Evaluation on MRC Diversity Reception of *M*-ary PSK and DPSK Signals on Wireless Fading Channels

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Abstract

The performances of *M*-ary PSK(MPSK) systems using *L*-branch maximum ratio combining(MRC) diversity reception in frequency-nonselective slow Rician fading channels are derived theoretically. Especially we investigate the effect of number of symbols on the difference between the approximation to bit error probability and the exact bit error rate(BER) in evaluating the performance of MPSK signals through numerical analyses. On the other hand, when *M*-ary DPSK(MDPSK) signals experience the Rician fading channels, the general formula for evaluating BER of MDPSK signals in the independent branch diversity system is presented using the integral-form expressions.

Key words : MRC, Ricain Fading, MPSK, MDPSK.

Ⅰ. Introduction

The statistical properties of mobile radio environments can be often specified by three propagation effects: 1) short-term fading, 2) long-term fading, 3) propagation path loss[1]. In short-term fading, the scattering mechanism only results in numerous reflected components[2]. The Rayleigh model is used to characterize this fading in small geographical areas and sometimes does not account for large scale effects like shadowing by building and hills[3]. In long-term fading, the change of effective height for mobile communication antenna exists due to the nature of the terrain. Its statistics follow the log-normal distribution. But the Rician model can be obtained from the direct wave and its scattering components, and both wave carry information[4]. By modeling the channel as a Rician fading channel, a result is obtained that is valid in the limit of large direct-to-diffuse power ratios for channels with no fading and in the limit of small direct-to-diffuse power ratios for Rayleigh environments as well as for the general case when neither the direct nor the diffuse components of the signal are negligible[5]. When Rician factor k is 0, the error performances lead to those of Rayleigh fading model. In a celluar system, Rayleigh fading is often the feature of large cells, whereas for the cells of smaller field, the envelope fluctuations of a received signal are closer to the Rician fading that is bounded by AWGN perturbations and Rayleigh fading[6].

An alternative solution to the problem of obtaining acceptable performances on a fading channel is the diversity technique, which is widely used to combat the fading effects of time-variant channels. When *M*-ary signals experience the fading channels, diversity schemes can minimize the effects of these fadings since deep fades seldom occur simultaneously during the same time intervals on two or more paths.

In this paper we can represent the average symbol error rate(SER) by MRC systems in receiving MPSK signals on Rician fading channels. We find the error performance of coherent MPSK over the slow and flat fading channels when additive white Gaussian noise (AWGN) is present, using the approximation, an upper bound on the bit error probability for large values of M signal waveforms. We then compare numerical analyses for the effect of increasing M on the difference between the approximation with the closed-form and the exact BER with the integral-form in the absence of the diversity branches. Next we present the performance of MRC diversity reception of MDPSK signals in slowly frequency-nonselective Rician fading channels with an AWGN. These performance evaluations allow designers to determine *M*-ary modulation methods for Rician fading environments. The analytical results presented in this paper are expected to provide important informations in designing radio systems under the fading channels.

Ⅱ. System Model with MRC Diversity Reception

We assume that there are L diversity branches in the frequency-nonselective and slow fadings, carrying the

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same information-bearing signal. The fading processes among L diversity branches are assumed to be mutually statistically independent. The signal in each channel is corrupted by an additive zero-mean white Gaussian noise.

The equivalent low-pass received signals for L branches can be expressed in the form

$$r_i(t) = \alpha_i e^{-j\phi_i} d_i(t) + n_i(t), \qquad i = 1, 2, \cdots, L \tag{1}$$

where $\{\alpha_i e^{-j\phi_i}\}$ represent the attenuation factors and phase shifts for L branches, $d_i(t)$ denotes the transmitted signal on the ith branch, and $n_i(t)$ denotes the AWGN on the ith branch. All signals in the set $\{d_i(t)\}$ have the same energy.

The optimum demodulator for the signal received from the ith branch has the impulse response

$$h_i(t) = d_i^*(T-t), \qquad 0 \le t \le T \tag{2}$$

where * denotes the complex conjugate.

The combiner that achieves the best performance is one in which each matched filter output is multiplied by the corresponding complex-valued (conjugate) branch gain $\alpha_i e^{j\phi_i}$. The effect of this multiplication is to cophase and to weight the low-pass received M-ary signal before being combined such that the weighted sum at the combiner output has the maximum signal-to-noise(SNR).

In spite of the complexity of MRC compared to other diversity techniques since it requires the knowledge of a fading amplitude in each signal branch, it is worth considering because it has the maximum possible improvement that a diversity system can attain through a fading channel.

Ⅲ. Error Rate Analysis

If each branch has equal fading parameter and average SNR γ_0, the conditional probability density function(PDF) of the instantaneous SNR r at the output of L-branch MRC on a Rician fading channel is [7]

$$f(\gamma) = \left(\frac{K+1}{\gamma_0}\right)^{\frac{L+1}{2}} \left(\frac{\gamma}{KL}\right)^{\frac{L-1}{2}} \cdot \exp\left[-KL - \frac{(K+1)\gamma}{\gamma_0}\right] \cdot I_{L-1}\left(2\sqrt{\frac{K(K+1)L\gamma}{\gamma_0}}\right), \qquad \gamma \ge 0 \tag{3}$$

where $I_{L-1}(\cdot)$ is the $(L-1)$th-order modified Bessel function of the first kind for statistically identical diversity branches and K is the ratio of the mean direct power to the mean diffused power.

The exact SER of coherent MPSK under a nonfading channel can be represented as [6]

$$P_{s,exact,MPSK} = \frac{1}{\pi}\int_{-\frac{\pi}{2}}^{\frac{\pi}{2}-\frac{\pi}{M}} \exp\left[-\gamma \sin^2\left(\frac{\pi}{M}\right)\sec^2\theta\right] d\theta. \tag{4}$$

Once the statistics of the instantaneous SNR are determined as the function of the fading parameter and the average SNR, the exact performance in the fading channels is accomplished by averaging the exact SER of coherent MPSK under a nonfading channel, i.e., (See Appendix A.)

$$\begin{aligned} P_{s,exact,MPSK,MRC} &= \int_0^\infty P_{s,exact,MPSK}\, f(\gamma)\, d\gamma \\ &= \frac{1}{\pi}\int_{-\frac{\pi}{2}}^{\frac{\pi}{2}-\frac{\pi}{M}} \left[\frac{K+1}{K+1+\gamma_0 \sin^2\left(\frac{\pi}{M}\right)\sec^2\theta}\right]^L \cdot \\ &\quad \exp\left[-\frac{KL\gamma_0 \sin^2\left(\frac{\pi}{M}\right)\sec^2\theta}{K+1+\gamma_0\sin^2\left(\frac{\pi}{M}\right)\sec^2\theta}\right] d\theta \end{aligned} \tag{5}$$

which can be written in the integral-form, not in the closed-form.

For the special case of $K=0$, we can thus find that the result of (5) corresponds to that of [6, Eq. (12)], when there is no diversity branch.

The approximation of coherent MPSK signals on the probability of symbol error for larger M may be represented as follows[8]:

$$P_{s,MPSK} = erfc\left(\sqrt{\gamma}\sin\frac{\pi}{M}\right) \tag{6}$$

where $erfc(\cdot)$ is the error function defined as

$$erfc(x) = \frac{2}{\sqrt{\pi}}\int_x^\infty e^{-t^2} dt. \tag{7}$$

Consequently, we can express the approximate form to the average SER of MPSK signals with MRC diversity reception in a Rician fading channel as (See Appendix B.)

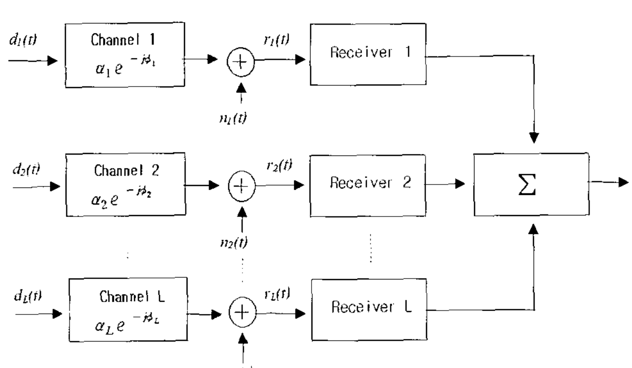

Fig. 1. Model of digital communication system with MRC diversity.

$$P_{s,MPSK,MRC} = \int_0^{\infty} P_{s,MPSK}\, f(\gamma)\, d\gamma$$
$$= \frac{1}{\sqrt{\pi}} e^{-KL} \sum_{l=0}^{\infty} \frac{1}{\Gamma(l+1)} (KL)^l \cdot \frac{\Gamma(L+l+1/2)}{(1/4)^{L+l}} \frac{\Gamma(L+l+1)}{\Gamma(2L+2l+1)} \cdot \beta(K)\left\{\frac{1}{\beta(K)} - \sum_{t=0}^{L+l-1} \binom{2t}{t} \left[\frac{(1+K)/4}{\mu^2+1+K}\right]^t\right\} \quad (8)$$

where $\beta(K) = \frac{\mu}{\sqrt{\mu^2+1+K}}$ and $\mu = \sqrt{\gamma_0}\sin\left(\frac{\pi}{M}\right)$.

For M=2, we can observe that the result of (8) is equivalent to [9, Eq. (A.5)].

The performance comparison among different alphabet size M is not fair at all, since the bit rate is different. It is then desirable to convert the probability of symbol error into the equivalent probability of bit error. For equiprobable orthogonal signals, the average probability of bit error, P_b, is related to the average probability of symbol error, P_s, by [4], [10]

$$P_b = \frac{M}{2(M-1)} P_s \quad (9)$$

where the per bit average SNR γ_b is related to the per symbol average SNR by

$$\gamma_b = \frac{L}{\log_2 M} \gamma_0 \,. \quad (10)$$

When MDPSK signals experience no fading, the expression for the conditional probability of error is given by [6]

$$P_{s,MDPSK} = \frac{\sin\frac{\pi}{M}}{2\pi} \cdot \int_{-\frac{\pi}{2}}^{+\frac{\pi}{2}} \frac{\exp\left[-\gamma\left(1-\cos\frac{\pi}{M}\cos\theta\right)\right]}{1-\cos\frac{\pi}{M}\cos\theta} d\theta. \quad (11)$$

We can represent the average SER in receiving MDPSK signals with MRC diversity branches on Rician fading channels by averaging (11) over underlying fading SNR as follows:

$$P_{s,MDPSK,MRC} = \int_0^{\infty} P_{s,MDPSK}\, f(\gamma)\, d\gamma \quad (12)$$

where $P_{s,MDPSK,MRC}$ is the average SER of MDPSK signals under the Rician fading model.

Next, substituting (3) and (11) into (12), we find the symbol error probability under the Rician fading model to be

$$P_{s,MDPSK,MRC} = \frac{\sin\frac{\pi}{M}}{2\pi} \left(\frac{K+1}{\gamma_0}\right)^L \int_{-\frac{\pi}{2}}^{+\frac{\pi}{2}} \cdot \frac{1}{\left(1+\frac{1+K}{\gamma_0} - \cos\frac{\pi}{M}\cos\theta\right)^L} \frac{1}{1-\cos\frac{\pi}{M}\cos\theta} \cdot \exp\left[-KL + \frac{\frac{K(K+1)L}{\gamma_0}}{1+\frac{1+K}{\gamma_0} - \cos\frac{\pi}{M}\cos\theta}\right] d\theta, \quad L \geq 1, \quad (13)$$

which can be written in the integral-form, not in the closed-form and converted to the BER, using (9) and (10). For the special case of Rician factor K=0, we can observe that the result of (13) for DPSK is equivalent to the result of [10, Eq. (B1)] for Nakagami fading index m=1, when there is no diversity branch. We can also find that the result of (13) for L=1 corresponds to [6, Eq. (5)].

Ⅳ. Numerical Results

For the particular case when L is 1, the results for coherent MPSK using the approximation and the exact BER in the fading channels are plotted with alphabet sizes M=4, 8 in Figs. 2 and 3, respectively. In each of these figures, we have the single value of M, K=0, 6 dB, ∞. We can find the difference between the approximation and the actual BER through numerical analyses. Given average SNR per bit, the performance of the exact BER almost becomes close to the approximation without the relation between the fading parameter. On the other hand, for larger SNR per bit, the performance for the exact BER of coherent M-ary FSK (MFSK) comes closer to the upper bound with increasing K. Also, the results of coherent MFSK indicate that, by increasing the number of M, the discrepancy between the upper bound and the exact BER becomes more apparent as K decreases[11].

Next, the selected numerical results to show the perfor-

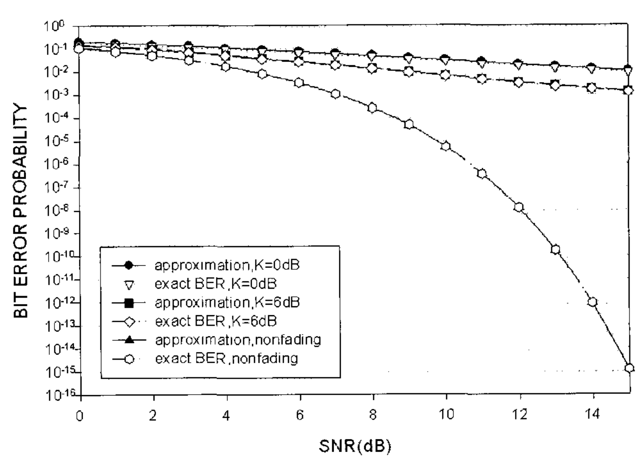

Fig. 2. Coherent MPSK performance comparison of the approximation and the exact BER for M=4.

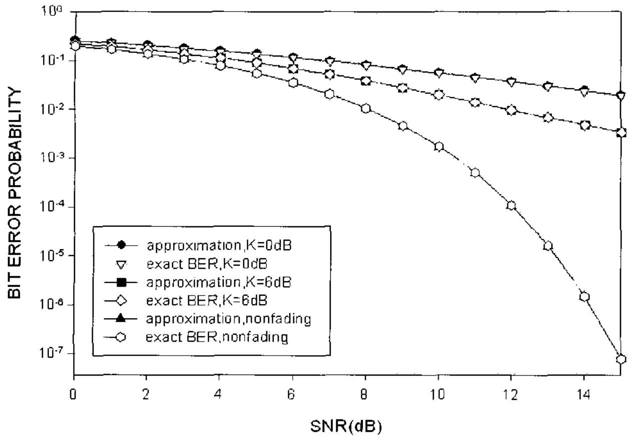

Fig. 3. Coherent MPSK performance comparison of the approximation and the exact BER for M=8.

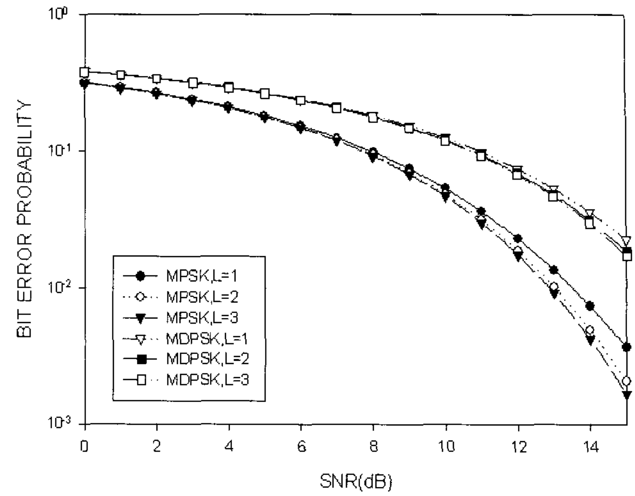

Fig. 5. Error performance comparisons of MPSK and MDPSK signals with MRC diversity receiver structures in Rician fading channels. These parameters for this figure are M=16, L=1, 2, 3, and K=12 dB.

mance of M-ary modulation systems in Rician fading channels with arbitrary fading parameters are presented in the presence of MRC diversity reception.

Let us assume that the performance of coherent MPSK in the fading channels has, first of all, an approximation. Fig. 4 shows MPSK and MDPSK performance comparison of the Rician channel for K=6 dB and M=8. Given the bit error probability of M-ary modulation systems, the SNR per bit is more deviated as the number of diversity branches decreases. It is noted that the performance of L=1 in MPSK signals is rather better than that of L=3 in MDPSK signals in a practical SNR range. Next in Fig. 5, the average SER performances are plotted against the order of diversity with K=12 dB and M=16. The performance is improved very restrictedly

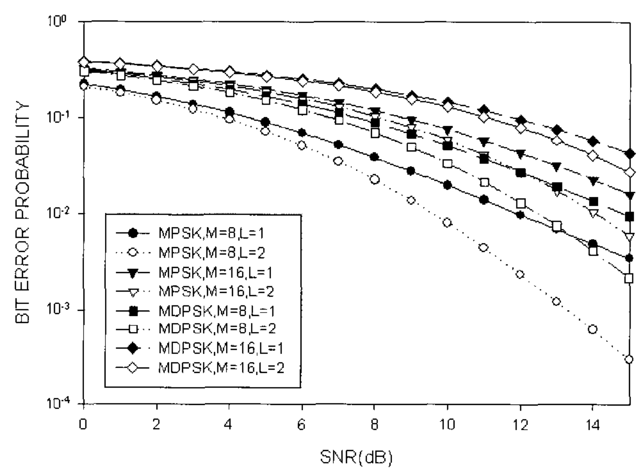

Fig. 6. Error performances of MPSK and MDPSK signals adopting MRC diversity technique. The parameters for this figure are M=8, 16, L=1, 2, and K=6 dB.

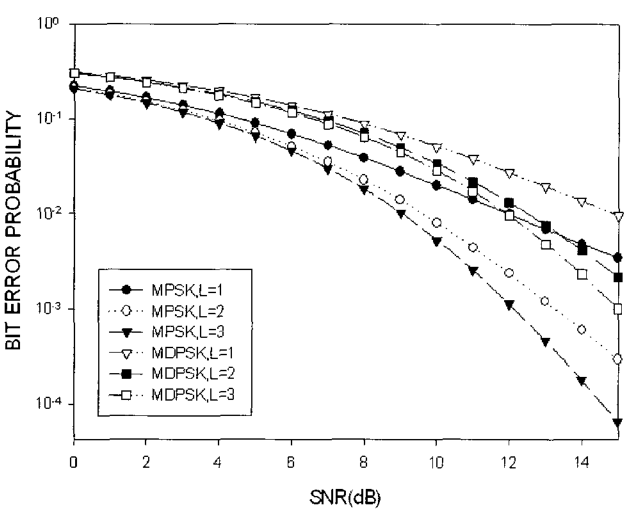

Fig. 4. Error performance comparisons of MPSK and MDPSK signals with MRC diversity receiver structures in Rician fading channels. These parameters for this figure are M=8, L=1, 2, 3, and K=6 dB.

in Rician fading conditions with increasing the diversity branches. Fig. 6 illustrates the performance of MRC diversity system in MPSK and MDPSK signals for the SNR per bit with K=6 dB, M=8, 16, and L=1, 2. Given the value of M signal waveforms, the discrepancy between the error performance of MPSK systems becomes more apparent than that of MDPSK signals as the order of diversity grows.

V. Conclusion

The performance for MPSK and MDPSK signals in

Rician fading channel environments has been evaluated. The approximation to a bit error probability and the exact SER for coherent MPSK in the fading channels have been presented, respectively.

The integral-form performances for MDPSK systems employing the multichannel MRC diversity in the presence of Rician-distributed slow and nonselective fadings have been analyzed.

It is expected result as the diversity branches L increase, the fading depth decreases. This result also shows that the restricted performance gain is achievable with increasing the number of the diversity branches, L. We can predict that the performance improvement saturates as L increases more than 4.

The results of the present works are sufficiently general in offering a convenient method to evaluate the performance of several current M-ary modulation systems that operate on channels with a wide variety of fading conditions in wireless personal communications.

Appendix A: The Integral-form Derivation of (5)

In this Appendix, given that μ, ν, α, and β are real numbers, the exact SER of coherent MPSK signals at the output of L-branch MRC in a Rician fading channel can be derived through some mathematical manipulations.

Invoking (3) and (4) and using the identity [12, p. 716, Eq. (6.631.1)]

$$\int_0^{\infty} x^{\mu} e^{-\alpha x^2} J_{\nu}(\beta x) dx = \frac{\beta^{\nu} \Gamma\left(\frac{\nu}{2}+\frac{\mu}{2}+\frac{1}{2}\right)}{2^{\nu+1} \alpha^{\frac{1}{2}(\mu+\nu+1)} \Gamma(\nu+1)} \, {}_1F_1 \cdot \left(\frac{\nu+\mu+1}{2};\ \nu+1;\ -\frac{\beta^2}{4\alpha}\right),\ Re\ \alpha>0,\ Re\ (\mu+\nu)>-1, \tag{A.1}$$

where $\Gamma(\cdot)$ is the gamma function [12, p. 933, Eq. (8.310.1)], ${}_1F_1(\cdot)$ is the Gauss hypergeometric function, which is expressed as [13]

$${}_1F_1(\alpha;\ \beta;\ x) = \sum_{n=0}^{\infty} \frac{\Gamma(\beta)\Gamma(\alpha+n)x^n}{\Gamma(\alpha)\Gamma(\beta+n)n!}, \quad \beta \neq 0, -1, -2, \cdots, \tag{A.2}$$

and $J_{\nu}(z)$ is the νth-order Bessel function of the first kind, defined as [12, p. 952, Eq. (8.406.3)]

$$I_{\nu}(z) = i^{-\nu} J_{\nu}(iz), \qquad i=\sqrt{-1} \tag{A.3}$$

where $I_{\nu}(z)$ is the νth-order modified Bessel function of the first kind, which may be presented by the infinite series [12, p. 961, Eq. (8.445)]

$$I_{\nu}(z) = \sum_{k=0}^{\infty} \frac{(z/2)^{\nu+2k}}{k!\,\Gamma(\nu+k+1)}, \qquad x \geq 0, \tag{A.4}$$

we can obtain (5) as

$$P_{s,\,exact,\,MPSK,\,MRC} = \frac{1}{\pi} \int_{-\frac{\pi}{2}}^{\frac{\pi}{2}-\frac{\pi}{M}} \left[\frac{K+1}{K+1+\gamma_0 \sin^2\left(\frac{\pi}{M}\right)\sec^2\theta}\right]^L \cdot \exp\left[-\frac{KL\gamma_0 \sin^2\left(\frac{\pi}{M}\right)\sec^2\theta}{K+1+\gamma_0\sin^2\left(\frac{\pi}{M}\right)\sec^2\theta}\right] d\theta. \tag{A.5}$$

For the special case of $K=0$, we can thus find that the result of (5) corresponds to that of [6, Eq. (12)], when there is no diversity branch.

Appendix B: The Approximated-form Derivation of (8)

In this Appendix, we can derive the approximate performance of MPSK signals under the effect of MRC diversity in a Rician fading channels.

Using the identity [12, p. 649, Eq. (6.286.1)]

$$\int_0^{\infty} erfc(\beta x) e^{\mu^2 x^2} x^{\nu-1} dx = \frac{\Gamma\left(\frac{\nu+1}{2}\right)}{\sqrt{\pi}\nu\beta^{\nu}} \, {}_2F_1\left(\frac{\nu}{2}, \frac{\nu+1}{2};\ \frac{\nu}{2}+1;\ \frac{\mu^2}{\beta^2}\right), \quad Re\ \beta^2 > Re\ \mu^2, Re\ \nu > 0, \tag{B.1}$$

where

$${}_2F_1(a, b;\ c;\ z) = \frac{\Gamma(c)}{\Gamma(a)\Gamma(b)} \sum_{n=0}^{\infty} \frac{\Gamma(a+n)\Gamma(b+n)}{\Gamma(c+n)} \frac{x^n}{n!}, \tag{B.2}$$

we can find the average SER with MRC in a Rician fading channel to be

$$P_{s,MPSK,MRC} = \int_0^{\infty} P_{s,MPSK}\, f(\gamma)\, d\gamma = \frac{1}{\sqrt{\pi}}\left(\frac{K+1}{\gamma_0}\right)^L e^{-KL} \sum_{l=0}^{\infty} \frac{1}{\Gamma(l+1)}\left[\frac{KL(K+1)}{\gamma_0}\right]^l \cdot \frac{1}{\left(\sin\frac{\pi}{M}\right)^{2(L+l)}} \cdot \frac{\Gamma(L+l+1/2)}{\Gamma(L+l+1)} \cdot {}_2F_1\left(L+l,\ L+l+\frac{1}{2};\ L+l+1;\ -\frac{1}{\sin^2\left(\frac{\pi}{M}\right)}\frac{K+1}{\gamma_0}\right) \tag{B.3}$$

Next, using identities [12, p. 1043, Eq. (9.131.1)], [14, Eq. (11)]

$${}_2F_1(\alpha, \beta;\ \gamma;\ Z) = (1-Z)^{-\beta}\, {}_2F_1\left(\beta, \gamma-\alpha;\ \gamma;\ \frac{Z}{Z-1}\right) \tag{B.4}$$

and

$${}_2F_1\left(1, p+\frac{1}{2};\ p+1;\ x\right) = \frac{1}{\binom{2p}{p}\left(\frac{x}{4}\right)^p}\left[\frac{1}{\sqrt{1-x}} - \sum_{s=0}^{p-1}\binom{2s}{s}\left(\frac{x}{4}\right)^s\right], \tag{B.5}$$

we can reduce the number of infinite-series including ${}_2F_1(\cdot)$ on the right-hand side of (B.3) and thus express (B.3) as follows:

$$P_{s,MPSK,MRC} = \frac{1}{\sqrt{\pi}} e^{-KL} \sum_{l=0}^{\infty} \frac{1}{\Gamma(l+1)} (KL)^l \cdot \frac{\Gamma(L+l+1/2)}{(1/4)^{L+l}} \frac{\Gamma(L+l+1)}{\Gamma(2L+2l+1)} \cdot \beta(K) \left\{ \frac{1}{\beta(K)} - \sum_{t=0}^{L+l-1} \binom{2t}{t} \left[\frac{(1+K)/4}{\mu^2+1+K} \right]^t \right\} \quad \text{(B.6)}$$

where $\beta(K) = \frac{\mu}{\sqrt{\mu^2+1+K}}$ and $\mu = \sqrt{\gamma_0} \sin\left(\frac{\pi}{M}\right)$.

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