Postdetection phase combining diversity

著者	安達　文幸
journal or publication title	IEEE Transactions on Vehicular Technology
volume	43
number	2
page range	298-303
year	1994
URL	http://hdl.handle.net/10097/46489

doi: 10.1109/25.293648

Postdetection Phase Combining Diversity

Masaharu Ikura, *Member, IEEE,* and Fumiyuki Adachi, *Senior Member, IEEE*

***Abstract*—We propose a postdetection phase combining (PC) scheme for the two branch diversity reception of differential phase shift keying (DPSK) over multipath fading channels. The receiver has a differential phase detector (DPD) in each diversity branch, and the combiner weights each detector output in proportion to the vth power of the signal envelope at the detector's input.**

For $\pi/4$-shift QDPSK over frequency-flat Rayleigh fading channels, we find via computer simulation that the optimum weight factor is $v = 2$, and that our simple, practical combining scheme performs almost as well as postdetection maximal ratio combining (MRC). We demonstrate similar relative performances for frequency-selective fading channels and for channels with co-channel interference (CCI).

I. Introduction

LINEAR differential phase shift keying (DPSK) is now attracting much attention in mobile radio fields because it is more bandwidth efficient than constant envelope digital FM, and because it supports differential detection. For example, symmetrical four level DPSK (or $\pi/4$-shift QDPSK) [1], [2] has been adopted in Japanese and North American digital TDMA cellular standards [3], [4]. Although coherent detection provides the best bit error rate (BER) performance in an additive white Gaussian noise (AWGN) channel, differential detection is preferred because it is less affected by multipath fading induced random phase variations. Fading is produced by interference among multipath signals as the mobile transceiver moves. In cellular systems, bit errors are caused by co-channel interference (CCI) and the time-varying intersymbol interference (ISI) produced by multipath delay spread as well as AWGN. Diversity reception [5] can be used to improve BER performance.

Predetection diversity, which coherently combines the received faded signals before detection, may be difficult to implement because of the fast phase fluctuations of faded signals (this is the same reason for preferring differential detection). If predetection selection combining (SC) is used, switching between two fading signals may cause an abrupt phase change that will produce bit errors. On the other hand, postdetection diversity combines the detector outputs which are all in phase; thus, no co-phasing function is required and when SC is used, no switching noise is produced. For all these reasons, postdetection diversity is preferred. In postdetection diversity, detector outputs are weighted according to each branch's channel condition before combination so that the contribution of the weaker signal branch is minimized. This type of postdetection combining was proposed for binary and quaternary DPSK systems and its performance is theoretically analyzed in [5, p. 522], [6]–[9].[1] The combiner output can be mathematically expressed as $I + jQ = \Sigma_k r_k(nT)r_k^*((n-1)T)$, where $r_k(t)$ is the kth branch received faded signal. Since $r_k^*((n-1)T)$ can be viewed as a weight that is analogous to that of predetection MRC, we call this combining postdetection maximal ratio combining (MRC) [9]. Weighted operation is implicit by quadrature differential detection (DD) and addition. Symbol decision for DPSK systems is then performed based on $\Delta\psi = \arg[I + jQ]$ or $\tan^{-1} Q/I$ (however, for binary and quaternary DPSK systems, simple binary decision based on polarity of I and Q can be applied).

Another differential detection scheme is differential phase detection (DPD). The phase difference of the received signal over one symbol duration is the output of the DPD detector. The above postdetection MRC assumes quadrature DD; however, it can be implemented by first taking cosine and sine of the DPD detector output $\Delta\psi_k$ and then, multiplying them by $|r_k(nT)| \cdot |r_k((n-1)T)|$ to restore $r_k(nT)r_k^*((n-1)T)$ before combination. Note that symbol decision is based on the phase $\Delta\psi$ of the combiner output. If a combiner that directly combines the DPD detector outputs is realized, then the cosine, sine, and arc-tangent operations can be eliminated and therefore, the diversity implementation can be less complex. In this paper, we propose a new, simple two branch[2] diversity, called postdetection phase combining (PC). It is suitable for DPD, and yields the output $w_1\Delta\psi_1 + w_2\Delta\psi_2 \bmod 2\pi$, where $\Delta\psi_k$ and w_k are the DPD detector output and weight of the kth branch, respectively. The difference between postdetection MRC and PC is that the former combines two faded signals vectorically (or in the complex plane) while the latter involves scalar combining in the phase plane. Section II describes the proposed PC diversity. A theoretical comparison of postdetection PC and postdetection MRC is presented in Section III. The effects of diversity on $\pi/4$-shift QDPSK transmission under Rayleigh fading are evaluated by computer simulations, and BER performance due to AWGN, CCI, and delay spread is presented in Section IV.

II. Postdetection PC Diversity

Fig. 1 shows a block diagram of the proposed postdetection PC diversity. We assume here the multiplicative fading process to simplify the explanation of the principle of operation PC

Manuscript received May 14, 1993; revised June 25, 1993.

The authors are with R&D Department, NTT Mobile Communications Network, Inc., Kanagawa-Ken, 238-03 Japan.

IEEE Log Number 9214106.

[1] Similar postdetection combining can also be implemented for narrowband digital FM systems [5, p. 525], [10].

[2] Proposed scheme works only in two branch diversity. Nevertheless, two branch diversity is more practical than multiple branch diversity because of the limited space available for multiple antennas especially at mobile tranceivers.

0018-9545/94$04.00 © 1994 IEEE

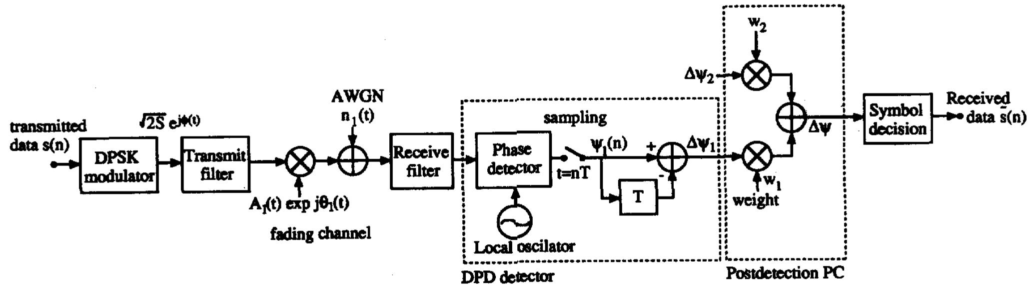


Fig. 1. Block diagram of postdetection PC diversity.

diversity (for the computer simulation described in Section IV, however, we consider frequency selective Rayleigh fading and CCI). Furthermore, we assume that the fading bandwidth (or Doppler spread) is very narrow compared to the receive filter bandwidth so that the fading process is not distorted by the receive filter.

Letting $h(t)$ be the impulse response of the overall transmit/receive filter, the DPD detector input $r_k(t)$ of the kth $(k = 1, 2)$ branch for $|t - nT| \leq 0.5T$ can be represented in complex form as $r_k(t) = \sqrt{2S}a_k(t) \exp j\,\theta_k(t)\, \Sigma_{m=-\infty}^{\infty} \{\exp j\,\phi_s(n - m)\}h(t - (n-m)T) + n_k(t)$, where $\phi_s(n)$ is the nth transmitted phase of the carrier, $a_k(t) \exp j\,\theta_k(t)$ is the fading complex envelope with $\langle a_k(t)^2 \rangle = 1, S$ is the average signal power, and $n_k(t)$ is the filtered AWGN. $\Delta\,\phi_s = \phi_s(n) - \phi_s(n-1)$ represents the transmitted $\log_2 M$-bit symbol $s(n).\, \Delta\,\phi_s$ takes on one of the equally spaced M values $\{(2\pi m/M) + \delta; m = 0, 1, 2, \cdots, (M-1)\}$, where $\delta = 0$ for asymmetrical DPSK and π/M for symmetrical DPSK or π/M-shift MDPSK. In DPD, the phase of the detector input $r_k(t)$ relative to the local oscillator output with constant frequency is detected and sampled at time $t = nT$. The sampled phase $\psi_k(n)$ is then reduced by the previous value $\psi_k(n-1)$ to obtain the phase difference $\Delta\,\psi_k (= \psi_k(n) - \psi_k(n-1) \bmod 2\,\pi)$ of the received signal over one symbol duration. Here we assume that the local oscillator has the same frequency as the unmodulated carrier. The oscillator is not necessarily phase coherent because the constant phase error is removed by the differential operation. Therefore, $\Delta\,\psi_k$ can be expressed as $\Delta\,\psi_k = \arg[r_k(nT)r_k^*((n-1)T)]$. Assuming a Nyquist filter response, $h(0) = 1$ and $h(mT) = 0$ if $m \neq 0$ and, therefore

$$\begin{aligned} r_k(nT) &= \sqrt{2S}a_k(nT)e^{j(\phi_s(n)+\theta_k(nT))} + n_k(nT) \\ &= R_k(nT)e^{j(\phi_s(n)+\theta_k(nT)+\eta_k(nT))} \end{aligned} \quad (1)$$

where $R_k(t)$ is the envelope of the fading signal plus noise and $\eta_k(t)$ is the phase noise due to AWGN. The DPD detector output $\Delta\,\psi_k$ defined over $[-\pi, \pi)$ can be represented as

$$\Delta\,\psi_k = \Delta\,\phi_s + \Delta\,\theta_k + \Delta\,\eta_k \bmod 2\,\pi \quad (2)$$

where $\Delta\,\theta_k = \theta_k(nT) - \theta_k((n-1)T)$ and $\Delta\,\eta_k = \eta_k(nT) - \eta_k((n-1)T)$.

The DPD detector outputs of the two branches are weighted and combined. The resultant combiner output is given by

$$\Delta\,\psi = w_1 \Delta\,\psi_1 + w_2 \Delta\,\psi_2 \bmod 2\,\pi \quad (3)$$

where w_k is the weight and $w_1 + w_2 = 1$. The phase noises, $\Delta\,\theta_k$ and $\Delta\,\eta_k$, become large when the received signal fades. Therefore, the weight should be chosen to minimize the contribution of the branch having the weaker signal. In this paper, we use the weight proportional to the vth power of $R_k(nT)$. Therefore,

$$w_k = \frac{R_k^v(nT)}{R_1^v(nT) + R_2^v(nT)}. \quad (4)$$

Substituting the above into (3), we obtain

$$\begin{aligned} \Delta\,\psi = {} & \Delta\,\phi_s \\ & + \frac{(\Delta\,\theta_1 + \Delta\,\eta_1)R_1^v(nT) + (\Delta\,\theta_2 + \Delta\,\eta_2)R_2^v(nT)}{R_1^v(nT) + R_2^v(nT)} \\ & \bmod 2\,\pi. \end{aligned} \quad (5)$$

The first term of the right-hand side of (5) is the transmitted phase difference and the second term is the sum of the weighted phase noises. The decision rule is

$$\text{choose}\, \Delta\,\tilde{\phi}_s = \frac{2\,\pi\,m}{M} + \delta \quad \text{if}\, \left|\Delta\,\psi - \frac{2\,\pi\,m}{M} - \delta \bmod 2\,\pi\right| \text{is minimum} \quad (6)$$

where $m = 0, 1, \cdots (M-1)$. The received symbol $\tilde{s}(n)$ is recovered from $\Delta\,\tilde{\phi}_s$. It should be noted that the weights are normalized so that $w_1 + w_2 = 1$, the decision rule is the same as for the no diversity case.

The above combining scheme is similar to that proposed for digital FM [5, p. 525], [10]. However, because of the $2\,\pi$ periodicity of the phase, the simple addition described by (3) may produce erroneous combining. Fig. 2 shows the relationship among the two DPD outputs, $\Delta\,\psi_1$ and $\Delta\,\psi_2$, and the combiner output $\Delta\,\psi$. Remember that $\Delta\,\psi_1$ and $\Delta\,\psi_2$ are distributed around $\Delta\,\phi_s$ and that the probabilities of $\Delta\,\psi_1$ and $\Delta\,\psi_2$ monotonically reduce as they deviate from $\Delta\,\phi_s$. This suggests that the combiner output phase should be inside the phase region sandwiched by $\Delta\,\psi_1$ and $\Delta\,\psi_2$. However, the resultant combined phase is outside the region. This erroneous combining always happens when $|\Delta\,\psi_1 - \Delta\,\psi_2| > \pi$. Fig. 2

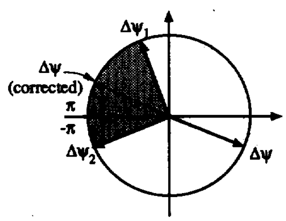


Fig. 2. Phase combiner output.

is an example of this. To avoid this problem, we introduce the following correction

$$\Delta\psi_2\,(\text{corrected}) = \begin{cases} \Delta\psi_2 + 2\pi & \text{if } \Delta\psi_1 - \Delta\psi_2 > \pi \\ \Delta\psi_2 - 2\pi & \text{if } \Delta\psi_1 - \Delta\psi_2 < -\pi \end{cases}. \quad (7)$$

In Fig. 2, corrected combiner output is shown as the dotted arrow. This correction brings the combiner output inside the region sandwiched by $\Delta\psi_1$ and $\Delta\psi_2$.

III. Theoretical Comparison with Postdetection MRC

We show below that postdetection PC with $v = 2$ is an approximate of postdetection MRC. Postdetection MRC weights and combines vectorically the quadrature differential detector (DD) outputs to yield the output $I + jQ = \Sigma_{k=1}^{2} r_k(nT)r_k^*((n-1)T)$ [9]. Using (1), we have

$$I + jQ = \sum_{k=1}^{2} w_{\text{MRC},k} \exp j(\Delta\phi_s + \Delta\theta_k + \Delta\eta_k) \quad (8)$$

where $w_{\text{MRC},k}$ is normalized so that $w_{\text{MRC},1} + w_{\text{MRC},2} = 1$ and is given by $w_{\text{MRC},k} = R_k(nT)R_k((n-1)T)/(R_1(nT)R_1((n-1)T) + R_2(nT)R_2((n-1)T))$. To recover the transmitted phase difference, the phase decision rule described by (6) can be applied if we use $\Delta\psi_{\text{MRC}} = \arg[I + jQ]$. Since the sum of $\Delta\theta_k$ and $\Delta\eta_k$ can be assumed to be much less than unity most of the time, $\exp j(\Delta\phi_s + \Delta\theta_k + \Delta\eta_k)$ can be approximated as $[1 + j(\Delta\theta_k + \Delta\eta_k)]\exp j\Delta\phi_s$. Therefore, we have $I + jQ \approx [1 + j\,\Sigma_{k=1}^{2} w_{\text{MRC},k}(\Delta\theta_k + \Delta\eta_k)]\exp j\Delta\phi_s$ since $w_{\text{MRC},1} + w_{\text{MRC},2} = 1$. Since the weighted sum of phase noises is still small, $1 + j\,\Sigma_{k=1}^{2} w_{\text{MRC},k}(\Delta\theta_k + \Delta\eta_k)$ can be approximated as $\exp j\,\Sigma_{k=1}^{2} w_{\text{MRC},k}(\Delta\theta_k + \Delta\eta_k)$. As a result, we have

$$\begin{aligned}\Delta\psi_{\text{MRC}} &= \arg[I + jQ] \\ &\approx \Delta\phi_s + \sum_{k=1}^{2} w_{\text{MRC},k}(\Delta\theta_k + \Delta\eta_k) \bmod 2\pi \quad (9)\end{aligned}$$

which is similar to (5). In many practical situations, fading is very slow compared to the symbol transmission rate $1/T$ and thus $R_k(nT) \approx R_k((n-1)T)$.[3] This means that the MRC weight can be well approximated by the PC weight given by (4) with $v = 2$. The above discussion implies that postdetection PC with $v = 2$ is an approximation of postdetection MRC. So, it is anticipated that PC yields almost the same BER performance as postdetection MRC. This is confirmed by computer simulation in Section IV.

[3] Suppose that the carrier frequency is 1 GHz and traveling speed of mobile tranceiver is 100 km/h, the maximum Doppler frequency is $f_D = 93$ Hz. For transmission symbol rates above 10 k symbols/s, the normalized Doppler frequency $f_D T$ is less than 0.01. In this situation, fading can be considered to be very slow.

IV. Computer Simulation

Cellular systems efficiently utilize the limited radio spectrum resources by reusing the same radio frequencies at spatially separated cells. Thus, CCI performance is important becasue it determines the reuse distance of the same frequency and thus, the spectrum efficiency. The ISI due to multipath channel delay spread cannot be ignored in high speed digital signal transmission. The effect of postdetection PC diversity on the BER performance of $\pi/4$-shift QDPSK under the influence of AWGN, CCI, and delay spread was evaluated by computer simulation.

A. Transmission System Model

The transmitted data is a 9-stage PN sequence, and that for CCI is a 15-stage PN sequence. These are input to the two independent $\pi/4$-shift QDPSK modulators. The Gray mapping rule of di-bit symbol $s(n)$ into the differential phase $\Delta\phi_s = \phi_s(n) - \phi_s(n-1)$ is assumed: $\Delta\phi_s = \pi/4$ if $s(n) = (1,1), = -\pi/4$ if $s(n) = (1,0), = 3\pi/4$ if $s(n) = (0,1)$, and $= -3\pi/4$ if $s(n) = (0,0)$. The modulated desired signal and CCI are transmitted over Rayleigh fading channels. If the multipath channel rms delay spread normalized by the symbol duration T is smaller than about 0.2, the delay profile shape is of no importance [9]. Therefore, for the desired signal fading, we assume the double-spike delay profile (or two ray model) with equal average power, each ray is subjected to independent Rayleigh fading. When the time difference between the two rays is τ seconds, the rms delay spread τ_{rms} becomes $\tau_{\text{rms}} = \tau/2$ s. When $\tau_{\text{rms}} = 0$, it becomes multiplicative (nonfrequency selective) Rayleigh fading. On the other hand, fading on the CCI is assumed to be multiplicative. The desired signal, CCI, and AWGN are added to form the input to the DPD detector. The recovered timing was assumed to track the first central moment of the double-spike delay profile as assumed in the theoretical analysis in [9]. The kth branch detector input sample corresponding to (1) can be rewritten as

$$\begin{aligned} r_k(nT) = \sqrt{S}\Big[&a_{k1}(nT)e^{j\theta_{k1}(nT)} \\ &\cdot \sum_{m=-\infty}^{\infty} e^{j\phi_s(n-m)} h(mT + \tau_{\text{rms}}) \\ &+ a_{k2}(nT)e^{j\theta_{k2}(nT)} \sum_{m=-\infty}^{\infty} e^{j\phi_s(n-m)} \\ &\cdot h(mT + \tau_{\text{rms}})\Big] + \sqrt{2I}b_k(nT)e^{j\xi_k(nT)} \\ &\cdot \sum_{l=-\infty}^{\infty} e^{j\phi_i(n-l)} h(lT + \Delta T) + n_k(nT) \quad (10)\end{aligned}$$

where I is the average CCI power. Average signal energy per bit-to-AWGN power sprectrum density ratio (E_b/N_0) and SIR are defined as $0.5(S/N)T$ and S/I, respectively. The independent Rayleigh fading samples $a_{k1}(nT)\exp j\,\theta_{k1}(nT), a_{k2}(nT)\exp j\,\theta_{k2}(nT)$, and $b_k(nT)$ $\exp j\,\xi_k(nT)$ are generated by assuming that 16 multipaths come from all directions (equally spaced arriving angles), with equal amplitude and that the mobile transceiver travels at constant speed. Fading speed can be determined by the normalized maximum Doppler frequency $f_D T$, where f_D = (mobile speed)/(carrier wavelength). In the simulation, we fixed the value of $f_D T$ to 1.9×10^{-3} which corresponds to $f_D = 40$ Hz and $T^{-1} = 21\,k$ symbols/s. In (10), ΔT is the modulation timing offset of the CCI relative to the desired signal. We assume square-root cosine Nyquist filters with a roll-off factor of α as the transmit and receive filters. $h(t)$ is given by

$$h(t) = \frac{\sin(\pi\,t/T)}{\pi\,t/T}\frac{\cos(\alpha\,\pi\,t/T)}{1-(2\alpha\,t/T)^2}. \tag{11}$$

The DPD detector outputs are combined based on the algorithm described in Section II and symbol decision is performed based on

$$\tilde{s}(n) = \begin{cases} (0,1) & \text{if } \pi/2 \le \Delta\,\psi \le \pi \\ (1,1) & \text{if } 0 \le \Delta\,\psi < \pi/2 \\ (1,0) & \text{if } -\pi/2 \le \Delta\,\psi < 0 \\ (0,0) & \text{if } -\pi < \Delta\,\psi(n) < -\pi/2 \end{cases}. \tag{12}$$

The number of bit errors are counted by comparing $s(n)$ and $\tilde{s}(n)$ and are divided by the total number of transmitted bits to obtain average BER.

B. Results

We first investigated the dependence of the diversity improvement on the weight factor v assuming that the desired signal was subjected to multiplicative fading and perturbed by AWGN only [$\tau_{\text{rms}} = 0$ and average SIR $\to \infty$, and thus, (10) reduces to (1)]. The results obtained at average E_b/N_0 values of 10 dB and 15 dB are shown in Fig. 3. For comparison, the BER's with postdetection SC are also shown. For small v values, PC diversity is inferior to SC. This is because two DPD detector outputs are combined with almost equal weight (since $R_k^v \to 1$ for $v \to 0$) and thus, large phase noise appears in the combiner outputs when one of the two branch signals fades. As the value of v is increased, the BER decreases at first and then increases. For large v values, the weight of the branch having the larger signal level becomes unity and the other becomes zero; therefore, the combining operation becomes similar to SC. Thus, the BER with postdetection PC diversity approaches that with postdetection SC as $v \to \infty$. This can be clearly seen in Fig. 3. The optimum weight factor that can minimize the BER is found to be 2. Fig. 3 also shows results for computer simulated postdetection MRC diversity. (In addition, we examined the effect of raising $w_{\text{MRC},k}$ in (8) to the power of $v/2$. $R_k((n-1)T)$ was approximated as $R_k(nT)$. We observed that the optimum weight factor is $v = 2$, verifying that MRC is optimum.) We find that postdetection PC with $v = 2$ performs almost as well as MRC, as predicted theoretically in Section III. The following computer simulations use the weight factor $v = 2$.

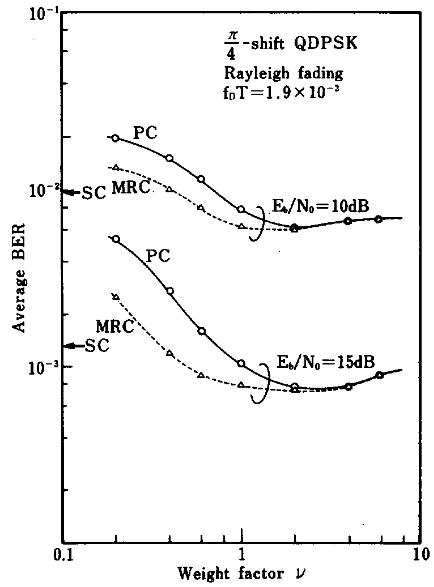


Fig. 3. Effect of weight factor.

The BER performance due to AWGN is shown in Fig. 4 as a function of average E_b/N_0. It can be seen again that in Rayleigh fading, postdetection PC diversity performs almost as well as MRC, achieving about a 1.5 dB larger diversity gain than SC. It is interesting to notice that under the no fading condition, a diversity gain of about 3 dB is obtained for PC while only a slight improvement is obtained for SC. This is because two DPD detector outputs are always combined for PC while only one of the detector outputs is always selected for SC. (In theory, however, there is no diversity benefit in SC from unfaded channels with identical noise statistics. The small performance difference observed in our results is due to the short-term noise power being different in each diversity branch.)

To investigate the BER performance in CCI limited channels, AWGN was ignored and the desired signal was assumed to be subjected to multiplicative Rayleigh fading ($\tau_{\text{rms}} = 0$). Since CCI is transmitted from a different cell, its modulation timing is not always identical to that of the desired signal. We assumed that the modulation timing offset $\Delta\,T$ was uniformly distributed within $|\Delta\,T| \le T/2$. Due to timing offset, ISI is produced on the CCI samples. Therefore, BER due to CCI depends on the CCI symbol sequence and the filter roll-off factor α. We took into account ISI from two adjacent symbols on each side when we calculated the detector input $r_k(nT)$, i.e., $|l| \le 2$ in (10). We first obtained average BER's for $\Delta\,T = 0, \pm T/4$, and $T/2$ and then averaged them.

Simulation results are plotted in Fig. 5 for $\alpha = 0.5$. A diversity gain of about 7 dB is observed for postdetection PC at a BER of 10^{-2} in fading, which is almost as goood as MRC (about 1 dB superior to SC). Also seen is that in no fading, a diversity gain of about 1 dB can be obtained at BER $= 10^{-2}$. However, for smaller BER's, the gain diminishes.

To determine the BER performance due to delay spread, the effects of AWGN and CCI were ignored. When delay spread

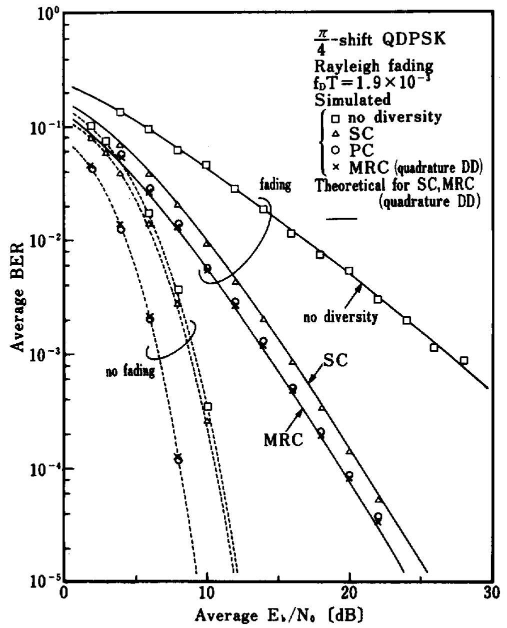


Fig. 4. BER performance due to AWGN.

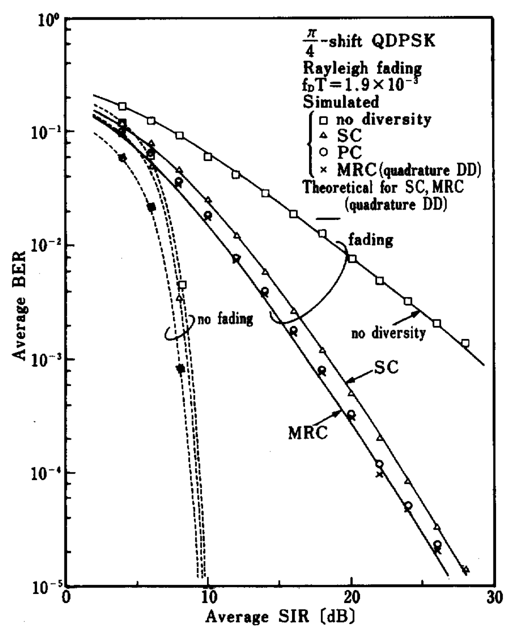


Fig. 5. BER performance due to CCI.

exists, ISI is produced on the detector input signal because the sampling timing is shifted from the ideal position. We took into account ISI from two adjacent symbols on either side, i.e., $|m| \leq 2$ in (10). The simulation results are shown in Fig. 6 as a function of normalized delay spread $\tau_{\rm rms}/T$. It is again seen that postdetection PC diversity performs almost as well as MRC; the allowable value of $\tau_{\rm rms}/T$ at $\mathrm{BER} = 10^{-2}$ can be enlarged from 0.09 without diversity to 0.19 which is 1.1 times larger than that using SC.

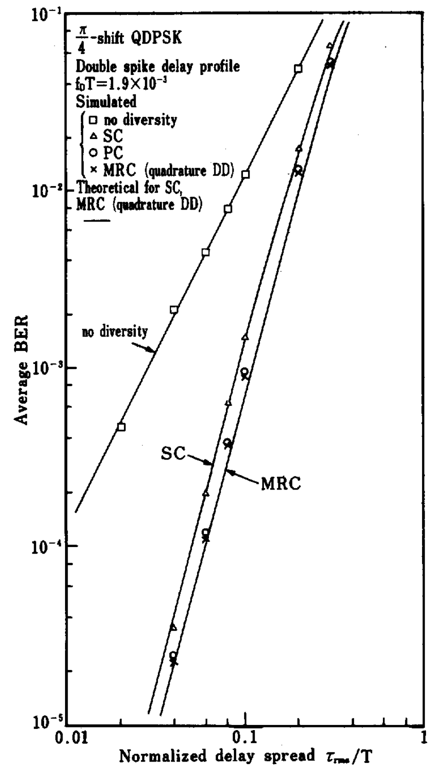


Fig. 6. BER performance due to delay spread.

V. Conclusion

This paper has proposed postdetection PC diversity which is simple and suitable for DPD. Each branch DPD output is weighted in proportion to the squared envelope of the detector input. The diversity improvement achievable was investigated by computer simulations for $\pi/4$-shift QDPSK signal transmission over mobile radio Rayleigh fading channels. It was shown that the diversity gain is almost identical to that of postdetection MRC proposed for quadrature DD. The DPD detector output $\Delta\,\psi_k$ is related to the quadrature DD detector output i_k+jq_k by $i_k+jq_k = \cos\,\Delta\,\psi_k + j\,\sin\,\Delta\,\psi_k$. Therefore, postdetection MRC can also be implemented for DPD by first taking the cosine and sine of the DPD detector output and weighting them in proportion to the squared envelope of the detector input prior to combination. However, the cosine and sine operation is eliminated in PC diversity with negligible performance degradation. Although simulation results were presented only for $\pi/4$-shift QDPSK, postdetection PC diversity is also applicable to other digital modulation schemes with differential detection.

References

[1] F. G. Jenks, P. D. Morgan, and C. S. Warren, "Use of four-level phase modulation for digital mobile radio," *IEEE Trans. Electromagnet. Compat.*, vol. EMC-14, pp. 113–128, Nov. 1972.

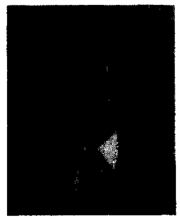
[2] Y. Akaiwa and Y. Nagata, "Highly efficient digital mobile communications with a linear modulation method," *IEEE J. Select. Areas Commun.*, vol. SAC-5, pp. 890–895, June 1987.

[3] J. Uddenfeldth, K. Raith, and B. Hedberg, "Digital technologies in cellular radio," in *Proc. 38th IEEE VTC,* Philadelphia, PA, June 15–17, 1989, pp. 516–519.
[4] N. Nakajima, M. Kuramoto, K. Kinoshita, and T. Utano, "A system design for TDMA mobile radio," in *Proc. 40th IEEE VTC,* Florida, May 6–9, 1990, pp. 295–298.
[5] W. C. Jakes, Jr., *Microwave Mobile Communications.* New York: Wiley, 1974.
[6] P. A. Bello and B. D. Nelin, "Effect of frequency selective fading on the binary error probability of incoherent and differentially coherent matched filter receivers," *IEEE Trans. Commun. Syst.,* vol. CS-11, pp. 170–186, June 1963.
[7] C. C. Bailey and J. C. Lindenlaub, "Further results concerning the effect of frequency selective fading on differentially coherent matched filter receivers," *IEEE Trans. Commun. Technol.,* vol. COM-16, p. 749, Oct. 1968.
[8] J. Horikoshi, "Error performance improvement of QDPSK in the presence of cochannel or multipath interference using diversity," (in Japanese) *Trans. IECE Japan,* vol. J67-B, pp. 24–31, Jan. 1984.
[9] F. Adachi and K. Ohno, "BER performance of QDPSK with postdetection diversity reception in mobile radio channels," *IEEE Trans. Veh. Technol.,* vol. 40, pp. 237–249, Feb. 1991.
[10] ——, "Experimental evaluation of postdetection diversity reception of narrow-band digital FM signals in Rayleigh fading," *IEEE Trans. Veh. Technol.,* vol. 38, pp. 216–221, Nov. 1989.

Masaharu Ikura (M'91) was born in Saitama, Japan, in 1963. He received the B.S. and M.S. degrees in mathematics from Waseda University, Tokyo, Japan, in 1986 and 1988, respectively.

Since joining NTT Radio Communication Systems Laboratories in 1989, he has been studying TDMA signal transmission techniques. In 1992, he transferred to the R&D Department in NTT Mobile Communications Network Inc., and was involved in the development of the base station modem based on the Japanese Personal Digital Cellular Standard. He is now an Assistant Manager in the Corporate Planning Department.

Mr. Ikura is a member of the Institute of Electronics, Information, and Communication Engineering of Japan.



Fumiyuki Adachi (M'79–SM'90) graduated from Tohoku University, Japan, in 1973 and was awarded Dr. Engineering from the same university in 1984.

In 1973, he joined the Nippon Telegraph & Telephone Corporation (NTT) Laboratories in Japan and in 1992, he transferred to NTT Mobile Communications Network, Inc. His major research activities center mobile communication digital signal processing, including digital modulation/demodulation, diversity reception, channel coding, and CDMA/TDMA. During academic year of 1984/1985, he was a United Kingdom SERC Visiting Research Fellow at the Department of Electrical Engineering and Electronics of Liverpool University. He authored chapters of three books: Y. Okumura and M. Shinji Eds., *Fundamentals of Mobile Communications* (in Japanese) (published by IEICE Japan, 1986), M. Shinji, Ed., *Mobile Communications* (in Japanese) (published by Maruzen Publishing Co., 1989), and M. Kuwabara ed., *Digital Mobile Communications* (in Japanese) (published by Kagaku Shinbun-sha, 1992).

Dr. Adachi was a Treasurer in 1989, and is now a Secretary of IEEE Vehicular Technology Society Tokyo Chapter. He is a co-recipient of the 1980 and 1990 IEEE Vehicular Technology Society Paper of the Year Awards. He is a member of the Institute of Electronics, Information, and Communication Engineers of Japan.