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Research Article

SSC Diversity Receiver over Correlated α-μ Fading Channels in the Presence of Cochannel Interference

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This paper studies the performances of a dual-branch switched-and-stay combining (SSC) diversity receiver, operating over correlated α-μ fading in the presence of cochannel interference (CCI). Very useful, novel, infinite series expressions are obtained for the output signal to interference ratio's (SIR's) probability density function (PDF) and cumulative distribution function (CDF). The performance analysis is based on an outage probability (OP) and an average bit error probability (ASEP) criteria. ASEP is efficiently evaluated for modulation schemes such as noncoherent frequency-shift keying (NCFSK) and binary differentially phase-shift keying (BDPSK). The effects of various parameters, such as input SIR unbalance, the level of correlation between received desired signals and interferences, nonlinearity of the environment, and fading severity on systems performances are graphically presented and analyzed.

1. Introduction

Various techniques for reducing the fading effects and the influence of the cochannel interference (CCI) are used in wireless communication systems [1]. The two main goals of the diversity techniques are (1) upgrading the transmission reliability without increasing transmission power and bandwidth and (2) increasing the channel capacity. Space diversity is an efficient method for amelioration of system's quality of service (QoS) when multiple receiver antennas are used. There are several principal types of combining techniques and division, that can be generally performed depending on the complexity restriction put on the communication system and the amount of channel state information available at the receiver.

However, switch and stay combining (SSC) is the least complex and can be used in conjunction with coherent, noncoherent, and differentially coherent modulation schemes. In general, with 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 [1–3]. In fading environments as in cellular systems, where the level of CCI is sufficiently high as compared to the thermal noise, SSC selects a particular branch until its signal-to-interference ratio (SIR) drops below a predetermined threshold (SIR-based switched diversity). Then the combiner switches to another branch and stays there regardless of SIR of that branch. This type of diversity, in which switching is based on the SIR value, can be measured in real time both in base stations (uplink) and in mobile stations (downlink) using specific SIR estimators.

The multipath fading in wireless communications is modelled by several distributions including Weibull, Nakagami-m Hoyt, Rayleigh, and Rice. By considering two important phenomena inherent to radio propagation, namely, nonlinearity and clustering, the α-μ fading model was proposed and discussed in [4–7]. The model provides a very good fit to measured data over a wide range of fading conditions. The α-μ distribution is written in terms of physicallybased fading parameters, namely, α and μ. Roughly speaking, α is related to the nonlinearity of the environment, whereas μ is associated with the number of multipath clusters. The diversity reception over α-μ fading channels was previously discussed in [8–11].

There is a number of papers concerning performance analysis of SSC receivers, for example, [12–15]. The performance analysis of SSC diversity receivers, operating over correlated Ricean fading satellite channels, can be found in [13], where the performance is evaluated based on a bunch of novel analytical formulae for the outage probability (OP), average symbol error probability (ASEP), channel capacity (CC), the amount of fading (AoF), and the average output SNR obtained in infinite series form. The similar performance analysis of the switched diversity receivers operating over correlated Weibull fading channels in terms of OP, ASEP, moments, and moment generating function (MGF) can be found in [14]. Reference [15] studies the performance of a dual-branch SSC diversity receiver with the switching decision based on SIR, operating over correlated Ricean fading channels in the presence of correlated Nakagami-m distributed CCI. Moreover to the best of author's knowledge, no analytical study of switch and stay combining involving assumed correlated α-μ fading for both desired signal and cochannel interference has been reported in the literature.

In this paper, desired signal and CCI are considered to be affected by α-μ fading distribution, which is adequate for multipath waves, propagating in a nonhomogenous environment. An approach to the performance analyses of given SSC diversity receiver is presented. Effectiveness of any modulation scheme and the type of diversity used is studied thought evaluating the system's performance over the channel conditions. Infinite series expressions for probability distribution function (PDF) and cumulative distribution function (CDF) of the output SIR for SSC diversity are derived. Furthermore, infinite series expression for important performance measure, such as OP, is obtained. An OP is shown graphically for different system parameters. Using infinite series formulae, ASEP is efficiently evaluated for several modulation schemes such as noncoherent frequency-shift keying (NCFSK) and binary differentially phase-shift keying (BDPSK). This analysis has a high level of generality because α-μ fading distribution model includes as special cases other important distributions such as Weibull and Nakagami-m (therefore the one-sided Gaussian and Rayleigh are also special cases of it).

2. System Model and Received Statistics

Independent fading assumes antenna elements in diversity systems to be placed sufficiently apart, which is not general case in practice due to insufficient spacing between antennas. When diversity system is applied on small terminals with multiple antennas, correlation arises between branches. Now, due to insufficient antennae spacing, desired signal envelopes R_1 and R_2 experience correlative α-μ fading, with joint distribution [5]:

$$
\begin{aligned}
f_{R_1,R_2}(R_1,R_2) = {} & \frac{\alpha_1 {\mu_1}^{\mu_1} {R_1}^{\alpha_1\mu_1-1}}{\hat{R}_1^{\alpha_1\mu_1}\Gamma(\mu_1)} \exp\left(-\mu_1 \frac{{R_1}^{\alpha_1}}{\hat{R}_1^{\alpha_1}}\right) \\
& \times \frac{\alpha_2 {\mu_1}^{\mu_1} R_2^{\alpha_2\mu_1-1}}{\hat{R}_2^{\alpha_2\mu_1}\Gamma(\mu_1)} \exp\left(-\mu_1 \frac{{R_2}^{\alpha_2}}{\hat{R}_2^{\alpha_2}}\right) \\
& \times \sum_{l=0}^{\infty} \frac{l!\Gamma(\mu_1)}{\Gamma(\mu_1+l)} {\rho_{12d}}^{l} {L_l}^{\mu_1-1} \\
& \times \left(\frac{\mu_1 {R_1}^{\alpha_1}}{\hat{R}_1^{\alpha_1}}\right) {L_l}^{\mu_1-1}\left(\frac{\mu_1 {R_2}^{\alpha_2}}{\hat{R}_2^{\alpha_2}}\right).
\end{aligned} \tag{1}
$$

Similarly, due to insufficient antennae spacing, envelopes of interference r_1 and r_2 are correlated with joint α-μ distribution [5]:

$$
\begin{aligned}
f_{r_1,r_2}(r_1,r_2) = {} & \frac{\alpha_1 {\mu_2}^{\mu_2} {r_1}^{\alpha_1\mu_2-1}}{\hat{r}_1^{\alpha_1\mu_2}\Gamma(\mu_2)} \exp\left(-\mu_2 \frac{{r_1}^{\alpha_1}}{\hat{r}_1^{\alpha_1}}\right) \\
& \times \frac{\alpha_2 {\mu_2}^{\mu_2} r_2^{\alpha_2\mu_2-1}}{\hat{r}_2^{\alpha_2\mu_2}\Gamma(\mu_2)} \exp\left(-\mu_2 \frac{{r_2}^{\alpha_2}}{\hat{r}_2^{\alpha_2}}\right) \\
& \times \sum_{k=0}^{\infty} \frac{k!\Gamma(\mu_2)}{\Gamma(\mu_2+k)} {\rho_{12c}}^{k} {L_k}^{\mu_2-1} \\
& \times \left(\frac{\mu_2 {r_1}^{\alpha_1}}{\hat{r}_1^{\alpha_1}}\right) {L_k}^{\mu_2-1}\left(\frac{\mu_2 {r_2}^{\alpha_2}}{\hat{r}_2^{\alpha_2}}\right).
\end{aligned} \tag{2}
$$

It is important to quote that ρ_{d12} and ρ_{c12} denote the desired and interfering signal power correlation coefficients defined as $\mathrm{cov}(R_i^{\alpha_i}, R_j^{\alpha_j})/(\mathrm{var}(R_i^{\alpha_i})\mathrm{var}(R_j^{\alpha_j}))^{1/2}$, and $\mathrm{cov}(r_i^{\alpha_i}, r_j^{\alpha_j})/(\mathrm{var}(r_i^{\alpha_i})\mathrm{var}(r_j^{\alpha_j}))^{1/2}$, respectively. For the desired signal with correlated envelopes R_i and the arbitrary power parameters α_i, connected with the nonlinearity of the environment, $\alpha_i > 0$, explained in [5], $\hat{R}_i = \sqrt[\alpha_i]{E({R_i}^{\alpha_i})}$ denote the α_i-root mean values of desired signal. Similarly $\hat{r}_i = \sqrt[\alpha_i]{E({r_i}^{\alpha_i})}$ denote the α_i-root mean values of interferer signal. Parameters μ_1 and μ_2 are associated with the number of multipath clusters through which desired and interference signals propagate, and they are defined in the terms of normalized variances as in [5, 11]. Cov$(\cdot)$ and var $(\cdot)$ are, respectively, the covariance and variance operators, and $\Gamma(\cdot)$ stands for Gamma function [16]. $L_n^k(x)$ defines generalized Laguerre polynomial with the property of [17]

$$
L_n^k(x) = \sum_{m=0}^{n} (-1)^m \frac{(n+k)!}{(n-m)!(k+m)!(m)!} x^m. \tag{3}
$$

Let $z_1 = R_1^2/r_1^2$ and $z_2 = R_2^2/r_2^2$ represent the instantaneous SIR on the diversity branches, respectively [18]. The joint PDF of z_1 and z_2 and can be expressed by [18, 19]

$$f_{z_1,z_2}(z_1,z_2) = \frac{1}{4\sqrt{z_1 z_2}} \int_0^\infty \int_0^\infty p_{R_1,R_2}(r_1\sqrt{z_1}, r_2\sqrt{z_2}) p_{r_1,r_2}(r_1,r_2) r_1 r_2 dr_1 dr_2. \tag{4}$$

Substituting (1), (2), and (3) in (4), we get

$$f_{z_1,z_2}(z_1,z_2) = \sum_{l=0}^{\infty}\sum_{k=0}^{\infty}\sum_{j=0}^{l}\sum_{m=0}^{l}\sum_{n=0}^{k}\sum_{p=0}^{k} \frac{G_1 {z_1}^{\alpha_1(\mu_1+j)/2-1} {z_2}^{\alpha_2(\mu_1+m)/2-1}}{\left(\mu_1 {z_1}^{\alpha_1/2} + {S_1}^{\alpha_1/2}\right)^{\mu_1+\mu_2+j+n} \mathcal{A}}, \tag{5}$$

where $\mathcal{A}$ denotes $(\mu_2 {z_2}^{\alpha_2/2} + {S_2}^{\alpha_2/2})^{\mu_1+\mu_2+m+p}$ with

$$G_1 = \frac{(-l)_m(-l)_j(-k)_n(-k)_p \mathcal{B}}{4 j! m! n! p! k! l!} \times \frac{\Gamma(\mu_1+\mu_2+n+j)\Gamma(\mu_1+\mu_2+p+m)}{\Gamma(\mu_1)\Gamma(\mu_2)\Gamma(\mu_1+l)\Gamma(\mu_2+k)} \times \frac{\mathcal{C}}{\mathcal{D}}, \tag{6}$$

where $\mathcal{B}$ denotes $\alpha_1\alpha_2{\mu_1}^{2\mu_1+j+m}{\mu_2}^{2\mu_2+n+p}{\rho_{d12}}^l{\rho_{c12}}^k$, $\mathcal{C}$ denotes $((\mu_1-1+l)!)^2((\mu_2-1+k)!)^2 \; {S_1}^{\alpha_1(\mu_1+n)/2-1} {S_2}^{\alpha_2(\mu_2+p)/2\,-1}$, $\mathcal{D}$ denotes $(\mu_1-1+j)!(\mu_1-1+m)!(\mu_2-1+n)!(\mu_2-1+p)!$, with $(a)_x$ denoting the Pochhammer symbol [16]. Derivation of this expression is presented in the appendix. Let z_{ssc} represent the instantaneous SIR at the SSC output, and z_τ the predetermined switching threshold for the both input branches. Following [20], the PDF of SIR is given by

$$f_{z_{\text{ssc}}}(z) = \begin{cases} v_{\text{ssc}}(z), & z \le z_\tau, \\ v_{\text{ssc}}(z) + f_z(z), & z > z_\tau, \end{cases} \tag{7}$$

where $v_{\text{ssc}}(z)$, according to [20], can be expressed as $v_{\text{ssc}}(z) = \int_0^{z_\tau} f_{z_1,z_2}(z,z_2)dz_2$. Moreover, $v_{\text{ssc}}(z)$ can be expressed as infinite series

$$v_{\text{ssc}}(z) = \sum_{l=0}^{\infty}\sum_{k=0}^{\infty}\sum_{j=0}^{l}\sum_{m=0}^{l}\sum_{n=0}^{k}\sum_{p=0}^{k} G_2 \frac{z^{\alpha_1(\mu_1+j)/2-1}}{\mathcal{E}} G_3^{\mu_2+m} \times {}_2F_1(\mu_2+m, 1-\mu_1-p; 1+\mu_2+m; G_3), \tag{8}$$

where $\mathcal{E}$ denotes $(\mu_1 z^{\alpha_1/2} + \mu_2 S_1^{\alpha_1/2})^{\mu_1+\mu_2+j+n}$ with ${}_2F_1(u_1,u_2;u_3;x)$, being the Gaussian hypergeometric function [16], and

$$G_2 = \frac{(-l)_m(-l)_j(-k)_n(-k)_p \alpha_1 \mu_1^{\mu_1+j}\mu_2^{\mu_2+n}\rho_{d12}^l\rho_{c12}^k}{2 j! m! n! p! l! k! (\mu_2+m)} \times \frac{\Gamma(\mu_1+\mu_2+n+j)\Gamma(\mu_1+\mu_2+p+m)}{\Gamma(\mu_1)\Gamma(\mu_2)\Gamma(\mu_1+l)\Gamma(\mu_2+k)} \times \frac{((\mu_1-1+l)!)^2 S_1^{\alpha_1(\mu_1+n)/2}}{(\mu_1-1+j)!(\mu_1-1+m)!} \times \frac{((\mu_2-1+k)!)^2}{(\mu_2-1+n)!(\mu_2-1+p)!},$$
$$G_3 = \frac{\mu_1 z_\tau^{\alpha_2/2}}{\mu_2 S_2^{\alpha_2/2} + \mu_1 z_\tau^{\alpha_2/2}}. \tag{9}$$

In the same manner, the $f_z(z)$ can be expressed as

$$f_z(z) = \frac{\alpha_1 \mu_1^{\mu_1}\mu_2^{\mu_2} z^{\alpha_1\mu_1/2-1}\Gamma(\mu_1+\mu_2)}{2\left(\mu_1 z^{\alpha_1/2} + \mu_2 S_1^{\alpha_1/2}\right)^{\mu_1+\mu_2}\Gamma(\mu_1)\Gamma(\mu_2)}. \tag{10}$$

Similarly, the CDF of the SSC output SIR, that is, the $F_{z_{\text{ssc}}}(z)$ is given by [11, 20]

$$F_{z_{\text{ssc}}}(z) = \begin{cases} F_{z_1,z_2}(z,z_\tau), & z \le z_\tau, \\ F_z(z) - F_z(z_\tau) + F_{z_1,z_2}(z,z_\tau), & z > z_\tau. \end{cases} \tag{11}$$

By substituting (5) in $F_{z_1,z_2}(z,z_\tau) = \int_0^z \int_0^{z_\tau} f_{z_1,z_2}(z_1,z_2)dz_1dz_2$ and (10) in $F_z(z) = \int_0^z f_z(z)dz$ and $F_z(z)$ can be expressed as the following infinite series, respectively,

$$F_{z_1,z_2}(z,z_\tau) = \sum_{l=0}^{\infty}\sum_{k=0}^{\infty}\sum_{j=0}^{l}\sum_{m=0}^{l}\sum_{n=0}^{k}\sum_{p=0}^{k} G_4 G_5^{\mu_1+l} G_3^{\mu_1+m} \times {}_2F_1(\mu_1+j, 1-\mu_2-n; 1+\mu_1+j; G_5) \times {}_2F_1(\mu_2+m, 1-\mu_1-p; 1+\mu_2+m; G_3), \tag{12}$$

$$F_z(z) = \frac{\Gamma(\mu_1+\mu_2)}{\Gamma(\mu_1)\Gamma(\mu_2)\Gamma(\mu_2+j)} G_5^{\mu_2+j} \times {}_2F_1(\mu_2+j, 1-\mu_1-n; 1+\mu_2+j; G_5), \tag{13}$$

$$F_z(z_\tau) = \frac{\Gamma(\mu_1+\mu_2)}{\Gamma(\mu_1)\Gamma(\mu_2)\Gamma(\mu_2+m)} G_3^{\mu_2+m} \times {}_2F_1(\mu_2+m, 1-\mu_1-p; 1+\mu_2+m; G_3), \tag{14}$$

Table 1: Terms need to be summed in (12) to achieve accuracy at the 7th significant digit.

$\mu_1 = 2, \mu_2 = 1.5, \alpha_1 = 1, \alpha_1 = 3, S_1 = S_2 = z\tau$		$S_1/z = 0\,\text{dB}$	$S_1/z = -10\,\text{dB}$
$\rho_d = 0.2$	$\rho_c = 0.2$	5	5
$\rho_d = 0.3$	$\rho_c = 0.3$	5	7
$\rho_d = 0.4$	$\rho_c = 0.2$	5	9
$\rho_d = 0.4$	$\rho_c = 0.3$	6	9
$\rho_d = 0.4$	$\rho_c = 0.4$	6	9
$\rho_d = 0.4$	$\rho_c = 0.5$	8	9
$\rho_d = 0.4$	$\rho_c = 0.6$	9	9
$\rho_d = 0.5$	$\rho_c = 0.5$	9	10
$\rho_d = 0.6$	$\rho_c = 0.6$	10	11

with

$$G_4 = \frac{(-l)_m(-l)_j(-k)_n(-k)_p \mu_1^{2\mu_1+j+p} \mu_2^{2\mu_2+n+m} \rho_{d12}^l}{j!m!n!p!l!k!} \times \frac{\rho_{c12}^k \Gamma(\mu_1+\mu_2+n+j)\Gamma(\mu_1+\mu_2+p+m)}{\Gamma(\mu_1)\Gamma(\mu_2)\Gamma(\mu_1+l)\Gamma(\mu_2+k)} \times \frac{((\mu_1-1+l)!)^2}{(\mu_1+j)(\mu_2+m)(\mu_1-1+j)!(\mu_1-1+m)!} \times \frac{((\mu_2-1+k)!)^2}{(\mu_2-1+n)!(\mu_2-1+p)!}, \tag{15}$$

$$G_5 = \frac{\mu_1 z^{\alpha_1/2}}{\mu_2 S_1^{\alpha_1/2} + \mu_2 z^{\alpha_1/2}}. \tag{16}$$

The nested infinite sums in (8) and (12) converge for any defined value of the parameters ρ_{d12}, ρ_{c12}, μ_1, μ_2, α_1, α_2, S_1, and S_2. Numbers of terms that need to be summed in (12) to achieve accuracy at the 7th significant digit are presented in Table 1. As is shown in this table, the number of the terms that need to be summed to achieve a desired accuracy depends strongly on the correlation coefficient ρ_{d12} and ρ_{c12}. The number of the terms increases as correlation coefficient increases. Convergence becomes slow and we need much more terms when ρ_{12d} and ρ_{12c} are higher and closer to 1. For the special cases of $\mu_1 = 1$, $\mu_2 = 1$, and for the special case of $\alpha_1 = 2$, $\alpha_1 = 2$, we can evaluate CDF and PDF expressions for Weibull and for Nakagami-m desired signal and cochannel interference, respectively. This generality of CDF of output SIR for a number of fading distributions is the main contribution of our work.

3. Outage Probability

Since OP is defined as probability that the instantaneous SIR of the system falls below a specified threshold value, it can be expressed in terms of the CDF of z_{ssc}, that is, as $P_{\text{out}} = F_{z_{\text{ssc}}}(z^*)$, where z^* is the specified threshold value. Using (11)–(15) the Pout performances results have been obtained. These results are presented in Figure 1, as the function of the normalized outage threshold (dB) for several values of ρ_{d12}, ρ_{c12}, μ_1, μ_2, α_1, and α_2. Normalized outage threshold (dB) is defined as being the average SIR's at the input branch of the balanced dual-branch switched-and-stay combiner, normalized by specified threshold value z^*. Results show that as the signal and interference correlation coefficients, ρ_{d12} and ρ_{c12}, increase and normalized outage threshold decreases, OP increases.

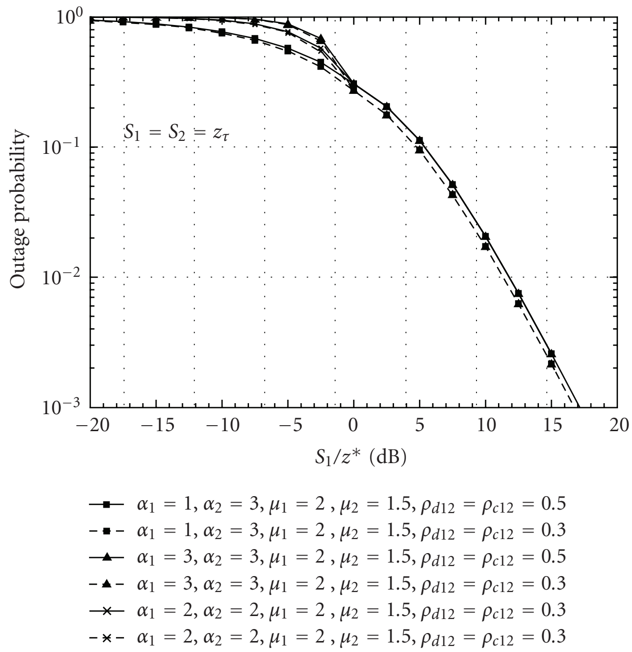

Figure 1: Outage probability (P_{out}) versus normalized outage threshold for the balanced dual-branch SSC diversity receiver and different values of ρ_{12d}, ρ_{12c}, μ_1, μ_2, α_1 and α_2.

4. Average Symbol Error Probability

The ASEP ($\overline{P}_e$) can be evaluated by averaging the conditional symbol error probability for a given SIR, that is, $P_e(z)$, over the PDF of z_{ssc}, that is, $f_{z_{\text{ssc}}}(z)$ [8]:

$$\overline{P}_e = \int_0^\infty P_e(z) f_{z_{\text{ssc}}}(z)\,dz, \tag{17}$$

where $P_e(z)$ depends on applied modulation scheme. For BDPSK and NCFSK modulation schemes the conditional symbol error probability for a given SIR threshold can be expressed by $P_e(z) = 1/2\exp(-\lambda z)$, where $\lambda = 1$ for BDPSK and $\lambda = 1/2$ for NCFSK [14]. Hence, substituting (7) into (17) gives the following ASEP expression for the considered dual-branch SSC receiver:

$$\overline{P}_e = \int_0^\infty P_e(z) v_{z_{\text{ssc}}}(z)\,dz + \int_{z_\tau}^\infty P_e(z) f_z(z)\,dz. \tag{18}$$

Using the previously derived infinite series expressions, we present representative numerical performance evaluation results of the studied dual-branch SSC diversity receiver, such as ASEP in case of two modulation schemes, NCFSK and

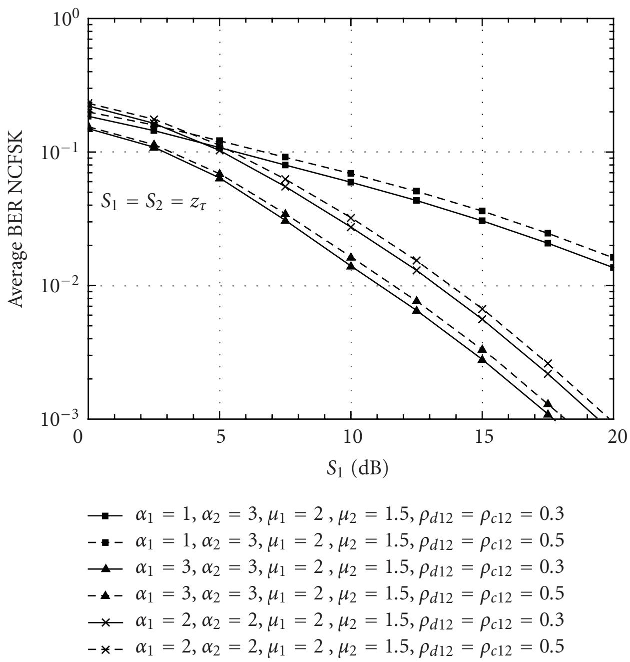

Figure 2: Average bit error probability (ABER) versus average SIRs at the input branches of the balanced dual-branch switched-and-stay combiner, S_1, for NCFSK modulation scheme and several values of ρ_{12d}, ρ_{12c}, μ_1, μ_2, α_1, and α_2.

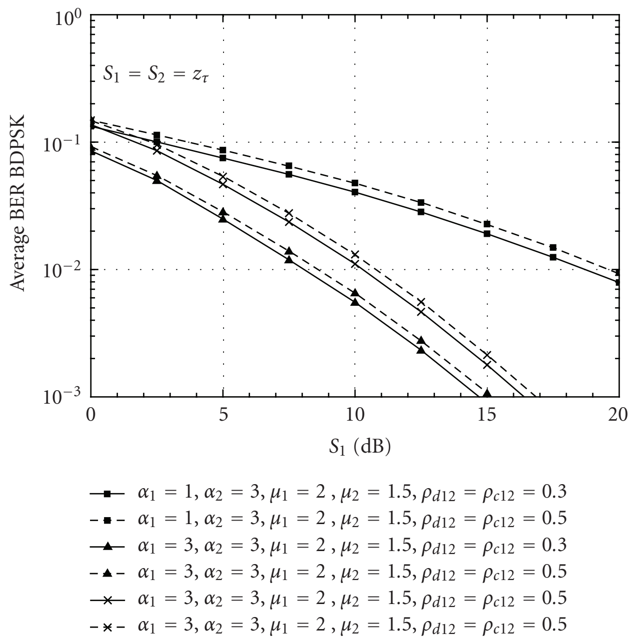

Figure 3: Average bit error probability (ABER) versus average SIR's at the input branches of the balanced dual-branch switched-and-stay combiner, S_1, for BDPSK modulation scheme and several values of ρ_{12d}, ρ_{12c}, μ_1, μ_2, α_1, and α_2.

BDPSK. Applying (18) on BDPSK and NCFSK modulation schemes, the ASEP performance results have been obtained, for several values of ρ_{d12}, ρ_{c12}, μ_1, μ_2, α_1 and α_2, as a function of the average SIRs at the input branches of the balanced dual-branch switched-and-stay combiner, that is, S_1. These results are plotted in Figures 2 and 3. It is shown that while as the signal and interference correlation coefficients, ρ_{d12} and ρ_{c12} increase and the average SIR's at the at the input branches increase, the ASEP increases at the same time. Comparison of Figures 2 and 3 shows better performance of BDPSK modulation scheme versus NCFSK modulation scheme.

5. Conclusion

In this paper, the system performances of a dual-branch SSC diversity receiver over correlated α-μ channels are analyzed based on SIR. Fading between the diversity branches and between interferers is correlated and α-μ distributed. The complete statistics for the SSC output SIR is given in the infinite series expressions form, that is, PDF, CDF, and OP. Using these new formulae, ASEP was efficiently evaluated for some modulation schemes such as DPSK and NCFSK. The main contribution of this analysis for dualbranch signal combiner is that it has been done for a general case of α-μ (Generalized Gamma) distribution, which includes as special cases important other distributions such as Weibull and Nakagami-m (therefore, the One-Sided Gaussian and Rayleigh are also special cases of it), so our analysis has a high level of generality.

Appendix

Substituting (1), (2), and (3) in (4) results in

$$p_{\lambda_1,\lambda_2}(\lambda_1,\lambda_2) = \sum_{l=0}^{\infty}\sum_{k=0}^{\infty}\sum_{j=0}^{l}\sum_{m=0}^{l}\sum_{n=0}^{k}\sum_{p=0}^{k} C_1 \times \int_0^\infty\int_0^\infty {r_1}^{\alpha_1(\mu_1+\mu_2+j+n)-1}{r_2}^{\alpha_2(\mu_1+\mu_2+m+p)-1} \times \exp\left(-\frac{\mu_1 {r_1}^{\alpha_1}{\lambda_1}^{\alpha_1/2}}{\widehat{R}_1^{\alpha_1}}\right)\exp\left(-\frac{\mu_1 {r_2}^{\alpha_2}{\lambda_2}^{\alpha_2/2}}{\widehat{R}_2^{\alpha_2}}\right) \times \exp\left(-\mu_2\frac{{r_1}^{\alpha_1}}{\widehat{r}_1^{\alpha_1}}\right)\exp\left(-\mu_2\frac{{r_2}^{\alpha_2}}{\widehat{r}_2^{\alpha_2}}\right)dr_1\,dr_2 \tag{A.1}$$

with constant C_1 given as

$$C_1 = \frac{(-1)^{j+m+n+p}(\alpha_1\alpha_2)^2{\mu_1}^{2\mu_1+j+m}{\mu_2}^{2\mu_2+n+p}l!k!{\rho_{d12}}^{l}{\rho_{c12}}^{k}}{4j!m!n!p!(l-j)!(l-m)!(k-n)!(k-p)!} \times \frac{\lambda_1^{\alpha_1(\mu_1+j)/2-1}\lambda_2^{\alpha_2(\mu_1+m)/2-1}}{\widehat{r}_1^{\alpha_1(\mu_2+n)}\widehat{R}_1^{\alpha_1(\mu_1+j)}\widehat{r}_2^{\alpha_2(\mu_2+p)}\widehat{R}_2^{\alpha_2(\mu_1+m)}} \times \frac{((\mu_1-1+l)!)^2((\mu_2-1+k)!)^2}{(\mu_1-1+j)!(\mu_1-1+m)!\mathcal{F}}, \tag{A.2}$$

where $\mathcal{F}$ denotes $(\mu_2 - 1 + n)!(\mu_2 - 1 + p)!$ Let $S_1 = \widehat{R}_1^2/\widehat{r}_1^2$ and $S_2 = \widehat{R}_2^2/\widehat{r}_2^2$ be the average SIRs at the first and second input branches of the dual-branch selection combiner. Then, the following integrals from the pervious expression can be presented in the form:

$$\begin{aligned} I_1 &= \int_0^\infty r_1{}^{\alpha_1(\mu_1+\mu_2+j+n)-1} \\ &\quad \times \exp\left(-r_1{}^{\alpha_1}\left(\frac{\mu_1\lambda_1{}^{\alpha_1/2}+\mu_2 S_1{}^{\alpha_1/2}}{\widehat{R}_1^{\alpha_1}}\right)\right)dr_1, \\ I_2 &= \int_0^\infty r_2{}^{\alpha_1(\mu_1+\mu_2+m+p)-1} \\ &\quad \times \exp\left(-r_2{}^{\alpha_2}\left(\frac{\mu_1\lambda_2{}^{\alpha_2/2}+\mu_2 S_2{}^{\alpha_2/2}}{\widehat{R}_2^{\alpha_2}}\right)\right)dr_2. \end{aligned} \tag{A.3}$$

Now, they can easily be solved using variable substitutions

$$\begin{aligned} u &= r_1{}^{\alpha_1}\left(\frac{\mu_1\lambda_1{}^{\alpha_1/2}+\mu_2 S_1{}^{\alpha_1/2}}{\widehat{R}_1^{\alpha_1}}\right), \\ v &= r_2{}^{\alpha_2}\left(\frac{\mu_1\lambda_2{}^{\alpha_2/2}+\mu_2 S_2{}^{\alpha_2/2}}{\widehat{R}_2^{\alpha_2}}\right), \end{aligned} \tag{A.4}$$

and well-known definition of Gamma function: $\Gamma(a) = \int_0^\infty t^{a-1}\exp(-t)dt$. Finally, (12) can be written as

$$\begin{aligned} &p_{\lambda_1,\lambda_2}(\lambda_1,\lambda_2) \\ &\quad = \sum_{l=0}^{\infty}\sum_{k=0}^{\infty}\sum_{j=0}^{l}\sum_{m=0}^{l}\sum_{n=0}^{k}\sum_{p=0}^{k}\frac{G_1\lambda_1{}^{\alpha_1(\mu_1+j)/2-1}\lambda_2{}^{\alpha_2(\mu_1+m)/2-1}}{\left(\mu_1\lambda_1{}^{\alpha_1/2}+S_1{}^{\alpha_1/2}\right)^{\mu_1+\mu_2+j+n}\mathcal{G}}, \end{aligned} \tag{A.5}$$

where $\mathcal{G}$ denotes $(\mu_2\lambda_2{}^{\alpha_2/2} + S_2{}^{\alpha_2/2})^{\mu_1+\mu_2+m+p}$ with

$$\begin{aligned} G_1 &= \frac{(-1)^{j+m+n+p}\alpha_1\alpha_2\mu_1{}^{2\mu_1+j+m}\mu_2{}^{2\mu_2+n+p}l!k!\rho_{d12}{}^l\rho_{c12}{}^k}{4j!m!n!p!(l-j)!(l-m)!(k-n)!(k-p)!} \\ &\quad \times \frac{\Gamma(\mu_1+\mu_2+n+j)\Gamma(\mu_1+\mu_2+p+m)}{\Gamma(\mu_1)\Gamma(\mu_2)\Gamma(\mu_1+l)\Gamma(\mu_2+k)} \\ &\quad \times \frac{\mathcal{C}}{\mathcal{D}} \end{aligned} \tag{A.6}$$

after using the properties of Pochhammer symbol [14, 17], G_1 can be written as

$$\begin{aligned} G_1 &= \frac{(-l)_m(-l)_j(-k)_n(-k)_p\mathcal{B}}{4j!m!n!p!k!l!} \\ &\quad \times \frac{\Gamma(\mu_1+\mu_2+n+j)\Gamma(\mu_1+\mu_2+p+m)}{\Gamma(\mu_1)\Gamma(\mu_2)\Gamma(\mu_1+l)\Gamma(\mu_2+k)} \\ &\quad \times \frac{\mathcal{C}}{\mathcal{D}}. \end{aligned} \tag{A.7}$$

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