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BER Performance of QDPSK with Postdetection Diversity Reception in Mobile Radio Channels

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Abstract—Bit error rate (BER) performance of Nyquist raised cosine filtered quaternary differential phase shift keying (RC-QDPSK) and $\pi/4$ -shift RC-QDPSK with postdetection diversity reception is theoretically analyzed for Rayleigh fading channels. Selection combiner (SC), equal-gain combiner (EGC), and maximal-ratio combiner (MRC) are considered for postdetection diversity reception. Diversity improvements on the average BER due to additive white Gaussian noise (AWGN), co-channel interference, random FM noise, and delay spread of the multipath channel are calculated. Diversity reception is shown to be a powerful technique for reducing the effects of delay spread (as well as for reducing the effects of AWGN, co-channel interference and random FM noise) if root mean square (rms) delay spread is small compared to the symbol duration. Also analyzed is the combined effect of diversity reception and error control. The required E_b/N_0 and allowable rms delay spread for achieving a certain BER and spectrum efficiency in cellular mobile systems are calculated when short BCH forward error correcting codes are used. Optimum code rate is shown to be relatively high (around 0.7) when diversity reception is used. Performance of automatic repeat request (ARQ) incorporated with time diversity reception is investigated.

I. INTRODUCTION

RECENTLY, demands for bandwidth efficient secure voice and data transmissions have been increasing in land mobile radio. Digital FM (or continuous phase modulation), such as GMSK [1] and GTFM [2], has been considered as a promising modulation scheme because of its advantage of using power-efficient class C power amplifiers. However, linear modulation, such as Nyquist raised cosine filtered quaternary differential phase shift keying (RC-QDPSK), is currently attracting a lot of attention because a much narrower bandwidth than digital FM is achieved [3]–[6]. Since the mobile radio channel is characterized by many different propagation paths with different time delays [7], multipath Rayleigh fading is produced. Errors are caused by time-varying intersymbol interference (ISI) from delay spread [8]–[11], as well as additive white Gaussian noise (AWGN) and random FM noise. Another cause of errors in cellular systems is cochannel interference produced by the reuse of the same radio frequency at the spatially separated cells. Diversity reception and error control (forward error correction (FEC) and automatic repeat request (ARQ)) are effective techniques for combating multipath fading. There are many possible implementations of diversity reception systems [12, ch. 6] but for mobile radio, postdetection diversity is attractive because the cophasing function necessary in predetection combiners is not required [13].

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Binary DPSK performance was analyzed by Bello and Nelin [8] assuming rectangular modulation pulse and Gaussian delay profile. Bailey and Lindenlaub [9], and Garber and Parsley [14] extended the analysis of [8] to show that the BER performance is not sensitive to the shape of the delay profile, but to the shape of modulation pulse. Diversity reception was considered in [8], [9], but the type of combiner assumed was the postdetection maximal-ratio combiner (MRC) considered in this paper. The simple and practical selection combiner (SC) was not considered. None of the above papers on DPSK included co-channel interference performance analysis. Performance analysis of QDPSK with postdetection MRC diversity is found in [15], however, the analysis was limited to the case of rectangular modulation pulse and double-spike delay profile. Unfortunately, there has been no unified performance analysis of RC-QDPSK with diversity reception taking into account the effects of AWGN, co-channel interference, random FM noise, and delay spread of the mobile multipath channel. Concerning application of error control techniques to mobile channels, only a few papers are available [16]–[19]; however, no performance analysis in frequency selective fading (where the errors are produced by delay spread) has been presented.

This paper presents a theoretical analysis of BER performance of ($\pi/4$ -shift) RC-QDPSK with postdetection diversity reception in the multipath Rayleigh fading channels. As for the diversity combiner, we consider the simple SC and a weighted combining of each branch differential detector (DD) output. The closed form expression for average BER is developed in Section II to allow investigation of the effects of AWGN, co-channel interference, random FM noise, and delay spread. Section III presents calculated BER's and shows that postdetection diversity reception can significantly reduce the effects of multipath fading. Section IV investigates the combined effects of diversity reception and FEC to find the optimum code rate. Also, ARQ incorporated with time diversity reception is investigated.

II. BER PERFORMANCE WITH DIVERSITY RECEPTION

A. Transmission system

A schematic diagram of a transmission system for ($\pi/4$ -shift) RC-QDPSK with diversity reception is shown in Fig. 1. Base-band equivalent (complex envelope) representation is used. The QDPSK modulation scheme groups input binary data sequence into di-bit symbols (a_n, b_n) that are mapped to the differential phase of the transmitted carrier. There are two mapping rules of the differential phase as shown in Fig. 2 (Gray encoding is employed). A $\pi/4$ -shift QDPSK modulation is preferred because its signal envelope has no zeros, and it has an advantage of less spectrum spreading due to nonlinearity of the transceivers.

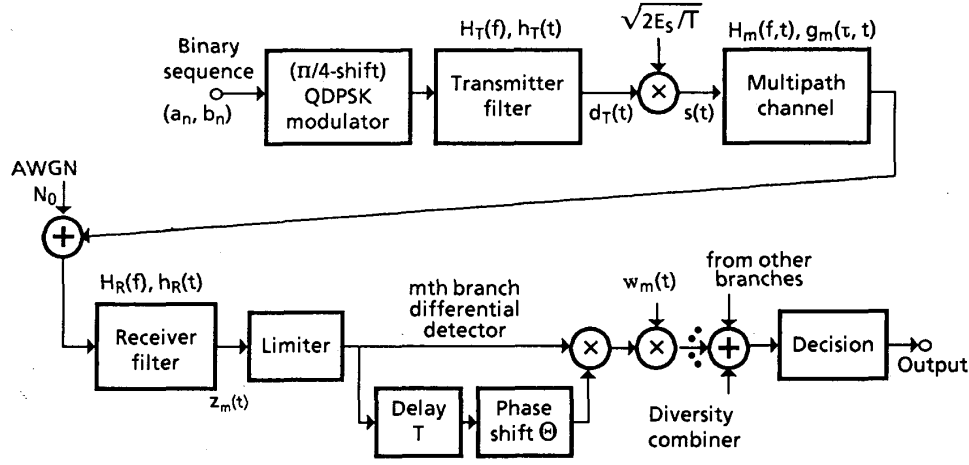


Fig. 1. Schematic diagram of $(\pi/4)$ -shift QDPSK in complex representation.

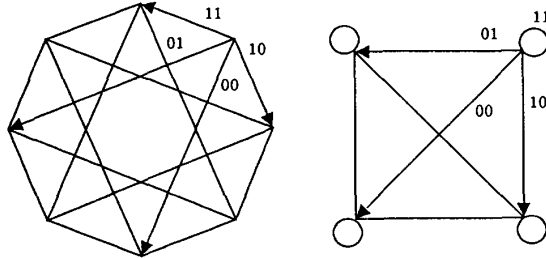


Fig. 2. Signal space diagram for two QDPSK modulations. (a) $\pi/4$ -shift QDPSK. (b) QDPSK.

Both the transmitter and receiver require channel filters, $H_T(f)$ and $H_R(f)$. The primary function of the transmitter filter is to achieve narrow-band spectrum. To achieve ISI free transmission, the RC transfer function $H^2(f)$ with roll-off factor α ($0 \leq \alpha \leq 1$) can be employed as an overall filter response:

$$H^2(f) = \begin{cases} 1, & 0 \leq |f| \leq \frac{1-\alpha}{2T} \\ \cos^2 \left[\frac{\pi T}{2\alpha} \left(f - \frac{1-\alpha}{2T} \right) \right], & \frac{1-\alpha}{2T} < |f| \leq \frac{1+\alpha}{2T} \\ 0, & \text{elsewhere} \end{cases} \quad (1)$$

where T is the symbol duration. In the AWGN channel (without fading and cochannel interference), the optimum performance is achieved when $H^2(f)$ is shared equally (thus, square root RC filter) at transmitter and receiver. The square root RC receiver filter at the IF stage is difficult to implement. Another possible combination of filters is that $H^2(f)$ is realized by the transmitter filter alone, and a narrow-band IF filter is used at the receiver. Narrow-band IF filters for current analog FM cellular systems can be applied at the cost of some performance degradation due to ISI introduced by the receiver filter (because overall transfer function is not RC). In this paper, a Gaussian receiver filter is assumed, and a performance comparison between the

square root RC filter case and the Gaussian receiver filter case is presented.

Although QDPSK signals can be demodulated by either differential detection or coherent detection with differential decoding, differential detection is attractive for mobile radio applications because its BER performance is less affected by rapid random phase fluctuations in the received signal due to fast fading. Moreover, it eliminates the need for carrier recovery.

B. Description of Receiver Signal

The transmitted signal of $(\pi/4)$ -shift RC-QDPSK can be represented in complex form as

$$s(t) = \sqrt{\frac{2E_s}{T}} d_T(t), \quad (2)$$

where

$$d_T(t) = \sum_{n=-\infty}^{+\infty} e^{j\Phi_n} \frac{h_T(t - nT)}{\sqrt{\frac{1}{T} \int_{-\infty}^{+\infty} h_T^2(t) dt}}, \quad (3)$$

E_s is the symbol energy, Φ_n is the carrier phase for the n th symbol transmission, and $h_T(t)$ is the impulse response of the transmitter filter $H_T(f)$. Di-bit symbol (a_n, b_n) to be transmitted is mapped to the differential phase $\Delta\Phi_n = \Phi_n - \Phi_{n-1}$:

$$\Delta\Phi_n = \Theta + \begin{cases} 3\pi/4 & (a_n, b_n) = (-1, 1) \\ \pi/4 & (1, 1) \\ -\pi/4 & (1, -1) \\ -3\pi/4 & (-1, -1) \end{cases} \quad (4)$$

with $\Theta = -\pi/4$ for RC-QDPSK and 0 for $\pi/4$ -shift RC-QDPSK.

Signal transmission between mobile and base stations take place over multipath channels. We assume that the fading variations are much slower compared to the symbol rate, thus the channel transfer function remains almost constant over a one-symbol duration T . The input to the m th branch DD of an M

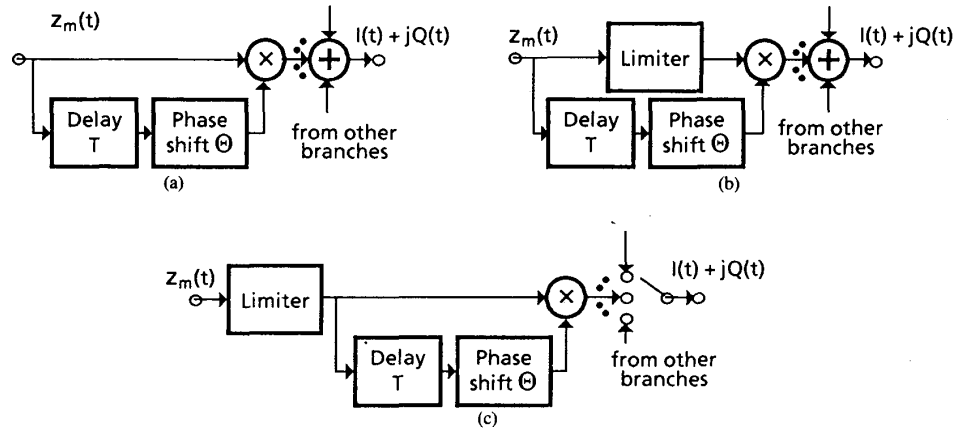


Fig. 3. Postdetection diversity combiners incorporated in DD. (a) MRC. (b) EGC. (c) SC.

branch postdetection diversity receiver can be written as [8]

$$\begin{aligned} z_m(t) &= z_{sm}(t) + z_{nm}(t) \\ &= \int_{-\infty}^{+\infty} S(f) T_m(f, t) H_R(f) e^{j2\pi f t} df \\ &\quad + z_{nm}(t), \end{aligned} \quad (5)$$

where $S(f)$ is the spectrum of $s(t)$, $H_R(f)$ is the baseband equivalent transfer function of the receiver filter (where $H_R(0) = 1$), $T_m(f, t)$ is the baseband equivalent transfer function of the multipath channel for m th branch at time t , and $z_{nm}(t)$ is the AWGN component with one-sided spectral density N_0 . Introducing the baseband equivalent channel impulse response, $g_m(\tau, t)$, measured from the instant of application of a unit impulse at the transmitter at t , $T_m(f, t)$ can be represented as

$$T_m(f, t) = \int_{-\infty}^{+\infty} g_m(\tau, t) e^{-j2\pi f \tau} d\tau. \quad (6)$$

Assuming that the impulse response at τ is due to the sum of many independent impulses produced by reflection from different buildings, $g_m(\tau, t)$ results in the zero-mean complex Gaussian process of t according to the central limit theorem [10], [11]. Therefore, $T_m(f, t)$ also becomes the zero-mean complex Gaussian process. The fading signals at the different diversity branches are independent and have identical statistical characteristics. Therefore, with the transmitting symbol sequence given, the M received signals are independent complex Gaussian time-processes.

The delay-time correlation function of the multipath channel for each branch is given by

$$\langle g_m(\tau, t) g_m^*(\tau - \lambda, t - \mu) \rangle = \xi_s(\tau, \mu) \delta(\lambda), \quad (7)$$

where $\langle \cdot \rangle$ denotes ensemble average, the asterisk denotes complex conjugate, and $\delta(\cdot)$ is the delta function. It was shown [8]–[13] that in frequency selective fading, BER is governed by the rms delay spread τ_{rms} , which is defined as

$$\tau_{rms} = \sqrt{\int_{-\infty}^{+\infty} (\tau - \tau_{mean})^2 \xi_s(\tau, 0) d\tau}, \quad (8)$$

where $\xi_s(\tau, 0)$ is called the delay profile, and

$$\tau_{mean} = \int_{-\infty}^{+\infty} \tau \xi_s(\tau, 0) d\tau \quad (9)$$

is the mean delay. (We have assumed $\int \xi_s(\tau, 0) d\tau = 1$ without loss of generality.) It was reported [20] that the measured BER performance of QDPSK under frequency selective Rayleigh fading is in good agreement with the theoretically predicted performance.

C. Postdetection Diversity Combining

With differential detection, amplitude-limited input signal is multiplied by its delayed version with the time delay T and phase shift Θ . The complex DD output of m th branch is

$$I_m(t) + jQ_m(t) = e^{-j\Theta} \frac{z_m(t) z_m^*(t - T)}{|z_m(t)| |z_m(t - T)|}. \quad (10)$$

The postdetection diversity combiner weights each branch DD output to combine so that the contribution of the branch with weak signal is reduced. The postdetection diversity combining schemes for reception of digital FM signals [13] can be applied to our case. The weighting factor $w_m(t)$ for m th branch is chosen as $|z_m(t)| |z_m(t - T)|$ for MRC, and $|z_m(t - T)|$ for equal-gain combiner (EGC). The combiner output $I(t)$ and $Q(t)$ can be represented as

$$\begin{aligned} I(t) + jQ(t) &= \sum_{m=1}^M w_m(t) \{I_m(t) + jQ_m(t)\} \\ &= \begin{cases} e^{-j\Theta} \sum_{m=1}^M z_m(t) z_m^*(t - T), & \text{for MRC} \\ e^{-j\Theta} \sum_{m=1}^M z_m(t) \frac{z_m^*(t - T)}{|z_m(t)|}, & \text{for EGC.} \end{cases} \end{aligned} \quad (11)$$

Fig. 3 shows possible postdetection diversity combiner implementations.

These postdetection combiners can be viewed in a different way. The weighting factor of each branch can be considered to be $z_m^*(t - T)$ for MRC and $z_m^*(t - T) / |z_m(t)|$ for EGC,

and thus these combiners are analogous to the predetection case [12] except that the postdetection diversity weighting factor bears the modulation information. This is the reason that these combiners are referred to as postdetection MRC and EGC in this paper. In each branch, however, weighting for both diversity combining and detection are performed simultaneously. It has

Rayleigh distributed, the average BER can be easily obtained by averaging the conditional BER. (For more details on derivation, see [13].)

Exact expressions for average BER can be derived for SC and MRC. (For EGC, an approximate expression is presented in Section II-E.) The average BER expression can be given as¹

$$Pe = \begin{cases} \frac{1}{2} \left[1 - \frac{1}{2} \sum_{m=1}^M \binom{M}{m} (-1)^{m+1} \left\{ \frac{a_0 \rho_c}{\sqrt{\rho_c^2 + m(1 - |\rho|^2)}} + \frac{b_0 \rho_s}{\sqrt{\rho_s^2 + m(1 - |\rho|^2)}} \right\} \right], & \text{for SC} \\ \frac{1}{2} \left[1 - \frac{1}{2} \sum_{m=0}^{M-1} \frac{(2m-1)!!}{(2m)!!} \left\{ \frac{a_0 \rho_c}{\sqrt{1 - \rho_s^2}} \left(\frac{1 - |\rho|^2}{1 - \rho_s^2} \right)^m + \frac{b_0 \rho_s}{\sqrt{1 - \rho_c^2}} \left(\frac{1 - |\rho|^2}{1 - \rho_c^2} \right)^m \right\} \right], & \text{for MRC} \end{cases} \quad (12)$$

where

$$\rho = \rho_c + j\rho_s = e^{-j\theta} \cdot \rho' = e^{-j\theta}$$

$$\frac{\Gamma \int \xi_s(\tau, T) d_R(-\tau) d_R^*(-T - \tau) d\tau + \Gamma/\Lambda \int \xi_i(\tau, T) d_i(\Delta T - t) d_i^*(\Delta T - T - \tau) d\tau + B_n T \xi_n(T)}{\sqrt{\Gamma \int \xi_s(\tau, 0) |d_R(-\tau)|^2 d\tau + \Gamma/\Lambda \int \xi_i(\tau, 0) |d_i(\Delta T - \tau)|^2 d\tau + B_n T}} \cdot \sqrt{\Gamma \int \xi_s(\tau, 0) |d_R(-T - \tau)|^2 d\tau + \Gamma/\Lambda \int \xi_i(\tau, 0) |d_i(\Delta T - T - \tau)|^2 d\tau + B_n T} \quad (13)$$

been shown [13] that for MSK transmissions, the BER performance achievable by two-branch postdetection MRC is only about 0.9 dB inferior to that of predetection MRC in terms of the required average E_b/N_0 .

On the other hand, postdetection SC selects the DD output associated with the branch having the largest input signal envelope, thus $w_m(t) = 1$ if $|z_m(t)|$ is the maximum, zero for other branches. Postdetection SC is of practical interest because switching noise due to abrupt phase change is not produced, and because it is much simpler to implement.

D. BER expression

In (5), co-channel interference was not included. We assume here a single interferer that uses the identical modulation scheme. The interference component is also complex Gaussian in Rayleigh fading when co-channel interference symbol sequence is given. Therefore, inclusion of co-channel interference effect is straightforward. We assume that the symbol $a_0 + jb_0$ is detected. When the phase mapping rule of Fig. 2 is used, a decision on a_0 and b_0 can be made independently based on the polarity of the sampled combiner outputs, $I_0 = I(0)$ and $Q_0 = Q(0)$, respectively. In Rayleigh fading, $z_m(0)$ and $z_m(-T)$ of different branches are independent zero-mean complex Gaussian variables when the desired signal and co-channel interference symbol sequences are given. The method presented in [13], which we apply for BER derivation here, uses the fact that in the Rayleigh fading, with $z_m(0)$ being given, $z_m(-T)$ is a complex Gaussian variable with the variance $\sigma^2(1 - |\rho'|^2)$ and the mean $\rho'^* z_m(0)$, where $\sigma^2 = 1/2(|z_m(0)|^2) = 1/2(|z_m(-T)|^2)$ and $\rho' = 1/2(z_m^*(0) \cdot z_m(-T))/\sigma^2$ for all m ; thus, the combiner output $I_0 + jQ_0$ becomes the complex Gaussian when all $z_m(0)$ are given. Therefore, I_0 and Q_0 are independent Gaussian, whose mean and variance can be determined from (11), and the conditional BER with all $z_m(0)$ being given can be derived. Since all of $|z_m(0)|$ are assumed to be independent and are

with

$$\begin{aligned} d_R(t) &= \int_{-\infty}^{+\infty} d_T(u) h_R(t - u) du \\ &= \sum_{n=-\infty}^{+\infty} e^{j\Phi_n} \frac{\int_{-\infty}^{+\infty} h_T(u - nT) h_R(t - u) du}{\sqrt{\frac{1}{T} \int_{-\infty}^{+\infty} h_T^2(t) dt}} \\ h_R(t) &= \int_{-\infty}^{+\infty} H_R(f) e^{j2\pi f t} df, \\ \xi_n(t) &= \int_{-\infty}^{+\infty} \frac{1}{B_n} |H_R(f)|^2 e^{j2\pi f t} df, \\ B_n &= \int_{-\infty}^{+\infty} |H_R(f)|^2 df, \end{aligned} \quad (14)$$

and $\Gamma = E_s/N_0$ is the average symbol energy-to-noise power spectral density ratio, and Λ is the average signal-to-interference power ratio. $d_i(t)$ is the overall filter response to co-channel interference ($d_i(t)$ is identical to $d_R(t)$ except that a_n and b_n are replaced by the interference symbol sequence), and ΔT is the timing offset because of the assumption that interference modulation timing is independent from desired signal modulation timing, $h_R(t)$ is the baseband equivalent impulse response of the receiver filter, $\xi_n(t)$ is the time correlation function of the bandlimited AWGN, and B_n is the noise bandwidth. $\xi_i(\tau, \mu)$ is the delay-time correlation function for co-channel interference. The effects of ISI produced by the delay spread and by the receiver filter $H_R(f)$ (in the case of Gaussian receiver filter) are included in the expression for ρ in (13).

¹In order to obtain the final results (presented in this paper), (12) must be further averaged over the desired signal and co-channel interference symbol sequences.

E. Approximations

For a low bit rate transmission with a bandwidth narrower than the coherence bandwidth of the multipath channel (or the rms delay spread is much smaller than the symbol duration T), the received signal is subject to multiplicative Rayleigh fading, with errors being produced by the AWGN, co-channel interference, and the random FM noise. If the errors due to random FM noise are not negligible, the fading is called time selective. As the transmission bit rate increases and the modulation bandwidth approaches the coherence bandwidth, the received signal suffers from frequency selective fading. The errors are due to delay spread instead of random FM noise. Average BER's can be approximately represented as the sum of four parts [12], i.e., those due to AWGN, co-channel interference, random FM noise, and delay spread. In the following, to provide good insight into diversity improvements on the cause of errors, we will derive simple approximate expressions for individual average BER's separately.

The average BERs at I and Q rails are identical for equiprobable transmission of di-bit symbols. The approximated BER for SC, EGC, and MRC can be expressed as [13]

$$P_e \approx K_M \left(\frac{1 - |\rho|^2}{2\rho_c^2} \right)^M, \quad (15)$$

where

$$K_M = \frac{(2M-1)!!}{2}, \quad \text{for SC, } \frac{1}{2} \frac{M^M}{M!}, \quad \text{for EGC}$$

$$\text{and } \frac{1}{2} \frac{(2M-1)!!}{M!}, \quad \text{for MRC.} \quad (16)$$

1) *Average BER due to AWGN*: For multiplicative fading, the transfer function of the multipath channel can be assumed to be constant over the bandwidth of interest, i.e., $T_m(f, t) \approx T_m(0, t)$ for $m = 1, 2, \dots, M$, and therefore, $\xi_s(\tau, \mu) \approx \xi_s(\mu) \cdot \delta(\tau)$, where $\xi_s(\mu)$ is the complex fading time correlation function. Average BER, P_{e1} , due to AWGN can be obtained by letting $\xi_s(\mu) \rightarrow 1$ (very slow fading assumption) and $\Lambda \rightarrow \infty$ in ρ of (13) and substituting into (15). We obtain

$$P_{e1} \approx K_M \left[\frac{B_n T}{2\Gamma} \cdot \frac{|d_R(0)|^2 + |d_R(-T)|^2 - 2\xi_n(T) \operatorname{Re} \{e^{-j\theta} d_R(0) d_R^*(-T)\}}{\operatorname{Re}^2 \{e^{-j\theta} d_R(0) d_R^*(-T)\}} \right]^M. \quad (17)$$

We have assumed that $H_R(f)$ is symmetrical with respect to $f = 0$; hence, $\xi_n(T)$ is real.

2) *Average BER due to random FM noise*: For large values of Γ and Λ , the errors are predominantly caused by random FM noise if the fading is multiplicative. Average BER, P_{e2} , due to random FM noise alone can be obtained by letting Γ and $\Lambda \rightarrow \infty$ in ρ , as

$$P_{e2} \approx K_M \left[\frac{1 - |\xi_s(T)|^2}{2} \cdot \frac{|d_R(0) d_R^*(-T)|^2}{\operatorname{Re}^2 \{e^{-j\theta} d_R(0) d_R^*(-T)\}} \right]^M. \quad (18)$$

3) *Average BER due to Co-channel Interference*: We assume that fading is very slow so that the random FM noise effect is negligible. An average BER, P_{e3} , to co-channel interference alone can be obtained by letting $\xi_s(T) \rightarrow 1$ and $\Gamma \rightarrow \infty$ in ρ , as

$$P_{e3} \approx K_M \left[\frac{1}{2\Lambda} \cdot \frac{|d_R(0)|^2 |d_i(\Delta T - T)|^2 + |d_R(-T)|^2 |d_i(\Delta T)|^2 - 2 \operatorname{Re} \{d_R(0) d_R^*(-T) d_i^*(\Delta T) d_i(\Delta T - T)\}}{\operatorname{Re}^2 \{e^{-j\theta} d_R(0) d_R^*(-T)\}} \right]^M. \quad (19)$$

4) *Average BER due to Delay Spread*: For frequency-selective fading, the time variations in multipath channel impulse response can be assumed to be negligible over T s and thus $\xi_s(\tau, \mu) \approx \xi_s(\tau, 0)$. Letting Γ and $\Lambda \rightarrow \infty$ in ρ , the average BER, P_{e4} , due to delay spread alone can be approximated as [13]

$$P_{e4} \approx K_M \left[T \frac{|d'_R(0) d_R(-T) - d_R(0) d'_R(-T)|}{\sqrt{2} \operatorname{Re} \{e^{-j\theta} d_R(0) d_R^*(-T)\}} \cdot \frac{\tau_{rms}}{T} \right]^{2M}, \quad (20)$$

where the prime denotes time derivative and τ_{rms} is given by (8).

F. Discussion

Equations (17) and (19) show that average BER due to AWGN (co-channel interference) decreases in inverse proportion to M th power of average E_s/N_0 (SIR). The required E_s/N_0 or SIR necessary to achieve a certain BER is the smallest for MRC and is about 1.5 dB larger for SC when $M = 2$ (this can be seen from examining (16)). The performance achieved by the use of EGC is between those of MRC and SC, and is about 0.9 dB superior to SC.

Average BER due to random FM noise (in time selective fading) depends on the fading time correlation $\xi_s(\mu)$ at $\mu = T$, where $\xi_s(\mu)$ is the inverse Fourier transform of the power spectrum $W(f)$ of $T_m(0, t)$:

$$\xi_s(T) = \int_{-\infty}^{+\infty} W(f) \exp(j2\pi fT) df. \quad (21)$$

If the fading is much slower compared to the symbol rate $1/T$, $\exp(j2\pi fT)$ can be expanded at the vicinity of $f = 0$, and we have

$$|\xi_s(T)|^2 \approx 1 - (2\pi T)^2 \int_{-\infty}^{+\infty} (f - f_{\text{mean}})^2 W(f) df$$

$$= 1 - (2\pi f_{rms} T)^2, \quad (22)$$

where f_{rms} is the rms Doppler frequency, and f_{mean} is the mean

Doppler frequency defined as

$$f_{\text{mean}} = \int_{-\infty}^{+\infty} f \cdot W(f) df. \quad (23)$$

We have assumed $\int W(f) df = 1$ without loss of generality. Substituting (22) into (18) shows that average BER is proportional to $2M$ th power of the normalized rms Doppler frequency $f_{rms}T$ and does not depend on the shape of the fading power spectrum if $f_{rms}T \ll 1$.

Commonly used measure of the degree of frequency selectivity of the channel is the rms delay spread. Equation (20) implies that BER performance in the frequency selective fading depends on the value of τ_{rms}/T and the shape of the delay profile is not important if $\tau_{rms}/T \ll 1$. (Also indicated is that the performance is sensitive to the overall modulation pulse shape intro-

duced by the combination of transmitter and receiver filters. BER dependence on pulse shape will be discussed in Section III.) Frequency selectivity can also be measured by the frequency correlation function, that is defined as $\Omega(\Delta f) = \langle T_m(f + \Delta f, t) \cdot T_m^*(f, t) \rangle$ with $\Omega(0) = 1$. The frequency correlation function is the Fourier transform of the delay profile:

$$\Omega(\Delta f) = \int_{-\infty}^{+\infty} \xi_s(\tau, 0) \exp(-j2\pi\Delta f\tau) d\tau. \quad (24)$$

Expansion of $\exp(-j2\pi\Delta f\tau)$ at the vicinity of $\tau = 0$ leads to

$$\begin{aligned} |\Omega(\Delta f)|^2 &\approx 1 - (2\pi\Delta f)^2 \int_{-\infty}^{+\infty} (\tau - \tau_{\text{mean}})^2 \xi_s(\tau, 0) d\tau \\ &= 1 - (2\pi\Delta f\tau_{\text{rms}})^2. \end{aligned} \quad (25)$$

If $\Delta f = 1/T$ is not very close to the coherence bandwidth of the multipath channel (which is often defined as the frequency displacement that gives $|\Omega(\Delta f)| = 0.5$), (25) is a good approximation. Substitution of (25) into (20) reveals that average BER due to delay spread is proportional to M th power of $1 - |\Omega(1/T)|^2$. Equation (18) shows that in time selective fading, average BER is proportional to M th power of $1 - |\xi_s(T)|^2$. As a consequence, the BER's in time selective fading (random FM noise as a cause of errors) and in frequency selective fading (delay spread as a cause of errors) can be expressed in a similar formula; the former is a function of the time correlation (or rms Doppler frequency) while the latter is a function of the frequency correlation (or rms delay spread).

III. NUMERICAL RESULTS

We first calculate BER performance for the case of square-root RC filter at the transmitter and receiver. Secondly, the calculated BER performance is shown for the Gaussian receiver filter case (the transmitter has RC filter). The exact calculations using (12) are presented for individual average BER due to AWGN, cochannel interference, random FM noise, and delay spread. The results for EGC are not presented. It can be seen from (16) that the average BERs for EGC are between those for SC and MRC and approximately $M^M/(2M-1)!!$ times larger than those for MRC.

A. Square Root RC Filter Case

In this case, $H_T(f) = T \cdot H(f)$, $H_R(f) = H(f)$ and the convolution of $h_T(t)$ and $h_R(t)$ in (14) is

$$\begin{aligned} &\int_{-\infty}^{+\infty} h_T(u - nT) h_R(t - u) du \\ &= \frac{\sin(\pi(t - nT)/T)}{\pi(t - nT)/T} \frac{\cos(\alpha\pi(t - nT)/T)}{1 - (2\alpha(t - nT)/T)^2}. \end{aligned} \quad (26)$$

Noise bandwidth $B_n = 1/T$, and noise correlation $\xi_n(T) = 0$. Figs. 4, 5, and 6 show the calculated BER performance due to AWGN, co-channel interference, and random FM noise, respectively.

In Fig. 4, BER performance due to AWGN is plotted as a function of the average signal energy per bit-to-noise power spectral density ratio $E_b/N_0 = (0.5E_s/N_0)$. Both RC-QDPSK and $\pi/4$ -shift RC-QDPSK exhibit identical performance because ISI free transmission is satisfied for both modulation cases. The BER's are not affected by the roll-off factor α of the filter. When $M = 2$, diversity gain (defined as the reduced value of

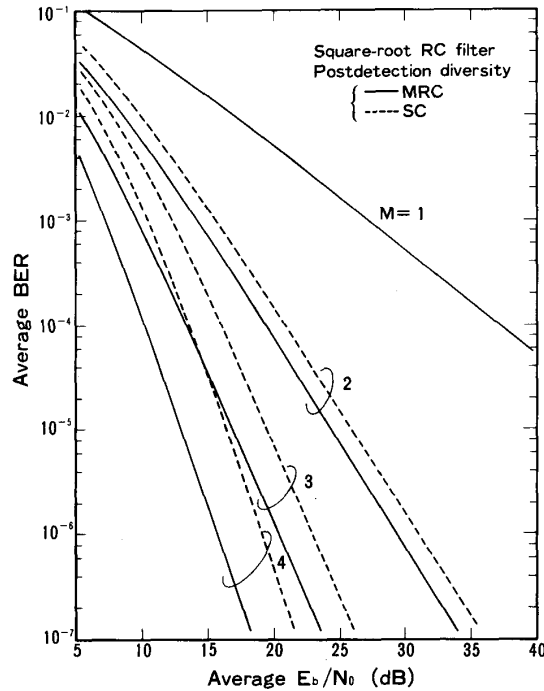


Fig. 4. Average BER due to AWGN for $(\pi/4)$ -shift RC-QDPSK using square-root RC filter.

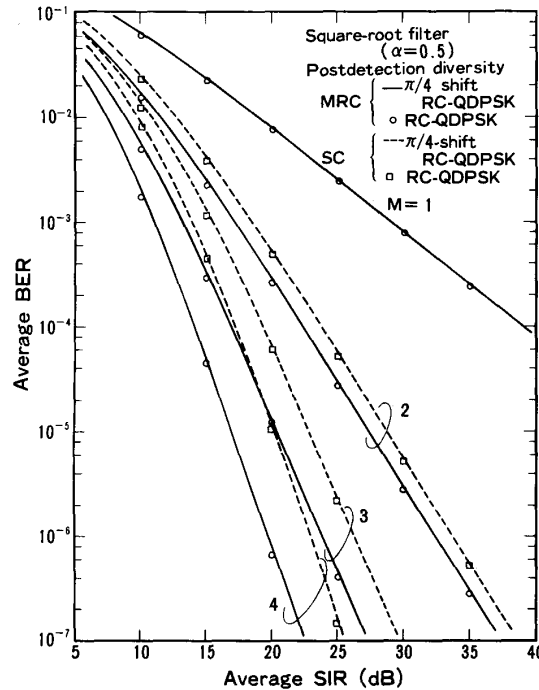


Fig. 5. Average BER due to cochannel interference for $(\pi/4)$ -shift RC-QDPSK using square-root RC filter ($\alpha = 0.5$).

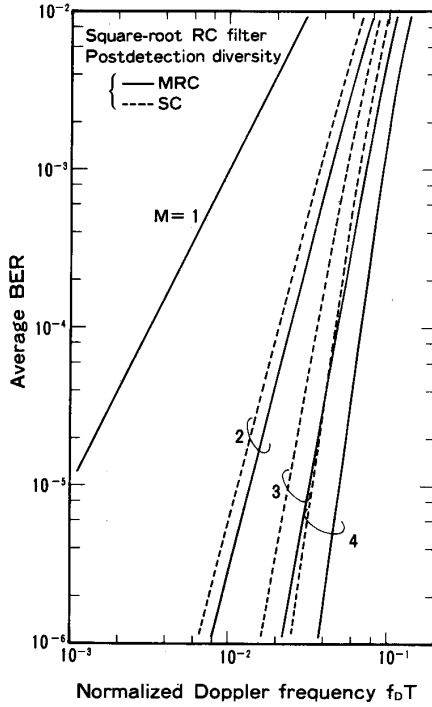


Fig. 6. Average BER due to random FM noise for $(\pi/4\text{-shift})$ RC-QDPSK using square root RC filter.

E_b/N_0 necessary to achieve a certain BER) of 7 dB for SC and 8.5 dB for MRC can be obtained at an average BER of 10^{-2} .

BER performance due to co-channel interference depends on the roll-off factor α and symbol sequence of the co-channel interference. In Fig. 5, $\alpha = 0.5$ was assumed. Sampling time offset ΔT was assumed to be uniformly distributed within $|\Delta T| \leq T/2$, thus, ISI from the co-channel interference component was considered. We took into account ISI from two symbols on each side (five-symbol sequence of the co-channel interference). The BER for each symbol sequence is first calculated using (12), then, the BER's are averaged over all sequence patterns. The BER performance of RC-QDPSK is slightly superior to that of $\pi/4\text{-shift}$ RC-QDPSK. This is because envelope variations of $\pi/4\text{-shift}$ RC-QDPSK signal are less than those of RC-QDPSK signal, and thus the slightly increased interference is produced under the same power condition. The BER performance due to cochannel interference are not much different from that due to AWGN. Performance similarity is because of Rayleigh fading, most errors are produced when the desired signal fades, irrespective of their cause (AWGN or co-channel interference).

For the evaluation of average BER due to random FM noise, we assume $\xi_s(\mu) = J_0(2\pi f_D \mu)$, where $J_0(\cdot)$ is the Bessel function and f_D is the maximum Doppler frequency (vehicle speed/carrier wavelength) [12]. According to Section II-F, the average BER depends on the rms Doppler frequency f_{rms} . In the above case, $f_{rms}T = f_D T/\sqrt{2}$. In Fig. 6, however, the BER's are plotted as a function of $f_D T$ for convenience. As in the case of AWGN channels, the two modulation schemes exhibit identical performance. For a bit rate of 8 kb/s and a vehicle speed of 48 km/h at a carrier frequency of 900 MHz, $f_D T = 1 \times 10^{-2}$. Average BER is below 10^{-5} when $M = 2$. As the bit rate

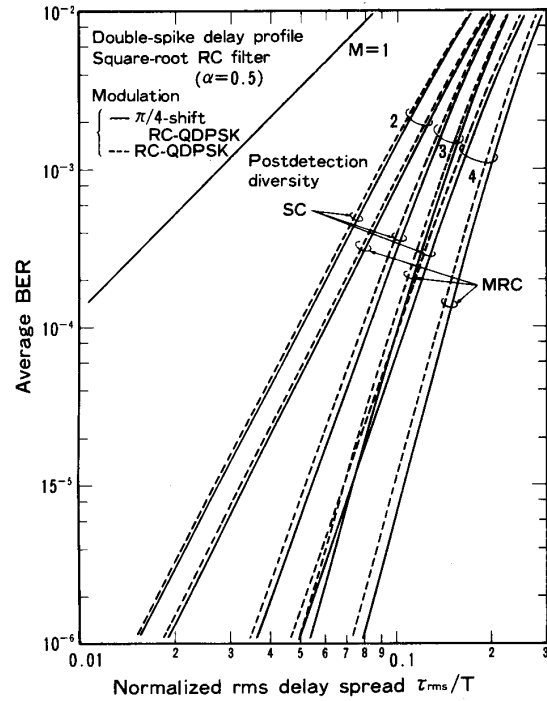


Fig. 7. Average BER due to delay spread for $(\pi/4\text{-shift})$ RC-QDPSK using square-root RC filter ($\alpha = 0.5$). Double-spike delay profile.

increases, random FM noise becomes less. For example, average BER at 64 kb/s can be made below 10^{-8} even when $M = 2$.

When a squaring timing recovery loop, whose bandwidth is much larger than f_D , is used, the recovered timing can be assumed to track the variations in the mean delay (or the first central moment of the delay profile) if $\tau_{rms}/T \ll 1$ [21]. Therefore, from this assumption, we assume $\tau_{mean} = 0$ so that sampling timing is $t = nT(\dots, -2, -1, 0, 1, 2, \dots)$ for calculation without loss of generality. A simple model of a delay profile may be a double-spike. Another widely used model is the Gaussian profile. However, according to mobile propagation measurements in urban areas [7], a one-sided exponential profile sometimes with several spikes exists. In this paper, to examine the effects of the delay profile shape, a one-sided exponential profile, a double-spike profile, and a Gaussian profile are used, and treated separately. The three delay profiles considered here are:

$$\xi_s(\tau, 0) = \begin{cases} \frac{1}{\tau_{rms}} \exp[-(\tau + \tau_{rms})/\tau_{rms}] & n \tau \geq -\tau_{rms} \text{ (one-sided exponential)} \\ \frac{1}{2} \delta(\tau + \tau_{rms}) + \frac{1}{2} \delta(\tau - \tau_{rms}) & \text{(double spike)} \\ \frac{1}{\sqrt{2\pi} \tau_{rms}} \exp[-\tau^2/2\tau_{rms}^2] & \text{(Gaussian)} \end{cases} \quad (27)$$

The average BER using postdetection SC and MRC is shown in Fig. 7 as a function of τ_{rms}/T for a double-spike delay profile. For the calculation of average BER, we consider two adjacent symbols on each side for ISI produced by delay spread (the calculated BER using (12) is averaged over five-symbol sequence of the desired signal). Comparison of the two modulation schemes shows that $\pi/4$ -shift RC-QDPSK is slightly superior to RC-QDPSK; however, the difference is very small. Particularly, without diversity ($M = 1$), the two modulations exhibit identical performance. The use of diversity can enlarge the allowable normalized rms delay spread that can achieve a certain BER; for $\pi/4$ -shift RC-QDPSK, the allowable value of τ_{rms}/T for $BER = 10^{-3}$ can be increased to 0.11 (corresponding to $3.4 \mu s$ for 64 kb/s) with two branch MRC, while this value is 0.028 without diversity.

Since the received signal is composed of many multipath components with different time delays, the modulation phase of each multipath component at the sampling instant slightly differs if $\tau_{rms}/T \ll 1$. When the received signal is in a deep fade (or multipath components are added destructively), a slight change of the relative envelope and phase of different multipath components produces large phase fluctuations. Therefore, as the envelope of the received signal becomes smaller, errors are produced. As described earlier, the postdetection diversity combiner weights each DD output before combination so as to reduce the contribution of the branch with faded signal. Hence, diversity reception can be effective. However, as the rms delay spread approaches the symbol duration, the modulation phase difference among multipath components at the sampling instant becomes large and thus the resultant phase fluctuations become significant even for large received signal envelopes. Therefore, diversity improvements diminish. It can be concluded from Fig. 7 that diversity reception can effectively work if $\tau_{rms}/T < 0.2 \sim 0.3$. Approximate BER's were also calculated using (20), and compared with the exact result. It was found that a good approximation is obtained if $\tau_{rms}/T < 0.1$.

Average BER's using MRC for two other types of delay profile are also calculated and compared in Fig. 8 for $\pi/4$ -shift RC-QDPSK. It can be seen, as was anticipated from the approximate expression in (20), that BER performance strongly depends on τ_{rms}/T and the profile shape has a negligible impact if $\tau_{rms}/T < 0.1$. When τ_{rms}/T becomes larger, however, the BER curves increasingly diverge. The performance curves are the best (worst) for the one-sided exponential (double-spike) delay profile, as was seen in [10].

Glance and Greenstein [10] and Chuang [11] have shown that average BER can be reduced by increasing the value of α (or increasing the modulation bandwidth). The reason for this BER reduction is that the overall modulation pulse response converges more rapidly as α increases and thus ISI caused by delay spread becomes less. The effect of the roll-off factor is shown in Fig. 9 for $\tau_{rms}/T = 0.1$. BER dependence on α , similar to [11], is observed, although differential detection is used instead of coherent detection (because of this, average BER is slightly larger). As α increases to 1.0 from 0.5 (bandwidth increases 1.3 times), average BER with two-branch MRC can be reduced to 3×10^{-5} from 7.6×10^{-4} for $\pi/4$ -shift RC-QDPSK.

B. Gaussian Receiver Filter Case

We assume that the transmitter filter $H_T(f) = T \cdot H^2(f)$ and that the receiver filter $H_R(f)$ has a Gaussian transfer function with a 3 dB bandwidth of B . In this case, $h_R(t)$ and

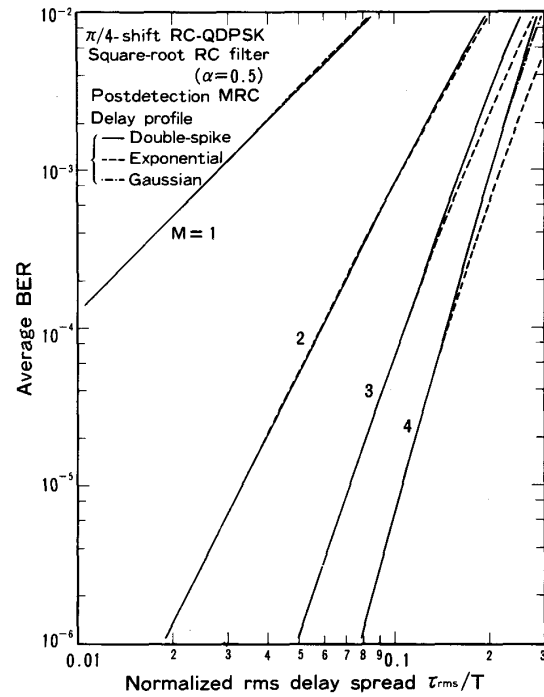


Fig. 8. Effect of delay profile for $\pi/4$ -shift RC-QDPSK using square-root RC filter ($\alpha = 0.5$). Postdetection MRC.

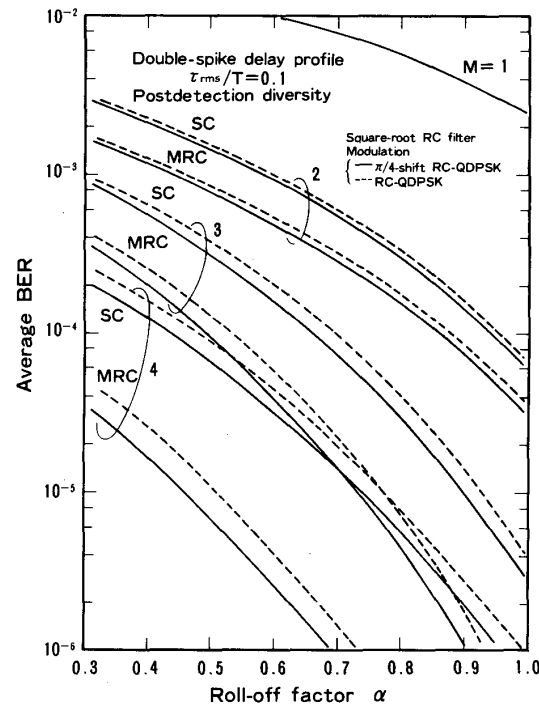


Fig. 9. Effect of roll-off factor on average BER due to delay spread for ($\pi/4$ -shift) RC-QDPSK using square root RC filter. Double-spike delay profile.

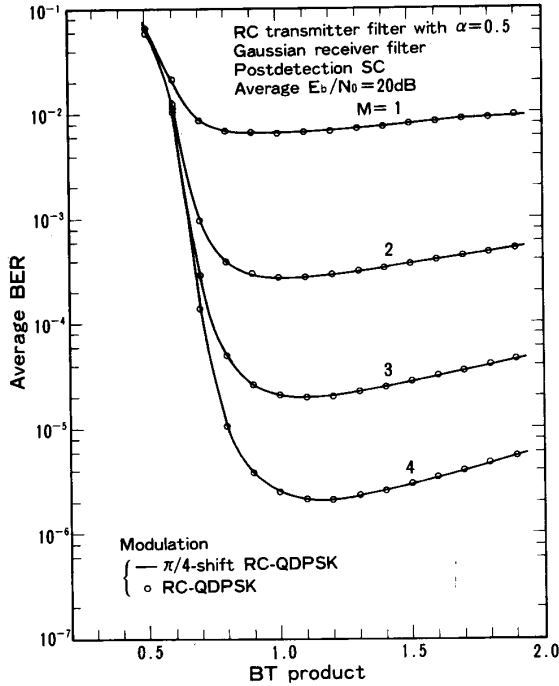


Fig. 10. Effect of BT product of Gaussian receiver filter on average BER due to AWGN with postdetection SC. RC transmitter filter with $\alpha = 0.5$.

$\xi_n(T)$ are given by

$$h_R(t) = \frac{\beta}{\sqrt{\pi} T} \exp \left[-\beta^2 \left(\frac{t}{T} \right)^2 \right],$$

$$\xi_n(t) = \exp \left[-\frac{\beta^2}{2} \left(\frac{t}{T} \right)^2 \right], \quad (28)$$

where $\beta = \pi BT / \sqrt{2 \ln 2} = \sqrt{2\pi} B_n T$. If too narrow a receiver filter is used, the ISI produced by the receiver filter degrades BER performance. On the other hand, as the receiver filter bandwidth becomes much wider, ISI free transmission is possible, but the increased noise power degrades performance. Hence, the optimum BT product exists.

We assume SC for the diversity combining because it is of much more practical interest than other combiners. For the calculation of $d_R(t)$, it is sufficient to take into account two adjacent symbols on each side for a receiver filter BT product > 0.5 . The calculated BER using (12) is averaged over five-symbol sequence of the desired signal. The effect of the receiver BT product is shown in Fig. 10 on average BER due to AWGN for an average E_b/N_0 of 20 dB. Because of relatively flat minimum, a BT product of 1.0 can be used as far as AWGN is concerned.

In an interference-limited cellular system, as the BT product becomes smaller, average BER increases. Careful selection of the BT product is required in order not to degrade the cochannel interference performance. How the receiver BT product affects average BER due to co-channel interference is shown in Fig. 11 for an average SIR of 20 dB. The degradation of the performance at BT = 1.0 is small, and therefore, we assume BT = 1.0 in the subsequent calculations.

Fig. 12 and 13 show BER performance due to AWGN and due to co-channel interference, respectively. Both RC-QDPSK

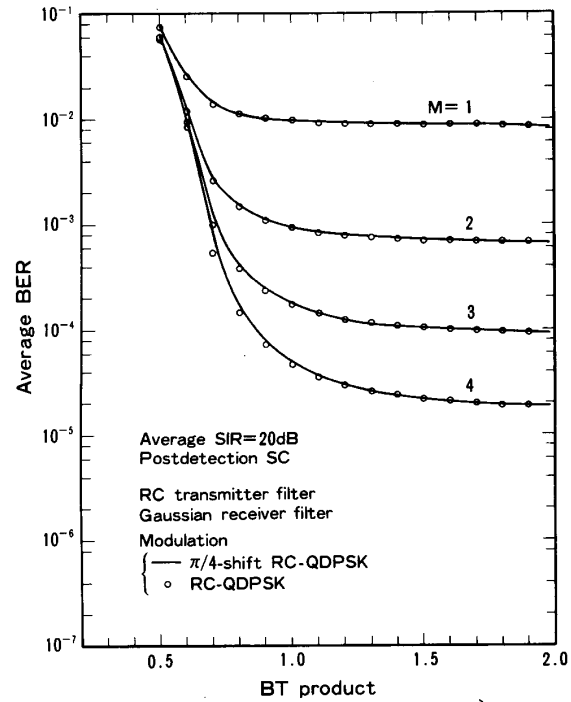


Fig. 11. Effect of BT product of Gaussian receiver filter on average BER due to co-channel interference with postdetection SC. RC transmitter filter with $\alpha = 0.5$.

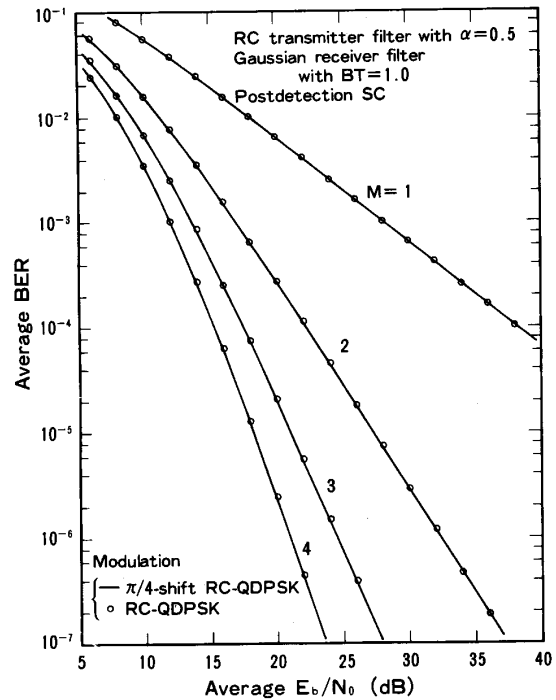


Fig. 12. Average BER due to AWGN for $(\pi/4)$ -shift RC-QDPSK with postdetection SC. RC transmitter filter with $\alpha = 0.5$ and Gaussian receiver filter with BT = 1.0.

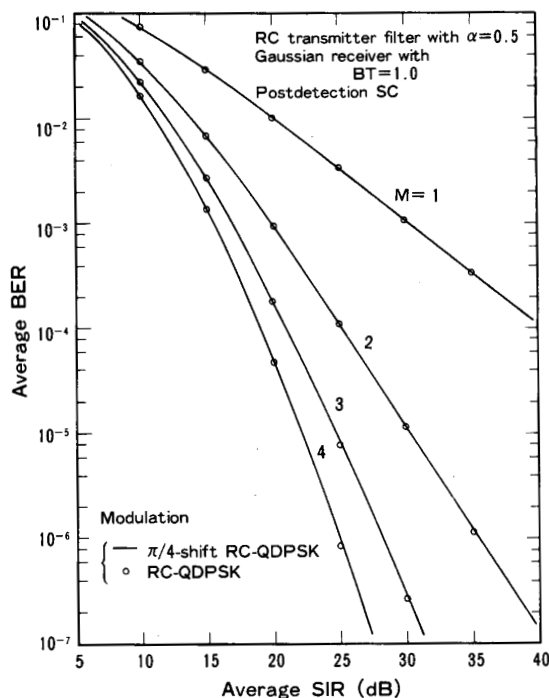


Fig. 13. Average BER due to co-channel interference for ($\pi/4$ -shift) RC-QDPSK with postdetection SC. RC transmitter filter with $\alpha = 0.5$ and Gaussian receiver filter with $BT = 1.0$.

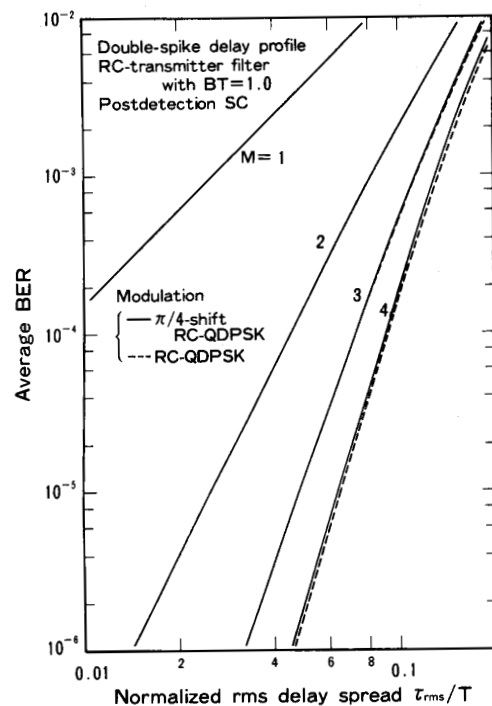


Fig. 14. Average BER due to delay spread for ($\pi/4$ -shift) RC-QDPSK with postdetection SC. RC transmitter filter with $\alpha = 0.5$ and Gaussian receiver filter with $BT = 1.0$. Double-spike delay profile.

and $\pi/4$ -shift RC-QDPSK exhibit almost identical performance. In the Gaussian receiver filter case, because of the ISI introduced by receiver filter, the performance is degraded compared to the square root RC filter case. However, comparison of Figs. 12 and 13 with Figs. 4 and 5 shows that the performance degradation is about 1.5 dB both in E_b/N_0 and in SIR. The effect of random FM noise was also calculated, and it was found that the degradation is very small and the same BER reached only at about 5% smaller $f_d T$ value.

The dependence of average BER on rms delay spread is shown in Fig. 14 for $\alpha = 0.5$. Double-spike delay profile was assumed. In the square-root RC filter case, $\pi/4$ -shift RC-QDPSK was found to achieve slightly superior performance with diversity reception. However, in the Gaussian receiver filter case, noticeable performance difference is not seen. Performance is slightly degraded from the square root RC filter case and the same BER is reached with about 10% smaller rms delay spread. Average BER cannot be reduced by increasing roll-off factor α as much as in the square root RC filter case, because of the ISI introduced by the receiver filter. This is shown in Fig. 15.

IV. COMBINED DIVERSITY RECEPTION/ERROR CONTROL

Error control techniques, such as FEC and ARQ, are other powerful means for combating multipath fading [16], [17]. In this section, we investigate the combination of postdetection diversity reception and error control assuming $\pi/4$ -shift RC-QDPSK signal transmission.

A. FEC

From an implementation point of view, we consider a relatively short block FEC. BCH(63, k) codes ($k = 57, 51, 46, 36$,

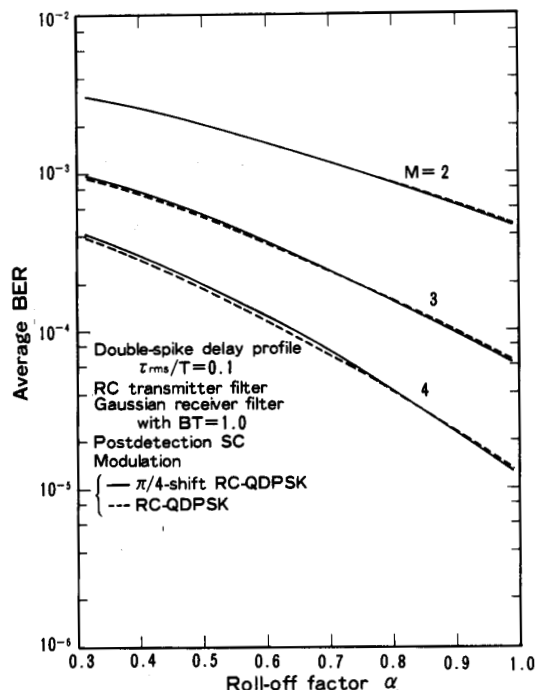


Fig. 15. Effect of roll-off factor on average BER due to delay spread for ($\pi/4$ -shift) RC-QDPSK with postdetection SC. RC transmitter filter and Gaussian receiver filter with $BT = 1.0$. Double-spike delay profile.

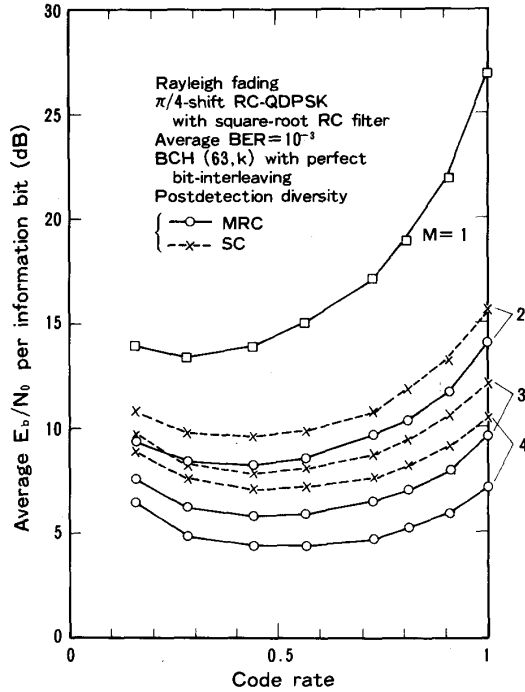


Fig. 16. Power efficiency of combined diversity reception/FEC for $\pi/4$ -shift RC-QDPSK using square root RC filter.

28, 18 and 10) are assumed. FEC can reduce the required E_b/N_0 necessary to achieve a certain BER at the cost of radio channel bandwidth expansion. In cellular mobile radio, however, improving spectrum efficiency as well as power efficiency is an important issue. How diversity reception affects the optimum code rate that minimizes the required E_b/N_0 or maximizes the spectrum efficiency is investigated.

For an (n, k) block code with minimum code-distance d and error correction capability $t = (d - 1)/2$, the decoded BER can be approximately calculated from [22]

$$P_{ed} \approx \frac{d}{n} \sum_{m=t+1}^d p(m, n) + \frac{1}{n} \sum_{m=d+1}^n mp(m, n), \quad (29)$$

where $p(m, n)$ is the probability of m bit errors occurring in an n -bit block. Assuming perfect bit-interleaving, $p(m, n)$ can be represented by a binomial distribution with a BER of P_e determined from (12).

The average channel E_b/N_0 necessary to obtain a certain decoded BER can be calculated by letting $f_D T \rightarrow 0$ and $\Lambda \rightarrow \infty$. Thus, the average E_b/N_0 per information bit is obtained as $(n/k) \times$ channel E_b/N_0 . An example of calculated values is shown in Fig. 16 for decoded BER = 10^{-3} with M as a parameter. The optimum code rate is found to be 0.29 ($k = 18$) for no diversity reception ($M = 1$). A significant coding gain of about 14 dB is observed. Combined with two-branch ($M = 2$) diversity reception, the optimum code rate becomes 0.44 ($k = 28$) and about 19 dB of overall gain can be achieved for MRC. However, selection of code rate is not too critical because of the relatively flat minimum.

The spectrum efficiency η of a cellular system can be represented by the product of spectrum efficiencies with respect to

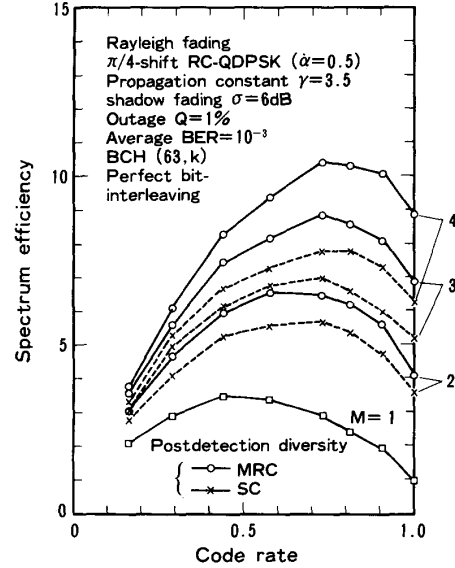


Fig. 17. Spectrum efficiency of combined diversity reception/FEC for $\pi/4$ -shift RC-QDPSK using square root RC filter ($\alpha = 0.5$).

time, frequency and space [18], i.e., $\eta = \eta_t \cdot \eta_f \cdot \eta_s$. The radio channel bandwidth (the transmission time-duration) increases by a factor of n/k if the information bit rate (the radio channel bandwidth) remains unchanged. Therefore, the value of $\eta_t \cdot \eta_f$ decreases by a factor of k/n . The value of η_s is inversely proportional to the number N of radio channel sets necessary to cover the entire service area. For a hexagonal cell layout, $N = 1/3(D/R)^2$, where R is the cell radius and D is the distance between the adjacent co-channel cells [12]. We consider the cell fringe to be where the worst co-channel interference is produced. Only the nearest interferer is considered for simplicity of calculation. Let Λ_c be the average SIR necessary to achieve a certain decoded BER when $f_D T \rightarrow 0$ and $\Gamma \rightarrow \infty$, D/R is obtained as $D/R = 1 + (\epsilon \cdot \Lambda_c)^{1/\gamma}$, where γ is the propagation constant (typical value of γ is 3.5 [23]) and ϵ is the fading margin for the allowable probability Q of geographical outage due to shadowing. Since both average powers (averaging is taken over a distance of several tens of carrier wavelengths) of desired signal and co-channel interference can be assumed to suffer independent log-normal shadowing with identical standard deviation σ , the variation of average SIR expressed in decibels follows a Gaussian distribution with a standard deviation of $\sqrt{2}\sigma$. Hence, the value of ϵ can be easily calculated from $Q = 1/2\text{erfc}(\epsilon/2\sigma)$. As a result, the spectrum efficiency can be determined from

$$\eta \propto \frac{k}{n} \{1 + (\epsilon \cdot \Lambda_c)^{1/\gamma}\}^{-2}. \quad (30)$$

An example of the calculated spectrum efficiencies is shown in Fig. 17 for decoded BER = 10^{-3} , $Q = 1\%$, $\gamma = 3.5$, and $\sigma = 6$ dB. Spectrum efficiency is normalized by that without diversity reception ($M = 1$) and FEC. The optimum code rate is 0.44 ($k = 28$) for no diversity reception and 0.73 ($k = 46$) for diversity reception. Spectrum efficiency with combined diversity reception/FEC is less sensitive to the code rate as was seen in the case of power efficiency. It was found [19] that the optimum code rate is almost independent of Q . This is because (30) can

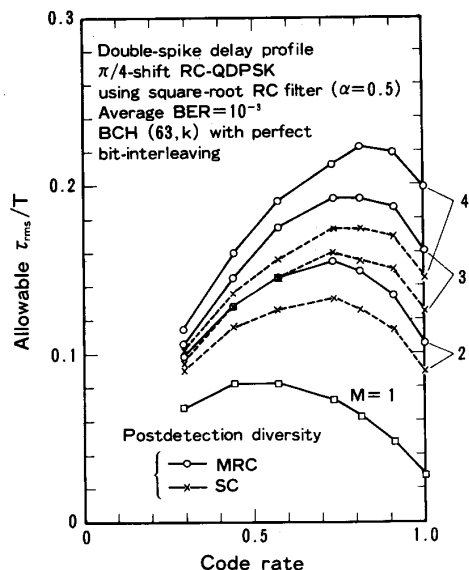


Fig. 18. Allowable of rms delay spread of combined diversity reception/FEC for $\pi/4$ -shift RC-QPSK using square root RC filter ($\alpha = 0.5$).

be approximated as $\eta \propto \epsilon^{-2/\gamma} \cdot (k/n) \cdot \Lambda_c^{-2/\gamma}$ and ϵ is not a function of the code rate.

As the code rate decreases for a given information bit rate, the channel bit rate increases, thus increasing average BER due to delay spread. However, the error correction power becomes larger. This means that there exists an optimum code rate that maximizes the allowable rms delay spread. Calculated result for an average BER of 10^{-3} is shown in Fig. 18. The combination of diversity reception and FEC can enlarge the allowable rms delay spread. When two-branch MRC is used, a code rate of 0.73 ($k = 46$) is optimum, and an allowable value of $\tau_{rms}/T = 0.15$ is achieved, which is about 1.4 times larger than without coding. Compared to the case without diversity and coding, the allowable rms delay spread can be 5.4 times larger.

We also calculated performance using BCH(31, k) codes and found that similar, but slightly degraded performance is obtained.

B. ARQ

The simplest implementation of ARQ schemes (referred to as the basic ARQ scheme) uses only error detection. Detected errors lead to a request for retransmission and discarding of information in erroneous blocks. The main drawback with such a basic ARQ scheme is an increase, due to fading, in the number of transmissions needed before receiving a block correctly [16]. To enhance the performance, a time diversity ARQ (TD-ARQ) scheme [24] was proposed which combines all retransmitted data blocks to construct a new reliable block by postdetection diversity combining. The number of transmissions is limited to avoid a large number of transmissions due to shadow fading. For simplicity, we assume that no error correcting code is used and that error detection code will detect all errors. Performance is compared based on the required E_b/N_0 taking into account the multiple retransmission necessary to achieve a certain BER, which is defined as

$$E_b/N_0 = \text{channel } E_b/N_0 \times \text{average number of transmissions.} \quad (31)$$

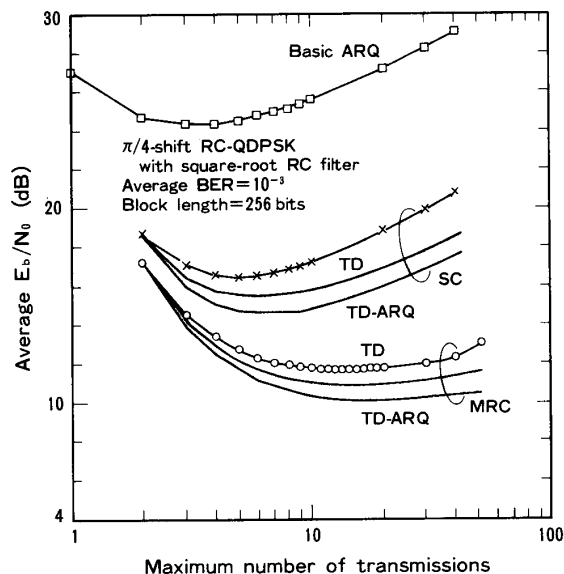


Fig. 19. Required average E_b/N_0 of TD-ARQ for $\pi/4$ -shift RC-QPSK using square root RC filter.

The average number of transmissions and the BER performance to calculate the required channel E_b/N_0 can be bound-estimated for a given maximum number M of transmissions. The calculated results are shown in Fig. 19 for BER = 10^{-3} and block length $n = 256$ bits. The results for the TD-ARQ scheme are presented by the shaded area, the upper (lower) curves of which are obtained using the upper (lower) bounds for the average number of transmissions and the BER performance. For a comparison, also shown is the performance of the TD scheme (the same data block is always transmitted M times and M received blocks are combined using postdetection diversity combining described in this paper). TD-ARQ using MRC offers the best performance; the minimum required E_b/N_0 is achieved around $M = 16$ and is about 11 dB, while the basic ARQ scheme requires a much larger value (24 dB achieved at $M = 4$). Because of a broad minimum, an E_b/N_0 close to the minimum can be attained with a small M .

V. CONCLUSION

BER performance of ($\pi/4$ -shift) RC-QPSK with postdetection diversity reception has been theoretically analyzed for mobile radio channels suffering from Rayleigh fading. Diversity reception is a powerful technique to reduce the effect of delay spread if $\tau_{rms}/T \ll 1$. The shape of the delay profile is of no importance and BER performance strongly depends on the value of τ_{rms}/T .

The effects of combined diversity reception/error control have also been analyzed. The optimum code rate for short BCH codes is around 0.7 when diversity reception is used. TD-ARQ, which combines all retransmitted data blocks by postdetection diversity combining, can achieve a significant improvement over the basic ARQ scheme.

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