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Research Article

Performance Analysis of SSC Diversity Receivers over Correlated Ricean Fading Satellite Channels

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This paper studies the performance of switch and stay combining (SSC) diversity receivers operating over correlated Ricean fading satellite channels. Using an infinite series representation for the bivariate Ricean probability density function (PDF), the PDF of the SSC output signal-to-noise ratio (SNR) is derived. Capitalizing on this PDF, analytical expressions for the corresponding cumulative distribution function (CDF), the moments of the output SNR, the moments generating function (MGF), and the average channel capacity (CC) are derived. Furthermore, by considering several families of modulated signals, analytical expressions for the average symbol error probability (ASEP) for the diversity receivers under consideration are obtained. The theoretical analysis is accompanied by representative performance evaluation results, including average output SNR (ASNR), amount of fading (AoF), outage probability (P_{out}), average bit error probability (ABEP), and average CC, which have been obtained by numerical techniques. The validity of some of these performance evaluation results has been verified by comparing them with previously known results obtained for uncorrelated Ricean fading channels.

1. INTRODUCTION

The mobile terrestrial and satellite communication channel is particularly dynamic due to multipath fading propagation, having a strong negative impact on the average bit error probability (ABEP) of any modulation scheme [1]. Diversity is a powerful communication receiver technique used to compensate for fading channel impairments. The most important and widely used diversity reception methods employed in digital communication receivers are maximal-ratio combining (MRC), equal-gain combining (EGC), selection combining (SC), and switch and stay combining (SSC) [2]. For SSC diversity considered in this paper, the receiver selects a particular branch until its signal-to-noise ratio (SNR) drops below a predetermined threshold. When this happens, the combiner switches to another branch and stays there regardless of whether the SNR of that branch is above or below the predetermined threshold. Hence, among the abovementioned diversity schemes, SSC is the least complex and can be used in conjunction with coherent, noncoherent, and differentially coherent modulation schemes. It is also well known that in many real life communication scenarios the combined signals are correlated [2, 3]. A typical example for such signal correlation exists in relatively small-size mobile terminals where typically the distance between the diversity antennas is short. Due to this correlation between the signals received at the diversity branches there is a degradation in the achievable diversity gain.

The Ricean fading distribution is often used to model propagation paths consisting of one strong direct line-of-sight (LoS) signal and many randomly reflected and usually weaker signals. Such fading environments are typically encountered in microcellular and mobile satellite radio links [2]. In particular for mobile satellite communications the Ricean distribution is used to accurately model the mobile satellite channel for single- [4] and clear-state [5] channel conditions. Furthermore, in [6] it was depicted that the Ricean K-factor characterizes the land mobile satellite channel during unshadowed periods.

The technical literature concerning diversity receivers operating over correlated fading channels is quite rich, for example, see [7–13]. In [7] expressions for the outage probability (P_{out}) and the ABEP of dual SC with correlated Rayleigh fading were derived either in closed form or in terms of single integrals. In [8] the cumulative distribution functions (CDF) of SC, in correlated Rayleigh, Ricean, and Nakagami-m fading channels were derived in terms of single-fold integrals and infinite series expressions. In [9] the ABEP of dual-branch EGC and MRC receivers operating over correlated Weibull fading channels was obtained. In [10] the performance of MRC in nonidentical correlated Weibull fading channels with arbitrary parameters was evaluated. In [11] an analysis for the Shannon channel capacity (CC) of dual-branch SC diversity receivers operating over correlated Weibull fading was presented. In [12], infinite series expressions for the capacity of dual-branch MRC, EGC, SC, and SSC diversity receivers over Nakagami-m fading channels have been derived.

Past work concerning the performance of SSC operating over correlated fading channels can be found in [14–17]. One of the first attempts to investigate the performance of SSC diversity receivers operating over independent and correlated identical distributed Ricean fading channels was made in [14]. However, in this reference only noncoherent frequency shift keying (NCFSK) modulation was considered and its ABEP has been derived in an integral representation form. In [15] the performance of SSC diversity receivers was investigated for different fading channels, including Rayleigh, Nakagami-m and Ricean, and under different channel conditions but dealt mainly with uncorrelated fading. For correlated fading in this reference only the Nakagami-m distribution was studied. In [16] the moments generating function (MGF) of SSC was presented in terms of a finite integral representation for the correlated Nakagami-m fading channel. In [17] expressions for the average output SNR (ASNR), amount of fading (AoF) and P_{out} for the correlated log-normal fading channels have been derived.

All in all, the problem of theoretically analyzing the performance of SSC over correlated Ricean fading channels has not yet been thoroughly addressed in the open technical literature. The main difficulty in analyzing the performance of diversity receivers in correlated Ricean fading channels is the complicated form of the received signal bivariate probability density function (PDF), see [14, Equation (17)], and the absence of an alternative and more convenient expression for the multivariate distribution. An efficient solution to these difficulties is to employ an infinite series representation for the bivariate PDF, such as those that were proposed in [18] or [19]. Such an approach was used in [20] to analyze the performance of MRC, EGC, and SC in the presence of correlated Ricean fading. Similarly here the most important statistical metrics and the capacity of SSC diversity receivers operating over correlated Ricean fading channels will be studied. In particular, we derive the PDF, CDF, MGF, moments and the average CC of such receivers operating over correlated Ricean fading channels. Furthermore, analytical expressions for the average symbol error probability (ASEP) of several modulation schemes will be obtained. Capitalizing on these expressions, a detailed performance analysis for the P_{out}, ASNR, AoF, and ASEP/ABEP will be presented.

The remainder of this paper is organized as follows. After this introduction, in Section 2 the system model is introduced. In Section 3, the SSC received signal statistics are presented, while in Section 4 the capacity is obtained. Section 5 contains the derivation of the most important performance metrics of the SSC output SNR. In Section 6, various numerical evaluation results are presented and discussed, while the conclusions of the paper can be found in Section 7.

2. SYSTEM MODEL

By considering a dual-branch SSC diversity receiver operating over a correlated Ricean fading channel, the baseband received signal at the ℓth ($\ell = 1$ and 2) input branch can be mathematically expressed as

$$\zeta_\ell = sh_\ell + n_\ell. \tag{1}$$

In the above equation, s is the transmitted complex symbol, h_ℓ is the Ricean fading channel complex envelope with magnitude $R_\ell = |h_\ell|$, and n_ℓ is the additive white Gaussian noise (AWGN) having single-sided power spectral density of N_0. The usual assumption for ideal fading phase estimation is made, and hence, only the distributed fading envelope and the AWGN affect the received signal. Moreover, the AWGN is assumed to be uncorrelated between the two diversity branches. The instantaneous SNR per symbol at the ℓth input branch is $\gamma_\ell = R_\ell^2 E_s/(2N_0)$, where $E_s = \mathbb{E}\langle |s|^2\rangle$ is the transmitted average symbol energy, where $\mathbb{E}\langle\cdot\rangle$ denoting expectation and $|\cdot|$ absolute value. The corresponding average SNR per symbol at both input branches is $\overline{\gamma} = \Omega E_s/N_0$, where $\Omega = \mathbb{E}\langle R_\ell^2\rangle$. The PDF of the SNR of the Ricean distribution is given by [2, Equation (2.16)]

$$f_\gamma(\gamma) = \frac{1+K}{\overline{\gamma}} \exp\left[-K - \frac{(1+K)}{\overline{\gamma}}\gamma\right] \times I_0\left[2\sqrt{\frac{K(K+1)}{\overline{\gamma}}}\gamma^{1/2}\right], \tag{2}$$

where K is the Ricean K-factor defined as the power ratio of the specular signal to the scattered signals and $I_0(\cdot)$ is the zeroth-order modified Bessel function of the first kind [21, Equation (8.406)]. The CDF of γ is given by [14, Equation (8)]

$$F_\gamma(\gamma) = Q_1\left[\sqrt{2K}, \sqrt{\frac{2(1+K)}{\overline{\gamma}}\gamma}\right], \tag{3}$$

where $Q_1(\cdot)$ is the first-order Marcum-Q function [2, Equation (4.33)].

The joint PDF of γ_1 and γ_2, presented in [14, Equation (17)], can be expressed in terms of infinite series by following a similar procedure as for deriving [18, Equation (9)]. Hence, substituting $I_0(\cdot)$ with its infinite series representation [21, Equation (8.445)], expanding the term $[\gamma_1 + \gamma_2 + 2\sqrt{\gamma_1\gamma_2}\cos(\theta)]^i$ using the multinomial identity [22, Equation (24.1.2)], using [21, Equation (3.389/1)] and after some mathematical manipulations the joint PDF of γ_1, γ_2 can be expressed as

$$f_{\gamma_1,\gamma_2}(\gamma_1,\gamma_2) = \sum_{\substack{i,h=0\\ v_1+v_2+v_3=i}}^{\infty} \mathcal{A}\exp\left[-\beta_1(\gamma_1+\gamma_2)\right] \times \left(\mathcal{B}\gamma_1^{\beta_2-1}\gamma_2^{\beta_3-1} + \mathcal{C}\overline{\gamma}^{-1}\gamma_1^{\beta_2-1/2}\gamma_2^{\beta_3-1/2}\right) \tag{4}$$

with

$$\mathcal{A} = \frac{2^{v_3+2h-1}(1+K)^{1+\beta_4}\rho^{2h}K^i\exp\left[-2K/(1+\rho)\right]}{\sqrt{\pi}\overline{\gamma}^{1+\beta_4}\left(1-\rho^2\right)^{1+2h}v_1!v_2!v_3!i!(1+\rho)^{2i}},$$

$$\mathcal{B} = \frac{\left[1+(-1)^{v_3}\right]\Gamma\left[h+(1+v_3)/2\right]}{\Gamma(h+1+v_3/2)\Gamma(1+2h)},$$

$$\mathcal{C} = \frac{\left[-1+(-1)^{v_3}\right]2\rho(1+K)\Gamma(1+h+v_3/2)}{(\rho^2-1)\Gamma(2+2h)\Gamma\left[h+(3+v_3)/2\right]},$$

$$\beta_1 = \frac{(1+K)}{(1-\rho^2)\overline{\gamma}}, \qquad \beta_2 = v_1 + \frac{v_3}{2} + h + 1,$$

$$\beta_3 = v_2 + \frac{v_3}{2} + h + 1, \qquad \beta_4 = i + 2h + 1, \tag{5}$$

where $\Gamma(\cdot)$ is the Gamma function [21, Equation (8.310/1)] and ρ is the correlation coefficient between γ_1 and γ_2. It can be proved that the above infinite series expression always converges [18].

3. RECEIVED SIGNAL STATISTICS

In this section, the most important statistical metrics, namely, the PDF, CDF, MGF, and moments of dual branch SSC output SNR diversity receivers operating over correlated Ricean fading channels will be presented.

3.1. *Probability density function (PDF)*

Let γ_{ssc} be the instantaneous SNR per symbol at the output of the SSC and γ_τ the predetermined switching threshold. Following [15], the PDF of γ_{ssc}, $f_{\gamma_{\text{ssc}}}(\gamma)$, is given by

$$f_{\gamma_{\text{ssc}}}(\gamma) = \begin{cases} r_{\text{ssc}}(\gamma), & \gamma \le \gamma_\tau, \\ r_{\text{ssc}}(\gamma) + f_\gamma(\gamma), & \gamma > \gamma_\tau. \end{cases} \tag{6}$$

Moreover, $r_{\text{ssc}}(\gamma)$ is given in [23, Equation (21b)] as

$$\begin{aligned} r_{\text{ssc}}(\gamma) &= \int_0^{\gamma_\tau} f_{\gamma_1\gamma_2}(\gamma,\gamma_2)d\gamma_2 \\ &= \int_0^{\infty} f_{\gamma_1\gamma_2}(\gamma,\gamma_2)d\gamma_2 - \int_{\gamma_\tau}^{\infty} f_{\gamma_1\gamma_2}(\gamma,\gamma_2)d\gamma_2. \end{aligned} \tag{7}$$

Hence, by substituting (4) in (7) and using [21, Equation (3.351/2-3)], these integrals can be solved and $r_{\text{ssc}}(\gamma)$ can be expressed as

$$r_{\text{ssc}}(\gamma) = \sum_{\substack{i,h=0\\ v_1+v_2+v_3=i}}^{\infty} \mathcal{A}\exp\left(-\beta_1\gamma\right)\gamma^{\beta_2-1/2} \times \left[\frac{\mathcal{B}\gamma(\beta_3,\beta_1\gamma_\tau)}{\sqrt{\overline{\gamma}}\beta_1^{\beta_3}} + \frac{\mathcal{C}\gamma(\beta_3+1/2,\beta_1\gamma_\tau)}{\overline{\gamma}\beta_1^{\beta_3+1/2}}\right], \tag{8}$$

where $\gamma(\cdot,\cdot)$ is the lower incomplete Gamma function [21, Equation (8.350/1)].

3.2. *Cumulative distribution function (CDF)*

Similar to [23, Equation (20)], the CDF of γ_{ssc}, $F_{\gamma_{\text{ssc}}}(\gamma)$, is given by

$$F_{\gamma_{\text{ssc}}}(\gamma) = \Pr\left(\gamma_\tau \le \gamma_1 \le \gamma\right) + \Pr\left(\gamma_2 < \gamma_\tau \wedge \gamma_1 < \gamma\right) \tag{9}$$

which after some manipulations can be expressed in terms of CDFs as

$$F_{\gamma_{\text{ssc}}}(\gamma) = \begin{cases} F_{\gamma_1,\gamma_2}(\gamma,\gamma_\tau), & \gamma \le \gamma_\tau, \\ F_\gamma(\gamma) - F_\gamma(\gamma_\tau) + F_{\gamma_1,\gamma_2}(\gamma,\gamma_\tau), & \gamma > \gamma_\tau. \end{cases} \tag{10}$$

Hence, by substituting (4) in $F_{\gamma_1,\gamma_2}(\gamma,\gamma_\tau) = \int_0^{\gamma}\int_0^{\gamma_\tau} f_{\gamma_1,\gamma_2}(\gamma_1,\gamma_2)d\gamma_1 d\gamma_2$ using [21, Equation (3.351/1)], $F_{\gamma_1,\gamma_2}(\gamma,\gamma_\tau)$ can be derived as

$$F_{\gamma_1,\gamma_2}(\gamma,\gamma_\tau) = \sum_{\substack{i,h=0\\ v_1+v_2+v_3=i}}^{\infty} \frac{\mathcal{A}}{\beta_1^{\beta_2+\beta_3}} \times \left[\mathcal{B}\gamma(\beta_2,\beta_1\gamma)\gamma(\beta_3,\beta_1\gamma_\tau) + \frac{\mathcal{C}}{\beta_1\overline{\gamma}}\gamma\left(\beta_2+\frac{1}{2},\beta_1\gamma\right)\gamma\left(\beta_3+\frac{1}{2},\beta_1\gamma_\tau\right)\right]. \tag{11}$$

In order to verify the validity of the above derivations, (10) and (11) have been numerically evaluated for the special case of uncorrelated, that is, $\rho = 0$, Ricean fading channels. The resulting CDF was found to be identical to the same CDF presented in [2, Equation 9.273], which was derived using a different mathematical approach as a closed-form expression.

3.3. *Moments generating function (MGF)*

Based on (6), the MGF of γ_{ssc}, $\mathcal{M}_{\gamma_{\text{ssc}}}(s) = \mathbb{E}\langle\exp(-s\gamma_{\text{ssc}})\rangle$, [24, Equation (5.62)], can be expressed in terms of two integrals as

$$\mathcal{M}_{\gamma_{\text{ssc}}}(s) = \int_0^{\infty}\exp(-s\gamma)r_{\text{ssc}}(\gamma)d\gamma + \int_{\gamma_\tau}^{\infty}\exp(-s\gamma)f_\gamma(\gamma)d\gamma = \mathcal{I}_1 + \mathcal{I}_2. \tag{12}$$

Using [21, Equation (3.381/4)], $\mathcal{I}_1$ can be expressed in terms of infinite series as

$$\mathcal{I}_1 = \sum_{\substack{i,h=0\\ v_1+v_2+v_3=i}}^{\infty} \mathcal{A}\Bigg[\frac{\Gamma(\beta_2)}{(\beta_1+s)^{\beta_2}}\mathcal{B}\beta_1^{-\beta_3}\gamma(\beta_3,\beta_1\gamma_\tau) + \mathcal{C}\beta_1^{-\beta_3-1/2}\frac{\Gamma(\beta_2+1/2)}{(\beta_1+s)^{\beta_2+1/2}}\gamma\Big(\beta_3+\frac{1}{2},\beta_1\gamma_\tau\Big)\Bigg]. \tag{13}$$

Setting $\psi = \sqrt{2\gamma[(1+K)/\overline{\gamma}+s]}$ and using [2, Equation (4.33)], $\mathcal{I}_2$ can be solved as

$$\mathcal{I}_2 = Q_1\left[\sqrt{\frac{2K(1+K)}{1+K+\overline{\gamma}s}},\sqrt{\frac{2(1+K+\overline{\gamma}s)\gamma_\tau}{\overline{\gamma}}}\right] \times \exp\left[\frac{K(1+K)}{1+K+s\overline{\gamma}}\right]\frac{(1+K)\exp(-K)}{1+K+\overline{\gamma}s}. \tag{14}$$

3.4. Moments

Based on (6), the moments for γ_{ssc}, $\mu_{\gamma_{\text{ssc}}}(n) = \mathbb{E}\langle\exp(\gamma_{\text{ssc}}^n)\rangle$, [24, Equation (5.38)], can be expressed in terms of two integrals as

$$\mu_{\gamma_{\text{ssc}}}(n) = \int_0^\infty \gamma^n r_{\text{ssc}}(\gamma)d\gamma + \int_{\gamma_\tau}^\infty \gamma^n f_\gamma(\gamma)d\gamma = \mathcal{I}_3 + \mathcal{I}_4. \tag{15}$$

Using again [21, Equation (3.381/4)], $\mathcal{I}_3$ can be expressed in terms of infinite series as

$$\mathcal{I}_3 = \sum_{\substack{i,h=0\\ v_1+v_2+v_3=i}}^{\infty} \mathcal{A}\Bigg[\mathcal{B}\gamma(\beta_3,\beta_1\gamma_\tau)\frac{\Gamma(n+\beta_2)}{\beta_1^{\beta_2+\beta_3+n}} + \frac{\mathcal{C}\gamma(\beta_3+1/2,\beta_1\gamma_\tau)}{\beta_1^{\beta_2+\beta_3+n+1}}\Gamma\Big(n+\beta_2+\frac{1}{2}\Big)\Bigg]. \tag{16}$$

Setting $\phi = \sqrt{2\gamma(1+K)/\overline{\gamma}}$ in $\mathcal{I}_4$, using [2, Equation (4.104)], after some straight-forward mathematical manipulations, yields

$$\mathcal{I}_4 = \frac{\overline{\gamma}^{n-1}}{2^n(1+K)^{n-1}}Q_{2n+1,0}\left(K,\sqrt{\frac{2(1+K)\gamma_\tau}{\overline{\gamma}}}\right), \tag{17}$$

where $Q_{m,n}(\cdot,\cdot)$ is the Nuttal Q-function defined in [25].

4. CHANNEL CAPACITY (CC)

CC is a well-known performance metric providing an upper bound for maximum errorless transmission rate in a Gaussian environment. The average CC, $\overline{C}$, is defined as [26]

$$\overline{C} \triangleq BW\int_0^\infty \log_2(1+\gamma)f_{\gamma_{\text{ssc}}}(\gamma)d\gamma, \tag{18}$$

where BW is transmission bandwidth of the signal in Hz. Hence, substituting (6) in (18), $\overline{C}$ becomes

$$\overline{C} = \int_0^\infty \log_2(1+\gamma)r_{\text{ssc}}(\gamma)d\gamma + \int_{\gamma_\tau}^\infty \log_2(1+\gamma)f_\gamma(\gamma)d\gamma = \mathcal{I}_5 + \mathcal{I}_6. \tag{19}$$

By representing $\ln(1+\gamma) = G_{2,2}^{1,2}\left[\gamma \,|\, {}^{1,1}_{1,0}\right]$, [27, Equation (01.04.26.0003.01)], and $\exp(-\gamma) = G_{0,1}^{1,0}\left[\gamma \,|\, {}^{0}_{-}\right]$, [27, Equation (01.03.26.0004.01)], where $G(\cdot)$ is Meijer's G-function [21, Equation (9.301)] and using [28], $\mathcal{I}_5$ can be solved as

$$\mathcal{I}_5 = \sum_{\substack{i,h=0\\ v_1+v_2+v_3=i}}^{\infty} \frac{\mathcal{A}}{\ln 2}\Bigg\{\mathcal{B}\frac{\gamma(\beta_3,\beta_1\gamma_\tau)}{\beta_1^{\beta_3+\beta_2}}G_{3,2}^{1,3}\left[\frac{1}{\beta_1}\,\middle|\,\begin{matrix}1,1,1-\beta_2\\1,0\end{matrix}\right] + \mathcal{C}\frac{\gamma(\beta_3+1/2,\beta_1\gamma_\tau)}{\beta_1^{\beta_3+\beta_2+3/2}} \times G_{3,2}^{1,3}\left[\frac{1}{\beta_1}\,\middle|\,\begin{matrix}1,1,1-\beta_2\\1,0\end{matrix}\right]\Bigg\}. \tag{20}$$

Due to the very complicated nature of $\mathcal{I}_6$, it is very difficult, if not impossible, to derive a closed-form solution for this integral. However, $\mathcal{I}_6$ can be evaluated via numerical integration using any of the well-known mathematical software packages, such as MATHEMATICA or MATLAB.

5. PERFORMANCE ANALYSIS

In this section a detailed performance analysis, in terms of P_{out}, ASEP, ASNR and AoF, for SSC diversity receivers operating over correlated Ricean fading channels will be presented.

5.1. Outage probability (P_{out})

P_{out} is the probability that the output SNR falls below a predetermined threshold γ_{th}, $P_{\text{out}}(\gamma_{\text{th}})$, and can be obtained by replacing γ with γ_{th} in (10) as

$$P_{\text{out}}(\gamma_{\text{th}}) = F_{\gamma_{\text{ssc}}}(\gamma_{\text{th}}). \tag{21}$$

5.2. Average symbol error probability (ASEP)

The ASEP, $\overline{P}_{\text{se}}$, can be evaluated directly by averaging the conditional symbol error probability, $P_e(\gamma)$, over the PDF of γ_{ssc} [29]

$$\overline{P}_{\text{se}} = \int_0^\infty P_e(\gamma)f_{\gamma_{\text{ssc}}}(\gamma)d\gamma. \tag{22}$$

For different families of modulation schemes, $P_e(\gamma)$ can be obtained as follows.

(i) For binary phase shift keying (BPSK) and square M-ary quadrature amplitude modulation (QAM) signaling formats and for high-input SNR, $P_e(\gamma) = D\,\text{erfc}(\sqrt{E\gamma})$, where

$\mathrm{erfc}(\cdot)$ is the complementary error function [21, Equation (8.250/1)] and D, E are constants the values of which depend on the specific modulation scheme under consideration. Using this expression, by substituting (6) in (22), yields

$$\overline{P}_{\mathrm{se}} = \int_0^\infty D\,\mathrm{erfc}\left(\sqrt{E\gamma}\right) r_{\mathrm{ssc}}(\gamma)d\gamma + \int_{\gamma_\tau}^\infty D\,\mathrm{erfc}\left(\sqrt{E\gamma}\right) f_\gamma(\gamma)d\gamma = \mathcal{I}_7 + \mathcal{I}_8. \tag{23}$$

Expressing $\mathrm{erfc}(\sqrt{E\gamma}) = \sqrt{\pi}^{-1} G_{1,2}^{2,0}\left[B\gamma \,|\, {}^{1}_{0,1/2}\right]$, [27, Equation (06.27.26.0006.01)], and $\exp(-\gamma) = G_{0,1}^{1,0}\left[\gamma \,|\, {}^{0}_{-}\right]$, [27, Equation (01.03.26.0004.01)], using [28] and after some straightforward mathematical manipulations $\mathcal{I}_7$ can be expressed as

$$\begin{aligned}\mathcal{I}_7 = \sum_{\substack{i,h=0\\ v_1+v_2+v_3=i}}^\infty & \frac{\mathcal{A}D\Gamma(\beta_2+1/2)}{\sqrt{\pi}\beta_1^{\beta_3}E^{\beta_2}} \\ &\times\Bigg\{\frac{\mathcal{B}\Gamma(\beta_2)}{\Gamma(\beta_2+1)}\gamma(\beta_3,\beta_1\gamma_\tau) \\ &\quad\times {}_2F_1\left(\beta_2,\beta_2+\frac{1}{2};\beta_2+1;-\frac{\beta_1}{E}\right) \\ &\quad+\frac{\mathcal{C}\gamma(\beta_3+1/2,\beta_1\gamma_\tau)\Gamma(\beta_2+1)}{(\beta_1E)^{1/2}\Gamma\left(\beta_2+\dfrac{3}{2}\right)} \\ &\quad\times {}_2F_1\left(\beta_2+\frac{1}{2},\beta_2+1;\beta_2+\frac{3}{2};-\frac{\beta_1}{E}\right)\Bigg\}\end{aligned} \tag{24}$$

with ${}_2F_1(\cdot,\cdot;\cdot;\cdot)$ being Gauss Hypergeometric function [21, Equation (9.100)]. Moreover, $\mathcal{I}_8 = \int_0^\infty D\,\mathrm{erfc}(\sqrt{E\gamma})f_\gamma(\gamma)d\gamma - \int_0^{\gamma_\tau} D\,\mathrm{erfc}(\sqrt{E\gamma})f_\gamma(\gamma)d\gamma = \mathcal{I}_{8,a} - \mathcal{I}_{8,b}$. Hence, substituting again $I_0(\cdot)$ with its infinite series representation [21, Equation (8.445)], $\mathcal{I}_{8,a}$ can be solved with the aid of [28] and $\mathcal{I}_{8,b}$ using [27, Equation (06.27.21.0019.01)]. Thus, using these solutions of $\mathcal{I}_{8,a}$ and $\mathcal{I}_{8,b}$ and after some mathematical manipulations, $\mathcal{I}_8$ can be expressed as in (25):

$$\begin{aligned}\mathcal{I}_8 = \frac{D(1+K)\exp(-K)}{\overline{\gamma}} & \sum_{k=0}^\infty (k!)^{-2}\left[\frac{K(K+1)}{\overline{\gamma}}\right]^k \\ &\times\Bigg\{\frac{\Gamma(k+1)\Gamma(k+3/2)}{\sqrt{\pi}E^{k+1}\Gamma(k+2)} \\ &\quad\times {}_2F_1\left[k+1,k+\frac{3}{2};k+2;-\frac{1+K}{\overline{\gamma}E}\right] \\ &\quad-\frac{2\sqrt{E/\pi}}{\left[\beta_1(1-\rho^2)\right]^{k+3/2}}\sum_{\rho=0}^\infty\frac{\left[-(1+K)/\overline{\gamma}\right]^\rho E^\rho}{(2\rho+1)\rho!} \\ &\quad\times\Gamma\left[k+\frac{3}{2}+\rho,\frac{(1+K)\gamma_\tau}{\overline{\gamma}}\right]-\frac{\Gamma\left[k+1,(1+K)\gamma_\tau/\overline{\gamma}\right]}{2\left[\beta_1(1-\rho^2)\right]^{k+1}}\Bigg\}.\end{aligned} \tag{25}$$

In (25), $\Gamma(\cdot,\cdot)$ is the upper incomplete Gamma function [22, Equation (6.51)].

(ii) For noncoherent binary frequency shift keying (BFSK) and binary differential phase shift keying (BDPSK), $P_e(\gamma) = D\exp(-D\gamma)$. Similar to the derivation of (12), that is, using [21, Equation (3.381/4)] and [2, Equation (4.33)], $\overline{P}_{\mathrm{se}}$ can be expressed as

$$\begin{aligned}\overline{P}_{\mathrm{se}} = \sum_{\substack{i,h=0\\ v_1+v_2+v_3=i}}^\infty & \mathcal{A}D \\ &\times\Bigg[\frac{\Gamma(\beta_2)\mathcal{B}}{(\beta_1+E)^{\beta_2}\beta_1^{\beta_3}}\gamma(\beta_3,\beta_1\gamma_\tau) \\ &\quad+\frac{\mathcal{C}\Gamma(\beta_2+1/2)}{(\beta_1+E)^{\beta_2+1/2}\beta_1^{\beta_3+1/2}}\gamma\left(\beta_3+\frac{1}{2},\beta_1\gamma_\tau\right)\Bigg] \\ &+Q_1\left[\sqrt{\frac{2K(1+K)}{1+K+\overline{\gamma}E}},\sqrt{\frac{2(1+K+\overline{\gamma}E)\gamma_\tau}{\overline{\gamma}}}\right] \\ &\times\exp\left[\frac{K(1+K)}{1+K+\overline{\gamma}E}\right]\frac{(1+K)\exp(-K)}{1+K+\overline{\gamma}E}.\end{aligned} \tag{26}$$

(iii) For Gray encoded M-ary PSK and M-ary DPSK, $P_e(\gamma) = D\int_0^\Lambda \exp[-E(\theta)\gamma]d\theta$, where Λ is constant. Thus, $\overline{P}_{\mathrm{se}}$ can be expressed as

$$\begin{aligned}\overline{P}_{\mathrm{se}} = \sum_{\substack{i,h=0\\ v_1+v_2+v_3=i}}^\infty & \mathcal{A}D \\ &\times\Bigg\{\frac{\mathcal{B}\gamma(\beta_3,\beta_1\gamma_\tau)}{\beta_1^{\beta_3}}\int_0^\Lambda\frac{\Gamma(\beta_2)}{\left[\beta_1+E(\theta)\right]^{\beta_2}}d\theta \\ &\quad+\frac{\mathcal{C}\gamma(\beta_3+1/2,\beta_1\gamma_\tau)}{\beta_1^{\beta_3+1/2}} \\ &\quad\times\int_0^\Lambda\frac{\Gamma(\beta_2+1/2)}{\left[\beta_1+E(\theta)\right]^{\beta_2+1/2}}d\theta\Bigg\} \\ &+\int_0^\Lambda Q_1\left[\sqrt{\frac{2K(1+K)}{g(\theta)}},\sqrt{\frac{2g(\theta)\gamma_\tau}{\overline{\gamma}}}\right] \\ &\times\exp\left[\frac{K(1+K)}{g(\theta)}\right]\frac{(1+K)\exp(-K)}{g(\theta)}d\theta,\end{aligned} \tag{27}$$

where $g(\theta) = 1+K+\overline{\gamma}E(\theta)$. The above finite integrals can be easily evaluated via numerical integration.

5.3. *Average output SNR (ASNR) and amount of fading (AoF)*

The ASNR, $\overline{\gamma}_{\mathrm{ssc}}$, is a useful performance measure serving as an excellent indicator for the overall system fidelity and can be obtained from the first-order moment of γ_{ssc} as

$$\overline{\gamma}_{\mathrm{ssc}} = \mu_{\gamma_{\mathrm{ssc}}}(1). \tag{28}$$

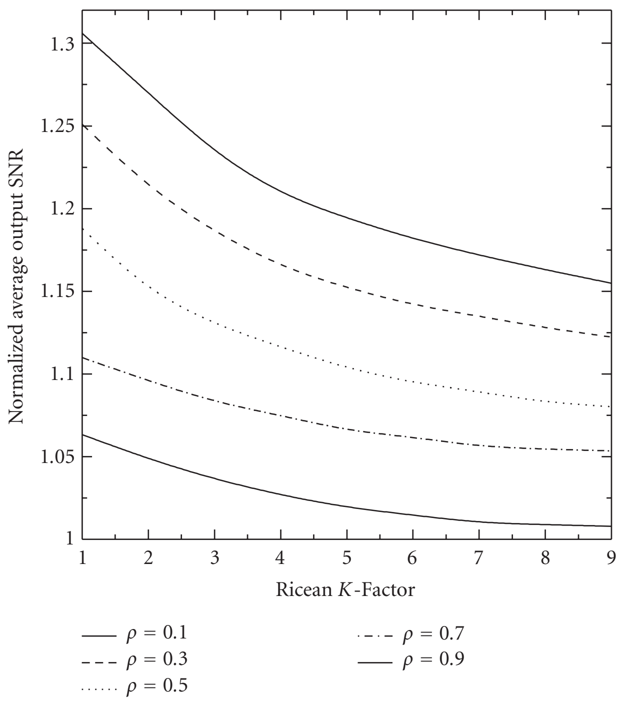

Figure 1: Normalized average output SNR (ASNR) versus the Ricean K-factor for several values of the correlation coefficient ρ.

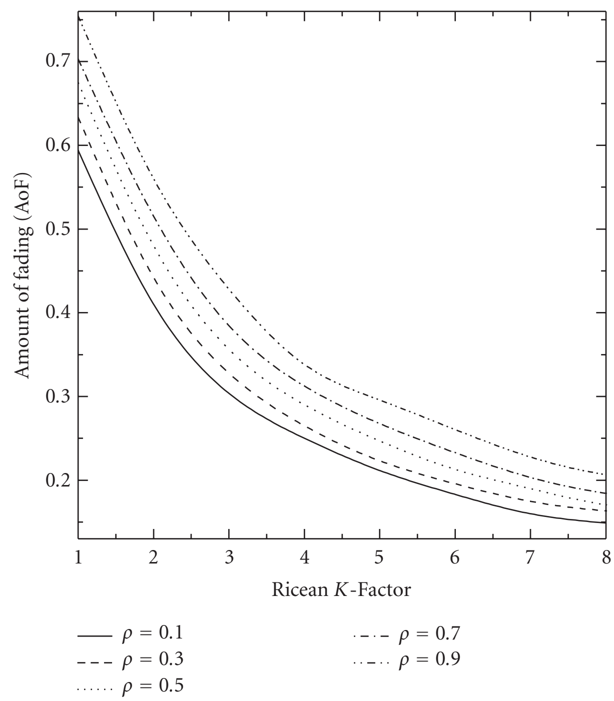

Figure 2: Amount of fading (AoF) versus the Ricean K-factor for several values of the correlation coefficient ρ.

The AoF, defined as $\text{AoF} \triangleq \text{var}(\gamma_{\text{ssc}})/\overline{\gamma}_{\text{ssc}}^2$, is a unified measure of the severity of the fading channel [2] and gives an insight to the performance of the entire system. It can be expressed in terms of first- and second-order moments of γ_{ssc} as

$$\text{AoF} = \frac{\mu_{\gamma_{\text{ssc}}}(2)}{\mu_{\gamma_{\text{ssc}}}(1)^2} - 1. \tag{29}$$

6. PERFORMANCE EVALUATION RESULTS

Using the previous mathematical analysis, various performance evaluation results have been obtained by means of numerical techniques and will be presented in this section. Such results include performances for the ASNR, AoF, P_{out}, ABEP[1], and $\overline{C}$ and will be presented for different Ricean channel conditions, that is, different values for K and ρ, as well as for various modulation schemes.

In Figures 1 and 2 the normalized ASNR ($\overline{\gamma}_{\text{ssc}}/\overline{\gamma}$) and AoF are plotted as functions of the Ricean K-factor for several values of the correlation coefficient ρ. These performance evaluation results have been obtained by numerically evaluating (15)–(17), (28), and (29). The results presented in Figure 1 show that as K increases, that is, the severity of the fading decreases, and/or ρ increases, the normalized ASNR decreases, resulting in a reduced diversity gain. We note that similar observations have been made in [3, 30]. Furthermore, the results presented in Figure 2 indicate that as K increases and/or ρ decreases, AoF is degraded.

Next the ABEP has been obtained using (23)–(27). In Figures 3 and 4 the ABEP is plotted as a function of the average input SNR per bit, that is, $\overline{\gamma}_b = \overline{\gamma}/\log_2 M$, for several values of K. Figure 3 considers the performance of DBPSK, BPSK, and M-ary PSK signaling formats and $\rho = 0.5$. As expected, when K increases, the ABEP improves and BPSK exhibits the best performance. Figure 4 presents the ABEP of 16-QAM for different values of ρ and K. For comparison purposes, the performance of an equivalent single receiver, that is, without diversity, is also included. Similar to the previous cases, it is observed that the ABEP improves as K increases and/or ρ decreases, while significant overall performance improvement, as compared to the no-diversity case, is also noted.

In Figure 5, P_{out} is plotted as a function of the normalized outage threshold per bit, $\gamma_{\text{th}}/\overline{\gamma}_b$, for several values of K and ρ. These performance results have been obtained by numerically evaluating (10), (11), and (21) and for $\rho = 0$ they are identical to the ones obtained by using [2, Equation 9.273]. It is observed that P_{out} decreases, that is, the outage performance improves, as K increases and/or ρ decreases.

Finally, the normalized average CC can be obtained as $\widetilde{C} = \overline{C}/BW$ (in b/s/Hz) by employing (19) and (20). In Figure 6, $\widetilde{C}$ is plotted as a function of $\overline{\gamma}_b$ for several values

[1] For the consistency of the presentation from now on instead of the ASEP the ABEP performance will be used. As it is well known [2] for M-ary ($M > 2$) modulation schemes, assuming Gray encoding, the ABEP can be simply obtained from the ASEP as $\overline{P}_{\text{be}} \cong \overline{P}_{\text{se}}/\log_2 M$, since $E_s = E_b \log_2 M$, where E_b denotes the transmitted average bit energy.

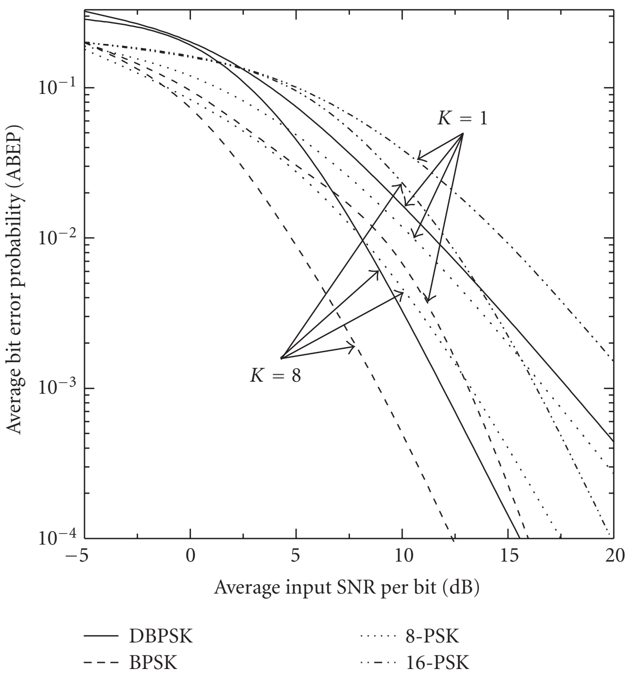

Figure 3: Average bit error probability (ABEP) versus average input SNR per bit for DBPSK, BPSK, and M-PSK ($M = 8$ and 16) signaling formats, for different values of the Ricean K-factor.

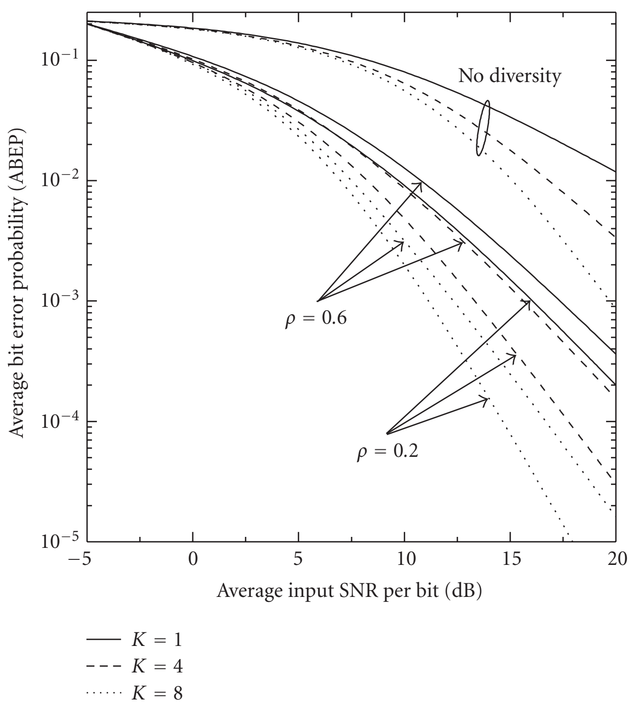

Figure 4: Average bit error probability (ABEP) versus average input SNR per bit for 16-QAM signaling format for different values of the Ricean K-factor and the correlation coefficient ρ.

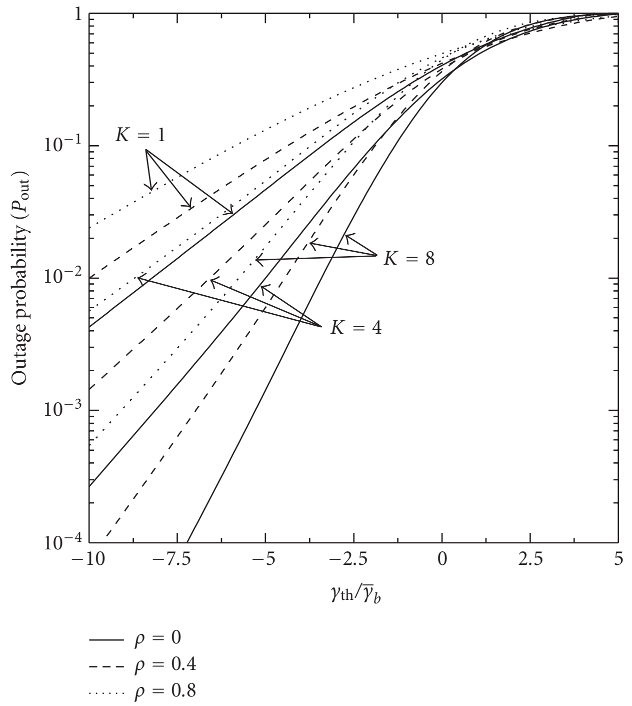

Figure 5: Outage probability (P_{out}) versus the normalized average input SNR per bit for several values of the Ricean K-factor and the correlation coefficient ρ.

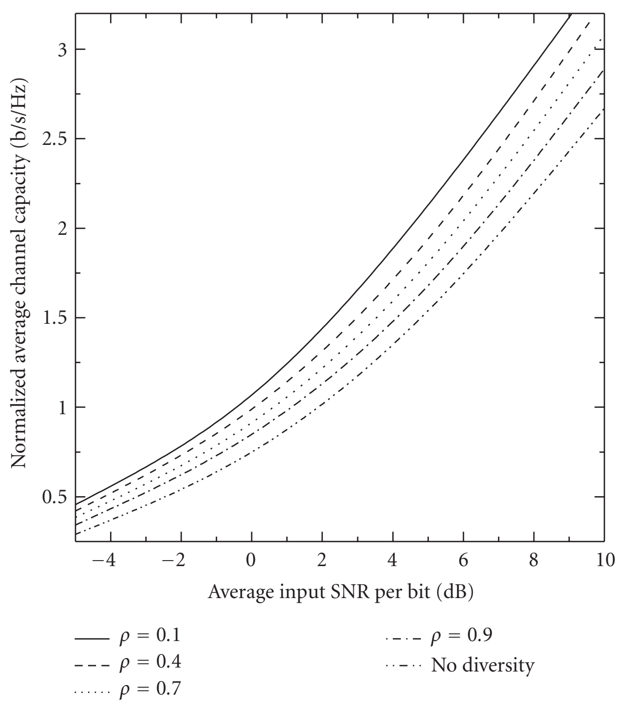

Figure 6: Normalized average channel capacity ($\overline{C}/BW$) versus the average input SNR per bit for several values of the correlation coefficient ρ.

of ρ and for $K = 1$. These results illustrate that as ρ increases, $\widetilde{C}$ decreases, as expected [12], and the receiver without diversity has always the worst performance.

7. CONCLUSIONS

In this paper, the performance of dual branch SSC diversity receivers operating over correlated Ricean fading channels has been studied. By deriving a convenient expression for the bivariate Ricean PDF, analytical formulae for the most important statistical metrics of the received signals and the capacity of such receivers have been obtained. Capitalizing on these formulas, useful expressions for a number of performance criteria have been obtained, such as ABEP, P_{out}, ASNR, AoF, and average CC. Various performance evaluation results for different fading channel conditions have been also presented and compared with equivalent performance results of receivers without diversity.

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