

Average Symbol Error Rate of Maximal Ratio Combining Scheme in the Presence of Multiple Cochannel Interferers

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ABSTRACT

The performance of digital cellular radio systems with maximal ratio combining (MRC) diversity is analyzed in a flat-fading channel with cochannel interference and additive white Gaussian noise (AWGN). It is assumed that the desired signal may experience Rice fading while the Rayleigh-faded interferers may have similar or dissimilar average mean powers. Closed-form expressions are derived for the average error probability in the presence of multiple independent Rayleigh-faded interferers. Both binary and M-ary linear modulation schemes are considered. The analysis is also extended to the case when the desired signal has a Rice distribution.

Keywords: Cochannel Interference, Maximal Ratio Combining, Rayleigh Fading, Rice Fading

1. INTRODUCTION

Multipath fading and cochannel interference are the major sources of performance impairment in wireless communication systems. It is well known that antenna diversity can significantly improve the capacity of many digital cellular radio systems, by combating multipath fading and reducing the effect of cochannel interference. In a system with optimum diversity combining, the signals received by several antenna elements can be properly weighted and combined to maximize the output signal-to-interference-plus-noise ratio (SINR) [1]. However, the implementation and performance analysis of the optimum combiner (OC) require channel estimation for the desired signal as well as for all the interfering signals, which are usually quite complicated. While maximal ratio combining (MRC), which in practice maximizes the output signal-to-noise ratio (SNR) and is suboptimal in the presence of cochannel interference, is usually employed instead [2], [3].

In many mobile radio systems, the assumption of Rayleigh fading channel is usually assumed to model the channel characteristics of both desired and interfering signals since the distance between mobile and the base station is usually random and long enough for the line of

sight (LOS) component of the desired signal to be neglected. Both desired and interfering signals, therefore, experience deep fading from mobile unit to the base station [1], [4], [5]. On the other hand, in microcellular mobile radio systems especially in an indoor environment, the distance of desired signal from the base station is usually not far enough to neglect the line of sight (LOS) component [7], [8]. In such systems, the assumption of Rayleigh fading for both the desired and interfering signals may not be appropriate and the Rice/Rayleigh fading model may be used instead [4], [5].

The performance of MRC is considerably easier to evaluate than that of the optimum combiner. Although the optimum combiner outperforms MRC, the performance difference is sometimes not substantial in some fading environments [1], [6]. The performance of an antenna array system using MRC in a fading environment with cochannel interference and an additive white Gaussian noise (AWGN) is usually given in terms of outage probability, which is the probability that the output SINR falls below a prescribed threshold value [2], [9]. Average error rate is another important measure that is commonly used to characterize the performance of many digital cellular radio systems. Closed-form results for average bit error rate (BER) of antenna array systems using MRC in a fading environment when the desired signal and the interferers fade independently with Rayleigh statistics were given for the special cases of dual-order diversity with one dominant cochannel interferer, assuming that the diversity channels are correlated [3]. Also the average BER for N-order diversity with an arbitrary number of identical independent interferers was derived in [3] and [10].

In this paper we derive semi-analytical expressions for the average symbol error rate (SER) for an N -element antenna array system employing MRC in the presence of independent arbitrary number of Rayleigh-faded cochannel interferers with either identical or distinct mean powers for some linear M-ary modulation schemes. The special cases of average BER of coherent and noncoherent binary phase shift keying is also provided. Finally, the extension to Rice-faded desired signal is also investigated for binary case as well.

The organization, in this paper, is as follow the system model of Rayleigh fading channel is provided in

section II. In section III, the derivation of semi-analytical expressions for the average symbol error rate for an N-element antenna array system employing MRC in the presence of independent arbitrary number of Rayleigh-faded cochannel interferers with either identical or distinct mean powers in some linear modulation schemes are derived. Some special cases of average bit error rate are investigated in this section as well. The extension to Rice/ Rayleigh fading model is studied in section IV. In addition, numerical considerations and conclusion of this manuscript are finally given in section V. and VI. respectively.

2. SYSTEM MODEL

In a cellular radio system with antenna array processing at the base station, the signals received on an N-element antenna array operating in the presence of L interferers may be written as

$$r(t) = \sum_{i=0}^L s_i(t) \mathbf{A}_i + \mathbf{n}, \quad (1)$$

where $s_i(t)$ ($i = 0, 1, 2, \dots, L$) are the transmitted signals and \mathbf{A}_i ($i = 0, 1, 2, \dots, L$) are the N-dimensional direction-of-arrivals of the desired signal ($i = 0$) and the interfering signals ($i = 1, 2, \dots, L$). We assume that each component of the vectors \mathbf{A}_i ($i = 0, 1, 2, \dots, L$) follows a complex Gaussian distribution with mean power P_i ($i = 0, 1, 2, \dots, L$). The vector \mathbf{n} is an additive Gaussian noise vector with zero mean and covariance matrix $\sigma^2 \mathbf{I}$. When the signal at each antenna is weighted by a complex-valued weight vector \mathbf{w} , the output SINR is given by [1]

$$\gamma = \frac{\mathbf{w}^H \mathbf{A}_0 \mathbf{A}_0^H \mathbf{w}}{\mathbf{w}^H \mathbf{R} \mathbf{w}}, \quad (2)$$

where the random covariance matrix \mathbf{R} is given by

$$\mathbf{R} = \sum_{i=1}^L \mathbf{A}_i \mathbf{A}_i^H + \sigma^2 \mathbf{I}. \quad (3)$$

The superscript H denotes the Hermitian transpose of the complex matrix. When the receiver employs MRC, the weight vector on each antenna is proportional to the desired signal vector (i.e., $\mathbf{w} = \mathbf{A}_0$) and the corresponding output SINR can be shown to be given by [3]

$$\gamma = \frac{x_0}{\sum_{i=1}^L z_i + 1}, \quad (4)$$

where $x_0 = \frac{1}{\sigma^2} \mathbf{A}_0^H \mathbf{A}_0$ is a Gamma distributed random variable with pdf given by [2], [9]

$$p_{x_0}(x_0) = \frac{x_0^{N-1}}{\Gamma(N) \Omega_0^N} \exp\left(-\frac{x_0}{\Omega_0}\right), \quad x_0 \geq 0, \quad (5)$$

and the random variable z_i has the exponential distribution given by

$$p_{z_i}(z_i) = \frac{1}{\Omega_i} \exp\left(-\frac{z_i}{\Omega_i}\right), \quad z_i \geq 0, \quad (6)$$

where $\Omega_i = P_i / \sigma^2$ ($i = 0, 1, 2, \dots, L$) is the signal-to-noise power ratio of the desired signal ($i = 0$) and i -th interfering signal. We assume throughout that the interfering powers are mutually independent. When the interferers have identical power (i.e., $P_i = P$ for $i = 1, 2, \dots, L$), the pdf of random variable $z = \sum_{i=1}^L z_i$ can be shown to be given by

$$p_z(z) = \frac{z^{L-1}}{\Gamma(L) \Omega^L} \exp\left(-\frac{z}{\Omega}\right), \quad z \geq 0. \quad (7)$$

When the interferers have dissimilar mean powers, the pdf of z is given by

$$p_z(z) = \sum_{i=1}^L \frac{\pi_i}{\Omega_i} \exp\left(-\frac{z}{\Omega_i}\right), \quad z \geq 0, \quad (8)$$

where

$$\pi_i = \prod_{\substack{k=1 \\ k \neq i}}^L \frac{P_i}{P_i - P_k} = \prod_{\substack{k=1 \\ k \neq i}}^L \frac{\Omega_i}{\Omega_i - \Omega_k}.$$

The pdf of the output SINR, $\gamma = \frac{x_0}{z+1}$ is then given by

$$p_\gamma(\gamma) = \int_0^\infty (\omega+1) p_{x_0}[(\omega+1)\gamma] p_z(\omega) d\omega, \quad \gamma \geq 0, \quad (9)$$

where we may substitute (5) for $p_{x_0}(x_0)$ and (7) or (8) for $p_z(\omega)$, and evaluate the resulting integral. When the interferers have equal mean power, the result is given by

$$p_\gamma(\gamma) = \frac{\gamma^{N-1} \exp(-\gamma/\Omega_0)}{\Gamma(N)\Gamma(L)\Omega_0^N} \sum_{k=0}^N C_k^N \frac{\Omega^k \Gamma(L+k)}{((\Omega/\Omega_0)\gamma+1)^{L+k}}, \quad \gamma \geq 0, \quad (10)$$

where $C_k^n = \frac{n!}{k!(n-k)!}$ is the binomial coefficient,

which agrees with $m_s = m_i = 0$ in [10, eq. (10)]. In the presence of interferers with nonidentical mean powers, we have

$$p_\gamma(\gamma) = \frac{\gamma^{N-1} e^{-\gamma/\Omega_0}}{\Gamma(N)\Omega_0^N} \sum_{i=1}^L \frac{\pi_i \Omega_0 / \Omega_i}{(\gamma + \Omega_0 / \Omega_i)} \sum_{k=0}^N C_k^N \left(\frac{\Omega_0}{\gamma + \Omega_0 / \Omega_i} \right)^k \times \Gamma(k+1), \quad \gamma \geq 0. \quad (11)$$

3. AVERAGE SYMBOL ERROR RATE

In this section we derive the average symbol error rate (SER), which is given by

$$\bar{P}_M = \int_0^\infty P_M(\gamma) p_\gamma(\gamma) d\gamma, \quad (12)$$

where $P_M(\gamma)$ denotes the conditional SER for the desired M-ary signaling schemes in an AWGN channel, depending on the type of detection scheme employed at the receiver. In particular, we consider coherent and differentially coherent M-ary PSK (MPSK and MDPSK), M-ary quadrature amplitude modulation (MQAM) and noncoherent M-ary frequency shift keying (MFSK). For a given SINR γ , let $P_M(\gamma)$ denote the conditional symbol error rate (SER) in an AWGN channel for the desired M-ary signaling scheme. Then, the average SER (averaging over the pdf of the SINR given in (9)) may be written as

$$\bar{P}_M = \int_0^\infty \int_0^\infty (y+1)^N \frac{\gamma^{N-1}}{\Gamma(N)\Omega_0^N} \exp\left\{-\gamma \frac{(y+1)}{\Omega_0}\right\} \times p_z(y) P_M(\gamma) dy d\gamma. \quad (13)$$

Interchanging the order of the integrations, (13) may be re-written as

$$\bar{P}_M = \int_0^\infty (y+1)^N p_z(y) \int_0^\infty \left[\frac{\gamma^{N-1}}{\Gamma(N)\Omega_0^N} \exp\left\{-\gamma \frac{(y+1)}{\Omega_0}\right\} P_M(\gamma) d\gamma \right] dy. \quad (14)$$

We note that the inner integral in (14) is the average SER, where the averaging is done over the pdf of the desired signal SNR (i.e., no interferers). Thus for a given M-ary signaling scheme with N-order MRC diversity operating in a Rayleigh fading channel with no interferers ($L=0$) and average desired signal SNR ϑ , we define the integral

$$\mathcal{J}(M, N, \vartheta) = \int_0^\infty \frac{\gamma^{N-1} e^{-\gamma/\vartheta}}{\Gamma(N)\Omega_0^N} P_M(\gamma) d\gamma. \quad (15)$$

Many researchers have studied the integral in (15) for several linear modulation schemes [12]-[15]. For identical mean-power Rayleigh-faded interferers, we substitute (7) in (14) and using (15), we have

$$\begin{aligned} \bar{P}_M &= \frac{1}{\Gamma(L)\Omega^L} \int_0^\infty (y+1)^N y^{L-1} e^{-y/\Omega} \\ &\quad \times \mathcal{J}(M, N, \Omega_0/(y+1)) dy \\ &= \frac{1}{\Gamma(L)\Omega^L} \sum_{k=0}^N C_k^N \int_0^\infty y^{L+k-1} e^{-y/\Omega} \\ &\quad \times \mathcal{J}(M, N, \Omega_0/(y+1)) dy. \end{aligned} \quad (16)$$

Similarly when the interferers have different mean powers, we have

$$\bar{P}_M = \sum_{i=1}^L \frac{\pi_i}{\Omega_i} \sum_{k=0}^N C_k^N \int_0^\infty y^k e^{-y/\Omega_i} \mathcal{J}(M, N, \Omega_0/(y+1)) dy. \quad (17)$$

It is worth noting that, for the special cases of noncoherent binary phase shift keying, the average bit error rate ($\bar{P}_{e,N}$) can be derived by substituting (15) in (16) with $M=2$ and

$$P_2(\gamma) = \frac{1}{2} \exp(-\alpha\gamma), \quad (18)$$

where

$$\alpha = \begin{cases} 1 & \text{for binary PSK} \\ \frac{1}{2} & \text{for binary FSK.} \end{cases}$$

Then, the result becomes

$$\begin{aligned} \bar{P}_{2,N} &= \frac{1}{2\Gamma(L)\Omega^N} \sum_{k=0}^N C_k^N \Omega^k \Gamma(L+k) \\ &\times \psi \left(N, N-L-k+1, \frac{1+\alpha\Omega_0}{\Omega} \right), \end{aligned} \quad (19)$$

which agrees with [10] as a special case of Nakagami fading channel with $m_s = m_i = 1$, where

$$\psi(\alpha, \beta, x) = \frac{1}{\Gamma(\alpha)} \int_0^\infty t^{\alpha-1} (1+t)^{\beta-\alpha-1} e^{-xt} dt, \quad \alpha > 0. \quad (20)$$

is the confluent hypergeometric function of the second kind [17]. In the presence of L independent Rayleigh-faded interferers with dissimilar mean power, we have

$$\bar{P}_{2,N} = \frac{1}{2} \sum_{i=1}^L \frac{\pi_i}{\Omega_i^N} \sum_{k=0}^N C_k^N \Gamma(k+1) \Omega_i^k \psi \left(N, N-k, \frac{1+\alpha\Omega_0}{\Omega_i} \right). \quad (21)$$

In addition, for the case of coherent binary phase shift keying modulation scheme with identical mean-power interferers, with a substitution of

$$P_2(\gamma) = \frac{1}{2} \operatorname{erfc}(\sqrt{\alpha\gamma}), \quad (22)$$

in (16) and some mathematical manipulation, the average bit error rate ($\bar{P}_{2,c}$) is then given by

$$\begin{aligned} \bar{P}_{2,c} &= \frac{1}{2} \left\{ 1 - e^{-N\kappa_0} \sqrt{\frac{\Omega_0}{\Omega}} \sum_{k=0}^{N-1} \frac{C_k^{2k}}{(4\Omega)^k} \sum_{j=0}^k C_j^k \frac{\Gamma(L+j)}{\Gamma(L)} \Omega^j \right. \\ &\times \psi \left(k + \frac{1}{2}, k-L-j + \frac{3}{2}, \frac{1+\Omega_0}{\Omega} \right) \left. \right\}, \end{aligned} \quad (23)$$

which also agrees with [10, eq.(24)] and we have

$$\begin{aligned} \bar{P}_{2,c} &= \frac{1}{2} \left\{ 1 - \sum_{i=1}^L \pi_i \sqrt{\frac{\Omega_0}{\Omega_i}} \sum_{k=0}^{N-1} C_k^{2k} \left(\frac{1}{4\Omega_i} \right)^k \sum_{j=0}^k C_j^k \Gamma(j+1) \right. \\ &\times \Omega_i^j \psi \left(k + \frac{1}{2}, k-j + \frac{1}{2}, \frac{1+\Omega_0}{\Omega_i} \right) \left. \right\}, \end{aligned} \quad (24)$$

in the case of Rayleigh-faded interferers with dissimilar mean powers [11]. Next we consider a number of M-ary

linear modulations for which the integral in (15) has been evaluated in the literature.

3.1 MDPSK

For MDPSK, we have [12], [13]

$$\begin{aligned} \mathcal{J}_{MDPSK}(M, N, \vartheta) &= \frac{\sin \pi/M}{\pi} \\ &\times \int_0^{\pi/2} \frac{d\theta}{(1 - \cos \pi/M \cos \theta) [y + 1 + \vartheta (1 - \cos \pi/M \cos \theta)]^N}. \end{aligned} \quad (25)$$

In the presence of equal mean-power interferers, we substitute (25) in (16) and interchange the order of integration. Upon simplifying the result, we have

$$\begin{aligned} \bar{P}_{MDPSK} &= \frac{\sin \pi/M}{\pi \Gamma(L)\Omega^N} \sum_{k=0}^N C_k^N \Gamma(L+k) \Omega^k \\ &\times \mathcal{P}_{MDPSK}(M, \nu_k, \Omega, \Omega_0). \end{aligned} \quad (26)$$

where

$$\begin{aligned} \mathcal{P}_{MDPSK}(M, \nu_k, \Omega, \Omega_0) &= \\ &\int_0^{\pi/2} \frac{\psi \left(N, \nu_k, \frac{1 + \Omega_0 (1 - \cos \pi/M \cos \theta)}{\Omega} \right)}{(1 - \cos \pi/M \cos \theta)} d\theta, \end{aligned} \quad (27)$$

and $\nu_k = N - L - k + 1$. Note that the final result in (27) is obtained by a simple change of variable $x = 2\theta / \pi$. Similarly, in the presence of nonidentical mean-power interferers, the result becomes

$$\begin{aligned} \bar{P}_{MDPSK} &= \frac{\sin \pi/M}{\pi} \sum_{i=1}^L \frac{\pi_i}{\Omega_i^N} \sum_{k=0}^N C_k^N \Omega_i^k \Gamma(k+1) \\ &\times \mathcal{P}_{MDPSK}(M, N-k, \Omega_i, \Omega_0). \end{aligned} \quad (28)$$

The expressions in (26) and (28) can be easily evaluated numerically with any desired degree of accuracy. For the case of binary DPSK, ($M = 2$), (26) and (28) reduce to (19) and (21), with $\alpha = 1$.

3.2 MPSK

For MPSK, we have [13], [15]

$$\mathcal{J}_{MPSK}(M, N, \vartheta) = \frac{2}{\pi} \int_0^{\pi(M-1)/M} \left[y + 1 + \vartheta \left(\sin^2 \frac{\pi}{M} \sec^2 \theta \right) \right]^{-N} d\theta. \quad (29)$$

Consequently, when equal mean-power interferers present at the receiver, we substitute (29) in (16) and interchange the order of integration, to obtain

$$\begin{aligned} \bar{P}_{MPSK} &= \frac{2}{\pi \Gamma(L) \Omega^N} \sum_{k=0}^N C_k^N \Omega^k \Gamma(L+k) \\ &\times \mathcal{P}_{MPSK}(M, \nu_k, \Omega, \Omega_0), \end{aligned} \quad (30)$$

where

$$\begin{aligned} \mathcal{P}_{MPSK}(M, \nu_k, \Omega, \Omega_0) &= \int_0^{\pi(M-1)/M} \\ &\times \psi \left(N, \nu_k, \frac{1 + \Omega_0 \sin^2 \pi/M \sec^2 \theta}{\Omega} \right) d\theta. \end{aligned} \quad (31)$$

When the interferers have distinct mean powers, the result becomes

$$\begin{aligned} \bar{P}_{MPSK} &= \frac{2}{\pi} \sum_{i=1}^L \frac{\pi_i}{\Omega_i^N} \sum_{k=0}^N C_k^N \Omega_i^k \Gamma(k+1) \\ &\times \mathcal{P}_{MPSK}(M, N-k, \Omega_i, \Omega_0). \end{aligned} \quad (32)$$

3.3 MQAM

For MQAM, we have [14], [15]

$$\begin{aligned} \mathcal{J}_{MQAM}(M, N, \vartheta) &= \frac{q}{\pi} \left[\int_0^{\pi/2} \left(y + 1 + \frac{\mu \vartheta}{\sin^2 \theta} \right)^{-N} d\theta \right. \\ &\left. - \frac{q}{4} \int_0^{\pi/4} \left(y + 1 + \frac{\mu \vartheta}{\sin^2 \theta} \right)^{-N} d\theta \right], \end{aligned} \quad (33)$$

where $q = 4(\sqrt{M} - 1)/\sqrt{M}$ and $\mu = 3 \log_2 M/2(M-1)$.

Then substituting (33) in (16) and interchanging the order of integration, we have

$$\bar{P}_{MQAM} = \frac{1}{\Gamma(L) \Omega^N} \sum_{k=0}^N C_k^N \Omega^k \Gamma(L+k) \mathcal{P}_{MQAM}(M, \nu_k, \Omega, \Omega_0), \quad (34)$$

where

$$\begin{aligned} \mathcal{P}_{MQAM}(M, \nu_k, \Omega, \Omega_0) &= \frac{q}{\pi} \left[\int_0^{\pi/2} \psi \left(N, \nu_k, \frac{1 + \mu \Omega_0 / \sin^2 \theta}{\Omega} \right) d\theta \right. \\ &\left. - \frac{q}{4} \int_0^{\pi/4} \psi \left(N, \nu_k, \frac{1 + \mu \Omega_0 / \sin^2 \theta}{\Omega} \right) d\theta \right], \end{aligned} \quad (35)$$

for the case of identical mean-power interferers. Also, when the interferers have distinct mean powers, the result becomes

$$\begin{aligned} \bar{P}_{MQAM} &= \sum_{i=1}^L \frac{\pi_i}{\Omega_i^N} \sum_{k=0}^N C_k^N \Omega_i^k \Gamma(k+1) \\ &\times \mathcal{P}_{MQAM}(M, N-k, \Omega_i, \Omega_0). \end{aligned} \quad (36)$$

3.4 Noncoherent MFSK

For noncoherent M-ary FSK, we have [13], [16]

$$\begin{aligned} \mathcal{J}_{MFSK}(M, N, \vartheta) &= \frac{1}{M} \sum_{j=2}^M (-1)^j C_j^M \\ &\times \left[y + 1 + \vartheta \left(1 - \frac{1}{j} \right) \right]^{-N}. \end{aligned} \quad (37)$$

When the mean powers of interferers are equal, we substitute (37) in (16) to obtain

$$\begin{aligned} \bar{P}_{MFSK} &= \frac{1}{M \Gamma(L) \Omega^N} \sum_{k=0}^N C_k^N \Omega^k \Gamma(L+k) \sum_{j=2}^M (-1)^j \\ &\times C_j^M \psi \left(N, \nu_k, \frac{1 + \Omega_0 (1-1/j)}{\Omega} \right). \end{aligned} \quad (38)$$

Similarly, when the interferers have dissimilar mean powers, the result becomes

$$\begin{aligned} \bar{P}_{MFSK} &= \frac{1}{M} \sum_{i=1}^L \frac{\pi_i}{\Omega_i^N} \sum_{k=0}^N C_k^N \Omega_i^k \Gamma(k+1) \sum_{j=2}^M (-1)^j \\ &\times C_j^M \psi \left(N, N-k, \frac{1 + \Omega_0 (1-1/j)}{\Omega_i} \right). \end{aligned} \quad (39)$$

For the special case of noncoherent binary FSK ($M=2$), (38) and (39) reduce to (19) and (21), respectively, with $\alpha = 1/2$.

4. EXTENTION TO RICE/RAYLEIGH FADING

In microcellular mobile radio systems, the assumption of Rayleigh fading for both the desired and interfering signals may not also be appropriate since the distance between the base station and mobile is usually not far enough to neglect the line of sight (LOS) components. Propagation measurements in urban microcellular environments as well as indoor radio channels have shown that the received signal envelope has a Rice distribution, indicating the presence of the LOS or specular components [7], [8]. Therefore, it is reasonable to assume that the interfering signals experience deeper fading than the desired signal since the interferers are usually further away from the base station than the desired signal. In this section, we assume that the desired signal has a Rice distribution while the interferers are Rayleigh distributed. We assume further that the Rice factors of the desired signal on the N antennas are identical and defined as $\kappa_0 = m^2/2P_0$, where m is the mean value of each component in \mathbf{A}_0 . The average SNR of the desired signal, x_0 is now a non-central Chi-square distributed random variable with pdf given by

$$p_{x_0}(x_0) = \frac{(x_0/\kappa_0)^{(N-1)/2}}{\Omega_0^{(N+1)/2}} e^{-\frac{x_0+\kappa_0\Omega_0}{\Omega_0}} I_{N-1} \left(2\sqrt{\frac{\kappa_0 x_0}{\Omega_0}} \right). \quad (40)$$

Substituting (7) and (40) in (9), the pdf of output SINR, when the interferers have equal mean power, is given by

$$p_\gamma(\gamma) = \frac{\gamma^{N-1} e^{-(\kappa_0+\gamma/\Omega_0)}}{\Gamma(L)\Omega_0^N} \sum_{j=0}^{\infty} \frac{(\kappa_0\gamma/\Omega_0)^j}{j!\Gamma(N+j)} \sum_{k=0}^{N+j} C_k^{N+j} \times \frac{\Gamma(L+k)\Omega^k}{[(\Omega/\Omega_0)\gamma+1]^{L+k}}, \gamma \geq 0. \quad (41)$$

When the interferers have nonidentical mean powers, by substituting (8) and (40) in (9), the result becomes

$$p_\gamma(\gamma) = \gamma^{N-1} e^{-(\kappa_0+\gamma/\Omega_0)} \sum_{j=0}^{\infty} \frac{(\kappa_0\gamma)^j}{j!} \sum_{k=0}^{N+j} \frac{C_k^{N+j} \Gamma(L+k)\Omega^k}{(\gamma+\Omega_0/\Omega_i)^{L+k}}, \gamma \geq 0. \quad (42)$$

Note that when $\kappa_0 = 0$, (41) and (42) reduce to the pdf for the Rayleigh fading cases given in (10) and (11), respectively. Next we derive the average

BER for the case of Rice-faded desired signal. The results may be easily extended to the case of M-ary signaling.

4.1 Noncoherent Detection

In the presence of a Rice-faded desired signal and L equal mean-power Rayleigh-faded interferers, we may substitute (18) and (41) in (12) to obtain the average BER for the noncoherent demodulation case as

$$\bar{P}_{2,N} = \frac{e^{-\kappa_0}}{2\Gamma(L)\Omega^N} \sum_{j=0}^{\infty} \frac{(\kappa_0/\Omega)^j}{j!} \sum_{k=0}^{N+j} C_k^{N+j} \Gamma(L+k)\Omega^k \times \psi \left(N+j, N+j-L-k+1, \frac{1+\alpha\Omega_0}{\Omega} \right). \quad (43)$$

When the interferers have nonidentical mean powers, the BER is obtained by substituting (18) and (42) in (12) to give

$$\bar{P}_{2,N} = \frac{e^{-\kappa_0}}{2} \sum_{j=0}^{\infty} \frac{\kappa_0^j}{j!} \Gamma(N+j+1) \sum_{i=1}^L \pi_i \sum_{k=0}^{N+j} \frac{(1/\Omega_i)^k}{k!} \times \psi \left(N+j, k, \frac{1+\alpha\Omega_0}{\Omega_i} \right). \quad (44)$$

We observe that (19) and (21) are the special cases of (43) and (44) correspondingly, when $\kappa_0 = 0$.

4.2 Coherent Detection

When the receiver employs coherent demodulation and the interferers have equal mean power, we may substitute (22) and (41) in (12) and use the relation of [18]

$$\int_0^\infty x^{n-1} \exp(-px) \operatorname{erf} c(\sqrt{ax}) dx = \frac{\Gamma(n)}{p^n} \left\{ 1 - \left(\frac{a}{a+p} \right)^{1/2} \sum_{k=0}^{n-1} C_k^{2k} \left(\frac{1}{4} \right)^k \left(\frac{p}{a+p} \right)^k \right\}, \quad p > 0, n > 0. \quad (45)$$

After some manipulation, the average BER is given by

$$\bar{P}_{e,c} = \frac{e^{-\kappa_0}}{2} \sum_{j=0}^{\infty} \frac{\kappa_0^j}{j!} \left[1 - \frac{\sqrt{\alpha\Omega_0}}{\Omega^L \Gamma(L)} \sum_{i=0}^{N+j-1} \frac{C_i^{2i} (1/4)^i}{(1+\alpha\Omega_0)^{i+1/2}} \times \sum_{k=0}^i C_k^i (1+\alpha\Omega_0)^{L+k} \Gamma(L+k) \right]$$

$$\times \psi \left(L + k, L + k - i + \frac{1}{2}, \frac{1 + \alpha \Omega_0}{\Omega} \right). \quad (46)$$

In the presence of distinct mean-power Rayleigh-faded interferers, the average BER is

$$\begin{aligned} \bar{P}_{e_c} = & \frac{e^{-\kappa_0}}{2} \sum_{j=0}^{\infty} \frac{\kappa_0^j}{j!} \left[1 - \sum_{i=1}^L \pi_i \sqrt{\frac{\alpha \Omega_0}{\Omega_i}} \sum_{k=0}^{N+j-1} C_k^{2k} \left(\frac{1}{4\Omega_i} \right)^k \right. \\ & \left. \times \sum_{n=0}^k C_n^k \Gamma(n+1) \Omega_i^n \psi \left(k + \frac{1}{2}, k - n + \frac{1}{2}, \frac{1 + \alpha \Omega_0}{\Omega_i} \right) \right]. \quad (47) \end{aligned}$$

5. NUMERICAL CONSIDERATIONS

In this section we present some numerical results for the average symbol error rate (SER) in some selected cases. In each case, we plot the SER versus the average output SINR, defined as the ratio of average power of the desired signal to the average power of interferers plus noise power. In Fig.1, the SER for 2, 4 and 16-DPSK is plotted against average SINR in the presence of L ($L = 1, 2, 3$) equal mean-power interferers for an MRC system with four diversity branches ($N = 4$). We observe that as the number of interferers in the system increases, the system performance deteriorates. In Fig.2, the SER of noncoherent 4-FSK is plotted against the number of equal-power interferers in the system, with the SNR fixed at 10dB and the interference-to-noise ratio (INR) fixed at 0 dB. We observe from the figure that when the system is interference-limited ($L \gg 0$), the system performance still improves substantially although the MRC system ignores the presence of the interferers and maximizes the output SNR. Also when there are no interferers ($L = 0$) in the system, the performance reaches the best since MRC is the optimum combining scheme in a noise-only system. Finally, for the Rice/Rayleigh fading model, we present the system performance for binary modulation schemes. In Fig. 3, the BER for DPSK and coherent BPSK are plotted against SINR for a system with six ($L = 6$) equal mean-power interferers and a Rice-faded desired signal ($\kappa_0 = 0$ and 10) with different values of the diversity order N . From the figure, we observe that the BER decreases as the Rice factor increases due to the presence of stronger LOS components between the mobile unit and the base station. For example, for a DPSK system at a BER of 10^{-4} , the system with a Rice factor of $\kappa_0 = 10$ has diversity gain of 1.07 dB in using $N = 3$ over $N = 2$ and has diversity gain of 6.19 dB in using $N = 3$ over $N = 2$ with a Rice factor of $\kappa_0 = 0$. Thus the use of diversity reception in the presence of Rayleigh-faded cochannel interference improves system performance more significantly in a Rayleigh fading channel than in a Rice fading channel.

6. CONCLUSION

In this paper we have studied the effect of cochannel interference on the performance of digital cellular mobile radio systems. The interferers were assumed to be subject to independent Rayleigh fading while the desired signal was subject to either Rayleigh or Rice fading. The average SER of maximal ratio combiner, which maximizes the output SNR in the presence of cochannel interference and noise, is analyzed. Semi-analytical expressions that provide a convenient tool for the system performance analysis are derived for both the pdf of the output SINR and the average SER. Some closed-form expressions of the special cases of BER are also studied as well. The analysis assumed an arbitrary number of cochannel interferers with identical or with different average powers.

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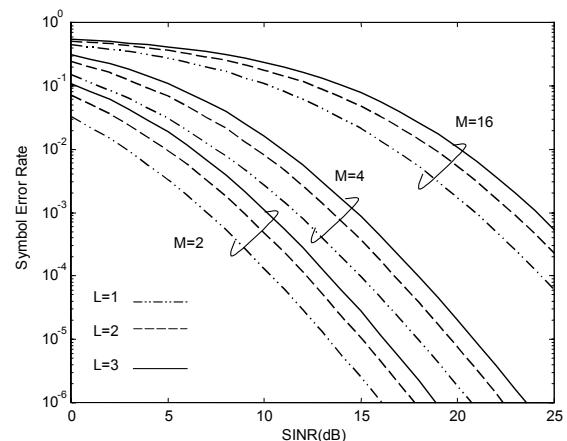


Fig. 1: Average Symbol Error Rate versus SINR of 2, 4 and 16-DPSK with a four-element MRC antenna array in the presence of L ($L = 1, 2, 3$) identical mean-power Rayleigh-faded interferers.

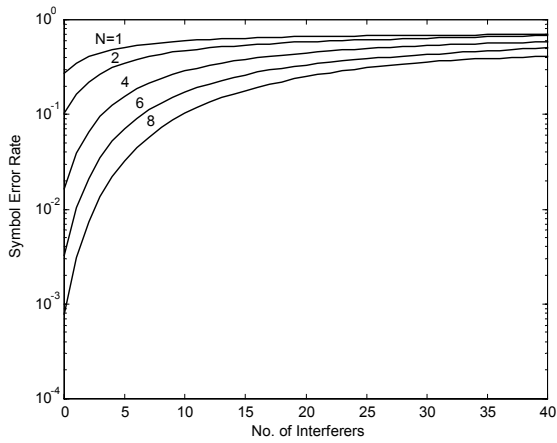


Fig. 2: Average Symbol Error Rate versus number of interferers L for Noncoherent 4-FSK with the diversity order N as a parameter, in the presence of identical mean-power Rayleigh-faded interferers.

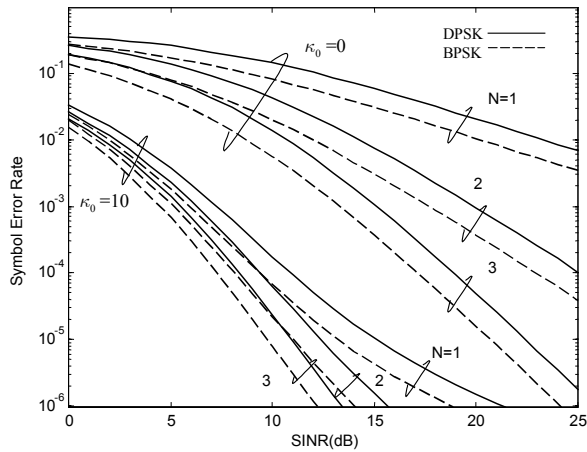


Fig. 3: Average BER versus SINR of DPSK and coherent BPSK with the diversity order N as a parameter, in a Rice/Rayleigh fading channel with six identical mean-power interferers.