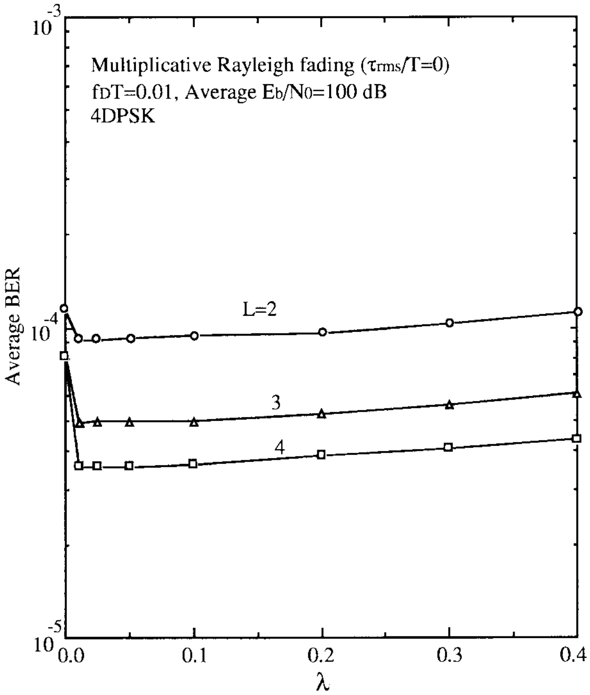


Fig. 5. Effect of LMS step-size λ on irreducible BER due to Doppler spread with average $E_b/N_0 = 100$ dB and $f_D T = 0.01$.

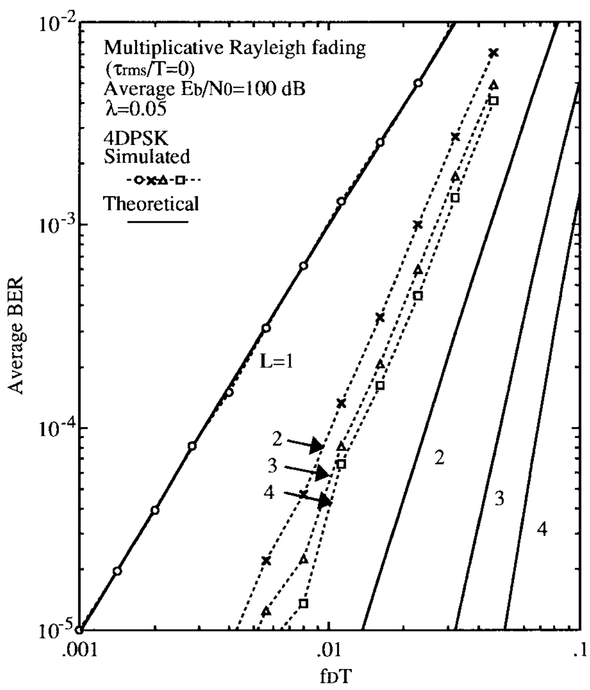


Fig. 6. Irreducible BER versus normalized Doppler spread with $\lambda = 0.05$ and average $E_b/N_0 = 100$ dB.

bol decision of adaptive PC diversity is based on minimizing the weighted sum of the phase errors of L DPD detectors, and the branch weights are adaptively estimated, based on feeding back the past data decisions, from each branch's DPD detector output-phase sequence. Since the proposed diversity scheme does not require the envelope measurement function, it is considered more practical than previously reported diversity schemes [6]–[8]. Computer simulations evaluated the average BER performance in the presence of AWGN, Doppler

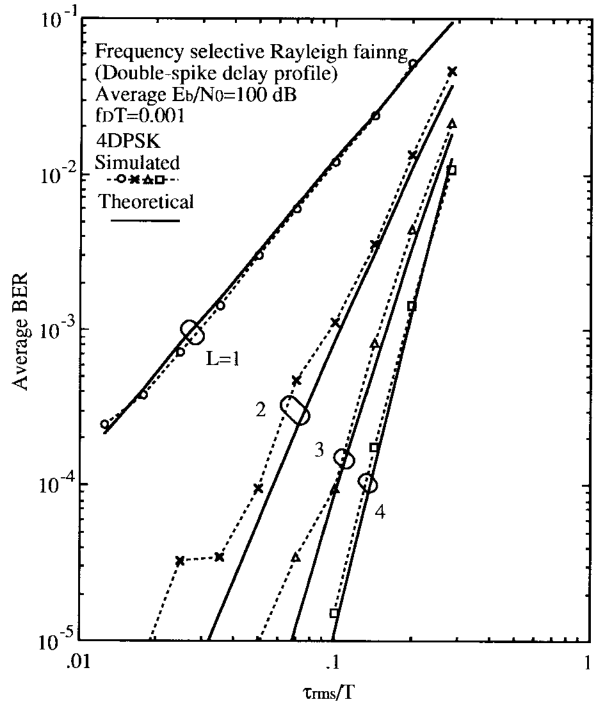


Fig. 7. Irreducible BER versus normalized delay spread with $\lambda = 0.6$, average $E_b/N_0 = 100$dB, and $f_D T = 0.001$.

spread, and multipath channel delay spread for 4DPSK transmission in Rayleigh fading channels. Computer simulations demonstrated that although the proposed scheme is somewhat inferior to optimal diversity, significant improvements can be achieved over the no-diversity system.

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