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Research Article

Performance Analysis of SSC Diversity Receiver over Correlated Ricean Fading Channels in the Presence of Cochannel Interference

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In this paper an approach to the performance analysis of a dual-branch switched-and-stay combining (SSC) diversity receiver, operating over interference-limited Ricean correlated fading environment, is presented. Infinite series expressions are obtained for the output signal-to-interference ratio's (SIR) probability density function (PDF) and cumulative distribution function (CDF). Using these new formulae, the outage probability (OP) and the average symbol error probability (ASEP) for modulation schemes such as noncoherent frequency-shift keying (NCFSK) and binary differentially phase-shift keying (BDPSK) are efficiently evaluated. Numerical results, presented into this paper, are graphically presented and analyzed, in order to point out the effects of fading severity and the level of correlation on the system performances.

1. Introduction

Space diversity reception, based on using multiple antennas at the reception, is being widely considered as a very efficient technique for mitigating fading and cochannel interference (CCI) effects and improving offered quality of service (QoS) in wireless communication systems. Various techniques for reducing the fading effects and the influence of the cochannel interference (CCI) are used in wireless communication systems [1–3]. Depending on the complexity restriction put on the communication system and the amount of channel state information available at the receiver, several principal types of space diversity techniques can be performed.

However, the least complex space diversity reception technique that can be used in conjunction with coherent, noncoherent, and differentially coherent modulation schemes and has application in many real life communication scenarios is switch-and-stay combining (SSC) technique. 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 can be easily performed, because SIR value is simply measurable in real time, by using specific SIR estimators, in base stations (uplink), mobile stations (downlink) and digital wireless systems.

In communications systems analysis a few statistical models are used to describe fading in wireless environments. The most frequently used distributions are Nakagami-m, Rice, Rayleigh, α-μ, and Weibull. The Rician 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. In particular, for mobile satellite communications, the Rician distribution is used to accurately model the mobile satellite channel for single [4], clear-state [5] channel conditions. Then the Rician fading is applicable for modeling the fading channels in frequency domain [6]. Also the Rician K factor characterizes the land mobile satellite channel, during unshadowed periods [7]. The diversity reception over Rician fading channels was previously discussed in [8–11]. The system performances of dual selection combining (SC) over correlated Rician channels are in the presence of CCI analyzed in [11].

There are a number of papers concerning performance analysis of SSC receivers, for example [12–16] Very useful, novel, infinite series expressions are obtained for the output SIR probability density function (PDF) and cumulative distribution function (CDF) of a dual-branch switched-and-stay combining (SSC) diversity receiver, operating over correlated α-μ fading in the presence of cochannel interference (CCI) derived in [12]. 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). However the α-μ fading model cannot be reduced only to Ricean fading model, so the analysis from [12] cannot be used for the physical scenario when strong direct (LoS) signal component is present. The performance analysis of the SSC diversity receivers, operating over correlated Ricean fading satellite channels, can be found in [14], but without the consideration of CCI effects. Reference [16] 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 the author's knowledge, no analytical study of switch-and-stay combining involving assumed correlated Ricean fading for both desired signal and cochannel interference has been reported in the literature.

In this paper, an approach to the performance analysis of given SSC diversity receiver over correlated Rician fading channels, in the presence of correlated CCI, is presented. In order to study the effectiveness of any modulation scheme and the type of diversity used, it is required to evaluate the system's performance over the channel conditions. Infinite series expressions for PDF and CDF of the output SIR for SSC diversity are derived. Furthermore, important performance measures, such as outage probability (OP) and Average Symbol Error Probability (ASEP) for several modulation schemes such as noncoherent frequency-shift keying (NCFSK) and binary differentially phase-shift keying (BDPSK) are efficiently evaluated and shown graphically for different system parameters in order to point out the effects of fading severity and the level of correlation on the system performances.

2. System Model and Received Statistics

Let us consider diversity system applied on a small terminal with multiple antennas. Due to insufficient antennae spacing, desired signal envelopes R_1 and R_2 will experience correlative α-μ fading, with joint distribution [16]:

$$\begin{aligned} f_{R_1,R_2}(R_1,R_2) = {} & \frac{R_1R_2(K_d+1)^2}{\beta_d^2(1-r^2)} \\ & \times \exp\left(-\frac{(R_1^2+R_2^2)(K_d+1)+4K_d\beta_d(1-r)}{2\beta_d(1-r^2)}\right) \\ & \times \sum_{k=0}^{\infty}\xi_k I_k\left(\frac{R_1R_2r(K_d+1)}{\beta_d(1-r^2)}\right) \\ & \times I_k\left(\frac{R_1}{(1+r)}\sqrt{\frac{2K_d(K_d+1)}{\beta_d}}\right) \\ & \times I_k\left(\frac{R_2}{(1+r)}\sqrt{\frac{2K_d(K_d+1)}{\beta_d}}\right). \end{aligned} \tag{1}$$

Similarly, due to insufficient antennae spacing, envelopes of interference r_1 and r_2 are correlated with joint Ricean distribution:

$$\begin{aligned} f_{r_1,r_2}(r_1,r_2) = {} & \frac{r_1r_2(K_i+1)^2}{\beta_i^2(1-r^2)} \\ & \times \exp\left(-\frac{(r_1^2+r_2^2)(K_i+1)+4K_i\beta_i(1-r)}{2\beta_i(1-r^2)}\right) \\ & \times \sum_{l=0}^{\infty}\xi_l I_l\left(\frac{r_1r_2r(K_i+1)}{\beta_i(1-r^2)}\right) \\ & \times I_l\left(\frac{r_1}{(1+r)}\sqrt{\frac{2K_i(K_i+1)}{\beta_i}}\right) \\ & \times I_l\left(\frac{r_2}{(1+r)}\sqrt{\frac{2K_i(K_i+1)}{\beta_i}}\right). \end{aligned} \tag{2}$$

It is important to quote that β_d and β_i, defined as $\beta_d = \overline{R_1^2}/2 = \overline{R_2^2}/2$ and $\beta_i = \overline{r_1^2}/2 = \overline{r_2^2}/2$, denote the average powers of desired and interference signals, respectively. $I_k(x)$ and $I_l(x)$ are the modified Bessel function of the first kind of the kth and lth order. K_d, known as Ricean factor, defines ratio of signal power in dominant component of desired signal over the scattered power while K_i is, similarly, Ricean factor of the interference signal. ξ_k and ξ_l are defined as $\xi_k = 1$ for $(k = 0)$ and $\xi_k = 2$ for $(k > 0)$ and similarly $\xi_l = 1$ for $(l = 0)$ and $\xi_l = 2$ for $(l > 0)$. Finally, with r the correlation coefficients are denoted.

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 [17, 18]. The joint PDF of z_1 and z_2 can be expressed by [17]

$$\begin{aligned} & f_{z_1,z_2}(z_1,z_2) \\ & = \frac{1}{4\sqrt{z_1z_2}}\int_0^{\infty}\int_0^{\infty} p_{R_1,R_2}(r_1\sqrt{z_1},r_2\sqrt{z_2})\,p_{r_1,r_2}(r_1,r_2)\,r_1r_2\,dr_1dr_2. \end{aligned} \tag{3}$$

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

$$f_{z_1,z_2}(z_1,z_2) = \sum_{k=0}^{\infty}\sum_{l=0}^{\infty}\sum_{m=0}^{\infty}\sum_{n=0}^{\infty}\sum_{p=0}^{\infty}\sum_{q=0}^{\infty}\sum_{s=0}^{\infty}\sum_{w=0}^{\infty} G_1 \times \frac{z_1^{m+n+k}S^{2q+2l+s+w+2}}{(z_1(K_d+1)+S(K_i+1))^{m+n+k+q+s+l+2}} \times \frac{z_2^{m+p+k}}{(z_2(K_d+1)+S(K_i+1))^{m+p+k+q+w+l+2}} \quad (4)$$

with

$$G_1 = \xi_k\xi_l \frac{r^{2m+2q+k+l}(1-r^2)^{(n+p+k+s+w+l+2)}}{\Gamma(m+k+1)\Gamma(n+k+1)\Gamma(p+k+1)} \times \exp\left(-\frac{2(K_d+K_i)}{1+r}\right) \times \frac{K_d^{n+p+k}K_i^{s+w+l}(K_d+1)^{2m+2k+n+p+2}(K_i+1)^{2q+2l+s+w+2}}{(1+r)^{2(n+p+k+s+w+l)}m!n!p!q!s!w!} \times \frac{\Gamma(m+n+k+q+s+l+2)\Gamma(m+p+k+q+w+l+2)}{\Gamma(q+l+1)\Gamma(s+l+1)\Gamma(w+l+1)}. \quad (5)$$

The 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 both input branches. Following [19], the PDF of SIR is given by

$$f_{z_{\text{ssc}}}(z) = \begin{cases} v_{\text{ssc}}(z), & z \le z_\tau, \\ v_{\text{ssc}}(z) + f_z(z), & z > z_\tau, \end{cases} \quad (6)$$

where $v_{\text{ssc}}(z)$, according to [19], 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:

$$v_{\text{ssc}}(z) = \sum_{k=0}^{\infty}\sum_{l=0}^{\infty}\sum_{m=0}^{\infty}\sum_{n=0}^{\infty}\sum_{p=0}^{\infty}\sum_{q=0}^{\infty}\sum_{s=0}^{\infty}\sum_{w=0}^{\infty} G_2 G_3^{m+p+k+1} \times \frac{z^{m+n+k}S^{q+l+s+1}}{(z(K_d+1)+S(K_i+1))^{m+n+k+q+s+l+2}} \times {}_2F_1(m+p+k+1,-q-w-l;2+m+p+k;G_3) \quad (7)$$

with ${}_2F_1(u_1,u_2;u_3;x)$ being the Gaussian hypergeometric function [20], and

$$G_2 = \xi_k\xi_l \frac{r^{2m+2q+k+l}(1-r^2)^{(n+p+k+s+w+l+2)}}{\Gamma(m+k+1)\Gamma(n+k+1)\Gamma(p+k+1)} \times \exp\left(-\frac{2(K_d+K_i)}{1+r}\right) \times \frac{K_d^{n+p+k}K_i^{s+w+l}(K_d+1)^{m+k+n+1}(K_i+1)^{q+l+s+1}}{(1+r)^{2(n+p+k+s+w+l)}m!n!p!q!s!w!(m+p+k+1)} \times \frac{\Gamma(m+n+k+q+s+l+2)\Gamma(m+p+k+q+w+l+2)}{\Gamma(q+l+1)\Gamma(s+l+1)\Gamma(w+l+1)},$$

$$G_3 = \frac{(K_d+1)z_\tau}{(K_d+1)z_\tau+(K_i+1)S}. \quad (8)$$

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

$$f_z(z) = \sum_{k=0}^{\infty}\sum_{l=0}^{\infty} \frac{(K_d+1)^{k+1}(K_i+1)^{l+1}z^kS^{l+1}\Gamma(k+l+2)}{(z(K_d+1)+S(K_i+1))^{k+l+2}\Gamma(k+1)\Gamma(l+1)k!l!} \times \exp\left(-\frac{2(K_d+K_i)}{1+r}\right). \quad (9)$$

Similarly, the CDF of the SSC output SIR, that is, the $F_{z_{\text{ssc}}}(z)$ is given by [19]

$$F_{z_{\text{ssc}}}(z) = \begin{cases} F_{z_1,z_2}(z,z_\tau), & z \le z_\tau, \\ F_z(z) - F_z(z_\tau) + F_{z_1,z_2}(z,z_\tau), & z > z_\tau. \end{cases} \quad (10)$$

By substituting (6) in

$$F_{z_1,z_2}(z,z_\tau) = \int_0^z\int_0^{z_\tau} f_{z_1,z_2}(z_1,z_2)dz_1dz_2 \quad (11)$$

$F_{z_1,z_2}(z,z_\tau)$ can be expressed as the following infinite series:

$$F_{z_1,z_2}(z,z_\tau) = \sum_{k=0}^{\infty}\sum_{l=0}^{\infty}\sum_{m=0}^{\infty}\sum_{n=0}^{\infty}\sum_{p=0}^{\infty}\sum_{q=0}^{\infty}\sum_{s=0}^{\infty}\sum_{w=0}^{\infty} G_4 G_5^{m+n+k+1} G_3^{m+p+k+1} \times {}_2F_1(m+n+k+1,-q-s-l;2+m+n+k+1;G_5) \times {}_2F_1(m+p+k+1,-q-w-l;2+m+p+k;G_3). \quad (12)$$

In the same manner by substituting (9) in

$$F_z(z) = \int_0^z f_z(z)dz \quad (13)$$

Table 1: Number of terms of (10) required for the 5th significant digit accuracy ($S/\gamma_{th} = 0$ dB).

r	$K_d = 1$	$K_d = 2$	$K_d = 3$
	$K_i = 1$	$K_i = 1$	$K_i = 1$
0.2	5	6	8
0.5	9	12	14

the expressions for $F_z(z)$ and $F_z(z_\tau)$ can be derived as

$$F_z(z) = \sum_{k=0}^{\infty}\sum_{l=0}^{\infty}\frac{\Gamma(l+k+2)}{\Gamma(k+1)\Gamma(l+1)l!k!}\exp\left(-\frac{2(K_d+K_i)}{1+r}\right)\frac{G_5^{k+1}}{k+1} \times {}_2F_1(k+1,-l;2+k;G_5),$$

$$F_z(z_\tau) = \sum_{k=0}^{\infty}\sum_{l=0}^{\infty}\frac{\Gamma(l+k+2)}{\Gamma(k+1)\Gamma(l+1)l!k!}\exp\left(-\frac{2(K_d+K_i)}{1+r}\right)\frac{G_3^{k+1}}{k+1} \times {}_2F_1(k+1,-l;2+k;G_3) \tag{14}$$

with

$$G_4 = \xi_k\xi_l\frac{r^{2m+2q+k+l}(1-r^2)^{(n+p+k+s+w+l+2)}}{\Gamma(m+k+1)\Gamma(n+k+1)\Gamma(p+k+1)} \times \exp\left(-\frac{2(K_d+K_i)}{1+r}\right) \times \frac{K_d^{n+p+k}K_i^{s+w+l}}{(1+r)^{2(n+p+k+s+w+l)}m!n!p!q!s!w!(m+p+k+1)(m+n+k+1)} \times \frac{\Gamma(m+n+k+q+s+l+2)\Gamma(m+p+k+q+w+l+2)}{\Gamma(q+l+1)\Gamma(s+l+1)\Gamma(w+l+1)},$$

$$G_5 = \frac{(K_d+1)z}{(K_d+1)z+(K_i+1)S}. \tag{15}$$

The nested infinite sums in (7) and (12) converge for any defined value of the parameters r, K_d, and K_i. Numbers of terms that need to be summed in (12) to achieve accuracy at the 5th significant digit are presented in Table 1. As it is shown in this table, the number of the terms needed to be summed to achieve a desired accuracy depends strongly on the correlation coefficient r. Here, the number of the terms increases as correlation coefficient increases.

3. Outage Probability

The outage probability (OP) is standard performance criterion of communication systems operating over fading channels, defined as probability that the instantaneous SIR of the system falls below a specified threshold value. The protection ratio depends on modulation technique and expected QoS. OP 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 (10), (12), and (14) the P_{out} 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 parameters r, K_d, and K_i. The normalized outage threshold (dB) is defined as being the average SIRs at the input branch of the balanced dual-branch switched-and-stay combiner, normalized by specified threshold value z^*.

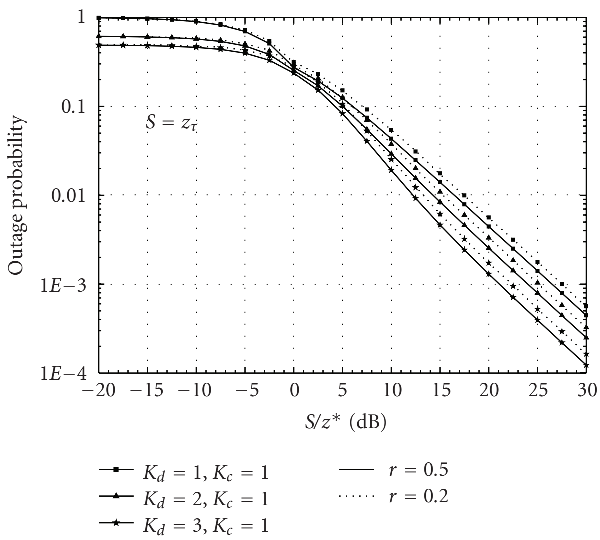

Figure 1: Outage probability (P_{out}) versus normalized outage threshold for the balanced dual-branch SSC diversity receiver and different values of parameters r and K_d.

It can be observed from that figure that OP deteriorates with decrease of the Rice factor K_d. Also, presented results show branch correlation influence on the OP. Namely, when correlation coefficients r increas, OP increases.

4. Average Symbol Error Probability

The average symbol error probability of the output of SSC ($\overline{P_e}$) can be evaluated by averaging the conditional symbol error probability for a given SIR, over the PDF of z_{ssc}, that is, $f_{z_{\text{ssc}}}(z)$ [19]:

$$\overline{P_e} = \int_0^\infty P_e(z)f_{z_{\text{ssc}}}(z)dz, \tag{16}$$

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 $\lambda = 1/2$ for NCFSK and $\lambda = 1$ for BDFSK [17]. Hence, substituting (7) into (16) gives the following ASEP expression for the considered dual-branch SSC receiver:

$$\overline{P_e} = \int_0^\infty P_e(z)v_{z_{\text{ssc}}}(z)dz + \int_{z_\tau}^\infty P_e(z)f_z(z)dz. \tag{17}$$

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 BDPSK. Applying (17) on BDPSK and NCFSK modulation schemes, the ASEP performance results have been obtained,

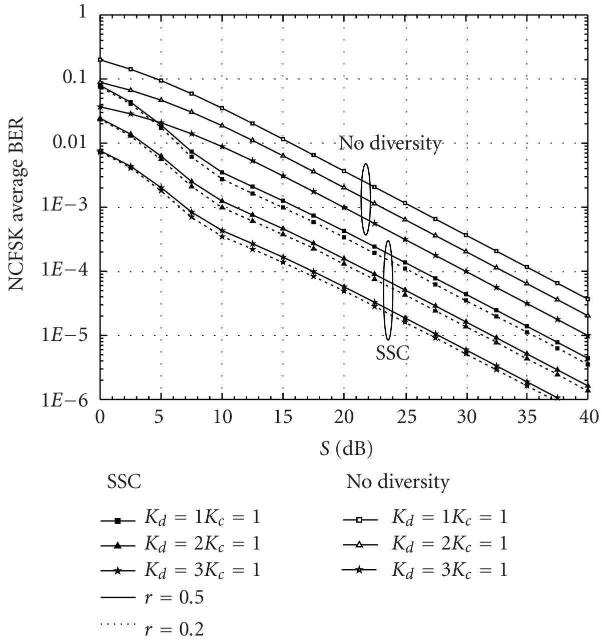

Figure 2: Average bit error probability (ABER) versus average SIRs at the input branches of the balanced dual-branch switched-and-stay combiner, S, for NCFSK modulation scheme and several values of parameters r and K_d.

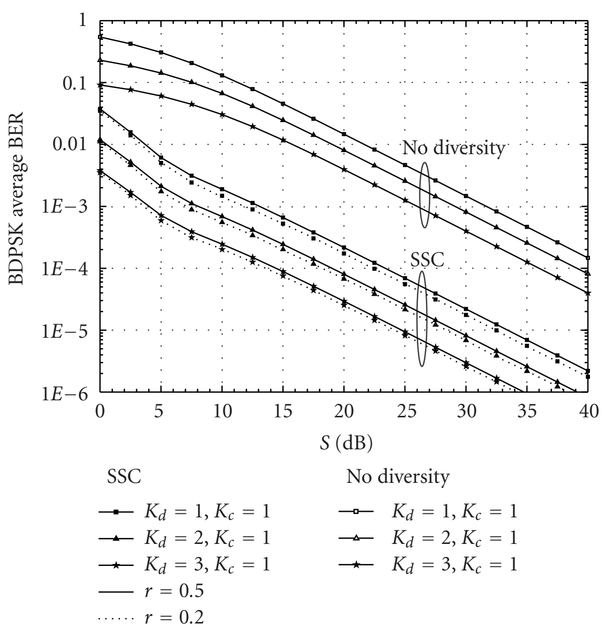

Figure 3: Average bit error probability (ABER) versus average SIRs at the input branches of the balanced dual-branch switched-and-stay combiner, S, for BDPSK modulation scheme and several values of parameters r and K_d.

for several values of r, K_d, and K_i, as a function of the average SIRs at the input branches of the balanced dual-branch switched-and-stay combiner, that is, S. These results are plotted in Figures 2 and 3. First the comparation was made between the no-diversity case and the SSC receiving technique. It is obvious that results are much better when diversity is applied. Then we can observe from the figures that while the average SIR at the input branches and values of Rice factor of desired signal, K_d, are increasing, the ASEP increases at the same time. Also, these figures show better error performances for greater distance between diversity branches, that is, for smaller values of the correlation coefficient r. The comparison of Figures 2 and 3 shows better performance of BDPSK modulation scheme versus NCFSK modulation scheme.

5. Conclusion

The system performances of dual SSC system over correlated Rician fading channels in the presence of correlated CCI are analyzed. Crucial statistics metrics for the SSC output SIR are given in the infinite-series form, that is, PDF and CDF. Capitalizing on this, outage probability and average symbol error probability have been obtained and graphically presented, describing their dependence on the correlation coefficient and the fading severity. The main contribution of this paper is analytical study of dual SSC diversity system assuming, for the first time, correlated Rician fading for both desired signal and interference.

Appendix

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

$$
\begin{aligned}
f_{z_1,z_2}(z_1,z_2) = {} & \sum_{k=0}^{\infty}\sum_{l=0}^{\infty}\sum_{m=0}^{\infty}\sum_{n=0}^{\infty}\sum_{p=0}^{\infty}\sum_{q=0}^{\infty}\sum_{s=0}^{\infty}\sum_{w=0}^{\infty} C_1 \\
& \times \int_0^{\infty} r_1^{2m+2n+2k+2p+2s+2l+3} \\
& \times \exp\left(-\frac{r_1^2 z_1 (K_d+1)}{2\beta_d(1-r^2)} - \frac{r_1^2 (K_i+1)}{2\beta_i(1-r^2)}\right) dr_1 \\
& \times \int_0^{\infty} r_2^{2m+2p+2k+2q+2w+2l+3} \\
& \times \exp\left(-\frac{r_2^2 z_2 (K_d+1)}{2\beta_d(1-r^2)} - \frac{r_2^2 (K_i+1)}{2\beta_i(1-r^2)}\right) dr_2.
\end{aligned}
\tag{A.1}
$$

Let $S = \beta_d/\beta_i$ be the average SIRs at the input branches of the dual-branch SC. Then, the following integrals from pervious expression can be presented in the form of

$$
\begin{aligned}
I_1 = {} & \int_0^{\infty} r_1{}^{2m+2n+2k+2p+2s+2l+3} \\
& \times \exp\left(-r_1^2 \frac{z_1(K_d+1)+S(K_i+1)}{2\beta_d(1-r^2)}\right) dr_1, \\
I_2 = {} & \int_0^{\infty} r_2{}^{2m+2p+2k+2q+2w+2l+3} \\
& \times \exp\left(-r_2^2 \frac{z_1(K_d+1)+S(K_i+1)}{2\beta_d(1-r^2)}\right) dr_2.
\end{aligned}
\tag{A.2}
$$

Now, they can easily be solved using variable substitutions:

$$u = r_1^2\left(\frac{z_1(K_d+1)+S(K_i+1)}{2\beta_d(1-r^2)}\right),$$
$$v = r_2^2\left(\frac{z_1(K_d+1)+S(K_i+1)}{2\beta_d(1-r^2)}\right), \tag{A.3}$$

and well-known definition of Gamma function: $\Gamma(a) = \int_0^\infty t^{a-1}\exp(-t)dt$. Finally, (11) can be written as

$$\begin{aligned} f_{z_1,z_2}(z_1,z_2) = &\sum_{k=0}^{\infty}\sum_{l=0}^{\infty}\sum_{m=0}^{\infty}\sum_{n=0}^{\infty}\sum_{p=0}^{\infty}\sum_{q=0}^{\infty}\sum_{s=0}^{\infty}\sum_{w=0}^{\infty} G_1 \\ &\times \frac{z_1^{m+n+k}S^{2q+2l+s+w+2}}{(z_1(K_d+1)+S(K_i+1))^{m+n+k+q+s+l+2}} \\ &\times \frac{z_2^{m+p+k}}{(z_2(K_d+1)+S(K_i+1))^{m+p+k+q+w+l+2}} \end{aligned} \tag{A.4}$$

with

$$\begin{aligned} G_1 &= \xi_k\xi_l \frac{r^{2m+2q+k+l}(1-r^2)^{(n+p+k+s+w+l+2)}}{\Gamma(m+k+1)\Gamma(n+k+1)\Gamma(p+k+1)} \\ &\times \exp\left(-\frac{2(K_d+K_i)}{1+r}\right) \\ &\times \frac{K_d^{n+p+k}K_i^{s+w+l}(K_d+1)^{2m+2k+n+p+2}(K_i+1)^{2q+2l+s+w+2}}{(1+r)^{2(n+p+k+s+w+l)}m!n!p!q!s!w!} \\ &\times \frac{\Gamma(m+n+k+q+s+l+2)\Gamma(m+p+k+q+w+l+2)}{\Gamma(q+l+1)\Gamma(s+l+1)\Gamma(w+l+1)}. \end{aligned} \tag{A.5}$$

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