Asymptotic Error Rate Analysis of Multi-Branch EGC and SC on Equally Correlated Rician Channels

Shuo Liu  
Dept. of ECE  
Univ. of Alberta, Edmonton, AB  
Canada T6G 2V4  
Email: sliu2@ualberta.ca

Julian Cheng  
School of Engineering  
UBC Okanagan, Kelowna, BC  
Canada V1V 1V7  
Email: julian.cheng@ubc.ca

Norman C. Beaulieu  
Dept. of ECE  
Univ. of Alberta, Edmonton, AB  
Canada T6G 2V4  
Email: beaulieu@ece.ualberta.ca

Abstract—Asymptotic bit-error rate expressions are derived for binary phase shift keying with multi-branch equal gain combining (EGC) and selection combining (SC) on equally correlated Rician channels. Numerical results indicate that these analytical error rate expressions can provide accurate error rate estimation at large signal-to-noise ratio (SNR). Furthermore, these analytical results can be used to reveal some important insights into the performance characteristics of EGC and SC operating on equally correlated Rician fading channels.

I. INTRODUCTION

Multi-branch diversity combining is a useful technique to mitigate the adverse effect of multipath fading in wireless communication systems. For example, equal gain combining (EGC) and selection combining (SC) are two important diversity combining techniques that provide substantial error rate performance improvement at relatively low implementation cost. Much work has been done on the error rate analysis of multi-branch diversity combining on independent fading channels [1]. However, fewer results for the correlated case have been published because error rate performance analysis is usually challenging for correlated diversity branches, particularly for EGC and SC. Yet, knowledge of the properties of diversity combining on correlated fading channels is important because, for practical reasons, the receiver diversity branches are often correlated due to insufficient antenna spacing.

Analytical approaches have been attempted for diversity combining on correlated channels. Unfortunately, with the exception of dual-branch SC [2], closed-form error rate solutions are not known for multi-branch EGC and SC on correlated fading channels. The closed-form solutions are desired because they can be used to compute the exact error rates efficiently. Furthermore, additional insight can often be gleaned from the closed-form solutions. The equally correlated Rician fading case has been analyzed in [3] by Chen and Tellambura. However, their solution consists of multiple integrations. The complex analytical solution can estimate the exact error rate, but it may not be practical for large signal-to-noise ratio (SNR) values because the accuracy of the error rate estimation is limited by the error associated with the chosen numerical integration algorithm. Further, no analytical insight can be obtained from the complex error rate expressions. It is therefore desirable to apply asymptotic techniques to analyze the error rate on fading channels in large SNR regions. Unlike the exact analysis, the asymptotic approach can usually lead to a simple closed-form error rate solution, which can be used to predict error rates at large SNRs accurately and to reveal the influence of different parameters on the system performance.

In the study of diversity combining on correlated channels, some counter-intuitive behavior has been reported. In [4], with the assumption of cophasing and fixed correlation, Chang and McLane found that a noncoherent frequency shift keying diversity receiver can achieve better error rate performance on a correlated Rayleigh fading channel than on a correlated Rician fading channel. In addition, in [5], Ma et al. observed that the bit error rate (BER) can be improved by increasing the fading correlation coefficient on the Rician fading channel. These counter-intuitive observations, while correct, contradict conventional expectation.

In a recent work, simple asymptotic closed-form error rate expressions have been derived for dual-branch EGC and SC on correlated Rician channels [6], [7]. Previous counter-intuitive observations have been unified and analytically explained. In this work, we extend the work in [6], [7] to N-branch EGC and SC on equally correlated Rician channels. Using our analytical results, we investigate the transmission characteristics of EGC and SC on equally correlated Rician channels.

II. SYSTEM MODEL

We consider a slow frequency-nonselective Rician fading channel model. After coherent demodulation, matched filtering, and sampling, the output for the $i$th data symbol, $d(i)$, can be expressed as [6], [7]

$$\tilde{r}(i) = \tilde{a}(i)d(i) + \tilde{n}(i)$$

(1)

where $\tilde{r}(i) = [r_1(i), \ldots, r_N(i)]^T$, and where $r_1(i), \ldots, r_N(i)$ are the matched-filter outputs from the $N$ diversity branches, and $[\cdot]^T$ denotes the matrix transpose. In (1), $\tilde{n}(i) = [n_1(i), \ldots, n_N(i)]^T$, where $n_1(i), \ldots, n_N(i)$ are independent symmetric complex Gaussian random variables (RVs) with zero mean and unit variance. If we denote the complex fading coefficients on the $k$th branch by $c_k(i) = a_k(i)e^{j\theta_k(i)}$, $k = 1, \ldots, N$, then $a_k(i)$ and $\theta_k(i)$, are the amplitude and phase of $c_k(i)$, respectively. In the remainder of this paper, we will omit the index $i$ for simplicity.
For the Rician fading channels, we have

\[ c_k = c_{Lk} + c_{Sk}, \quad k = 1, \ldots, N \]  

(2)

where \( c_{Lk} \) denotes the line-of-sight (LOS) component of \( c_k \), and \( c_{Sk} \) denotes the scattering component of \( c_k \) on the \( k \)th branch. We write \( c_{Lk} = |c_{Lk}|e^{j\phi_{Lk}} = a_{Lk}e^{j\theta_{Lk}} \) where \( a_{Lk} \) and \( \theta_{Lk} \) are the amplitude and phase of the LOS component on the \( k \)th branch, respectively. For a correlated Rician fading channel, the probability density function (PDF) for \( \vec{c} \) can be expressed as [8]

\[ f(\vec{c}) = \frac{1}{|R|} \exp\{-(\vec{c} - \vec{c}_L)^H R^{-1}(\vec{c} - \vec{c}_L)\} \]  

(3)

where \( \vec{c} = [c_1, \ldots, c_N]^T \) and \( \vec{c}_L = [c_{L1}, \ldots, c_{LN}]^T \). The symbol \(|R|\) is the determinant of the matrix \( R \) and \((\cdot)^H \) denotes the Hermitian operator. In (3), \( R \) is the correlation matrix of the scattering components for \( N \) branches, i.e.

\[ R = E\{(\vec{c}_S)(\vec{c}_S)^H\} = E\{(\vec{c} - \vec{c}_L)(\vec{c} - \vec{c}_L)^H\} \]  

(4)

where \( \vec{c}_S = [c_{S1}, \ldots, c_{SN}]^T \), and \( c_{S1}, \ldots, c_{SN} \) are the \( N \) scattering components on the \( N \) branches, and where \( E\{\cdot\} \) denotes expectation. We have the correlation matrix

\[ R = \sigma_S^2 \begin{pmatrix} 1 & \rho & \cdots & \rho \\ \rho & 1 & \cdots & \rho \\ \vdots & \vdots & \ddots & \vdots \\ \rho & \cdots & \rho & 1 \end{pmatrix} \]  

(5a)

and its inverse

\[ R^{-1} = m \begin{pmatrix} n & \rho & \cdots & \rho \\ \rho & n & \cdots & \rho \\ \vdots & \vdots & \ddots & \vdots \\ \rho & \cdots & \rho & n \end{pmatrix} \]  

(5b)

where \( m = \sigma_S^2 \frac{1}{(1 + \rho)(1 - \rho)(N - 1)} \) and \( n = \rho - 1 - (N - 1)\rho \). The determinant of \( R \) is given by

\[ |R| = \sigma_S^2 N \left[ 1 + (N - 1)\rho \right] (1 - \rho)^{N - 1} \]  

(5c)

where \( \sigma_S^2 = E\{|c_{S1}|^2\} = \cdots = E\{|c_{SN}|^2\} \) is the variance of the scattering components, and \( \rho = E\{c_{Sk}c_{Sk}^H(c_{Sk})\}/\sigma_S^2 \) is the complex correlation coefficient between any two branches. In this paper, for analytical convenience we do not consider the case when \( \rho = 1 \).

In our work, we further assume that the amplitudes of the LOS components on the \( N \) branches are equal, i.e., \( |c_{L1}| = \cdots = |c_{LN}| = a_L \). Therefore, the Rice factor is given by \( K = a_L^2/\sigma_S^2 \) and the SNR in each branch is given by \( \Omega = a_L^2 + \sigma_S^2 = (K + 1)\sigma_S^2 \).

In the case when \( \sigma_S^2 \) is very small, the pdf of \( |\vec{c}_S| \) is approximated as

\[ f(\vec{c}_S) \approx \frac{1}{|\vec{c}_S|} \exp\left\{-\frac{|\vec{c}_S|^2}{\sigma_S^2}\right\} \]  

(6)

where \( a_L = \frac{1}{\sqrt{N}} \sum_{k=1}^{N} a_k \) and \( n_L = \frac{1}{N} \sum_{k=1}^{N} n_k \). The instantaneous output SNR is \( \Omega_L = (\sqrt{N}a_L)^2/E\{n_L^2n_L\} = a_L^2 \).

We require the Laplace transform of \( a_L \) to derive the asymptotic average error rate. The Laplace transform can be defined as

\[ \mathcal{L}_{a_L}(s) = E_{a_L}\{e^{-sa_L}\} \]  

(7)

To evaluate (7), we must derive the joint PDF for \( a_1, \theta_1, \ldots, a_n, \theta_n \). From (3), one can write the joint PDF as

\[ f(a_1, \theta_1, \ldots, a_N, \theta_N) = \frac{H_0}{\pi N \Omega^N} \prod_{k=1}^{N} H_k(a_k, \theta_k, \ldots, a_N, \theta_N) \prod_{i=1}^{N} a_i \]  

(8)

where

\[ H_0 = \frac{\Omega^N}{|R|} \exp\left\{-xNa_L^2 - 2ga_L^2 \sum_{k=1}^{N-1} \sum_{j=k+1}^{N} \cos(\theta_Lk - \theta_Lj)\right\} \]  

(9a)

\[ H_k(a_k, \theta_k, \ldots, a_N, \theta_N) = \exp\{-x[a_k^2 - 2a_k a_L \cos(\theta_k - \theta_{Lk})]\} \]  

\[ + 2ya_k a_L \sum_{j=1, j \neq k}^{N} \cos(\theta_k - \theta_{Lj}) \]  

\[ - 2ya_k \sum_{j=k+1}^{N} a_j \cos(\theta_k - \theta_j) \]  

\[ k = 1, 2, \ldots, N \]  

(9b)

where \( x = \frac{\rho - 1 - (N - 1)\rho}{\sigma_S^2 (\rho - 1)(1 + (N - 1)\rho)} \) and \( y = \frac{\rho}{\sigma_S^2 (\rho - 1)(1 + (N - 1)\rho)} \).

For notational simplicity we use \( H_k \) to denote \( H_k(a_k, \theta_k, \ldots, a_N, \theta_N) \) in the sequel. Now the Laplace transform of \( a_L \) becomes

\[ \mathcal{L}_{a_L}(s) = \int_{0}^{\infty} \cdots \int_{0}^{\infty} \exp\left\{-s(a_1 + \cdots + a_N)\right\} da_1 \cdots da_n \]  

(10a)

\[ \times \int_{0}^{2\pi} \cdots \int_{0}^{2\pi} f(a_1, \theta_1, \ldots, a_N, \theta_N) da_1 \cdots, d\theta_N \]  

(10b)

\[ = \frac{H_0}{\pi N \Omega^N} \int_{0}^{2\pi} \cdots \int_{0}^{2\pi} A d\theta_1 \cdots, d\theta_N \]  

(10c)
where
\[
A = \int_0^\infty H_N e^{-\frac{s}{\sqrt{N}}} \int_0^\infty H_{N-1} e^{-\frac{s}{\sqrt{N}}} \cdots \int_0^\infty H_2 e^{-\frac{s}{\sqrt{N}}} \left( \int_0^\infty H_1 e^{-\frac{s}{\sqrt{N}}} a_1 da_1 \right) a_2 da_2 \cdots a_{N-1} da_{N-1} a_N da_N. 
\]

(10b)

Taking advantage of the quadratic form in the exponent of \( H_1 \) we can show
\[
\int_0^\infty H_1 \exp \left( -\frac{s a_1}{\sqrt{N}} \right) a_1 da_1 = \frac{1}{\beta^2} + o \left( \frac{1}{\beta^2} \right) \quad (11a)
\]
where
\[
\beta = \frac{s}{\sqrt{N}} - 2x a_L \cos(\theta_1 - \theta_{L1}) - 2y a_L \sum_{j=2}^N \cos(\theta_1 - \theta_{Lj}) + 2y \sum_{j=2}^N a_j \cos(\theta_1 - \theta_j)
\]
and the definition of \( o(\cdot) \) follows [9].

By an asymptotic approximation \( \beta = \frac{s}{\sqrt{N}} \) as \( s \to \infty \), and eqn. (11a) can be approximated as
\[
\int_0^\infty H_1 \exp \left( -\frac{s a_1}{\sqrt{N}} \right) a_1 da_1 = \frac{N}{s^2} + o \left( \frac{1}{s^2} \right). \quad (12)
\]

Substituting (12) into (10b) and integrating over \( a_2, \ldots, a_N \) in a similar way, one obtains
\[
A = \frac{N^N}{s^{2N}} + o \left( \frac{1}{s^{2N}} \right). \quad (13)
\]

Substituting (13) into (10a) and integrating over \( \theta_1, \theta_2, \ldots, \theta_N \), we finally get
\[
L_{a_E}(s) = \frac{H_0}{\pi N \Omega^N} \left[ \frac{N^N}{s^{2N}} + o \left( \frac{1}{s^{2N}} \right) \right] (2\pi)^N
= \frac{2p_t \Gamma(2t+2)}{\Omega^t+1|\Omega|^2t+1} + o \left( \frac{1}{\Omega^t} \right) \quad (14)
\]
where \( t = N - 1 \) and \( p_t = \frac{(2N)^N H_0}{2t(2N)^\Omega} \). Now we can apply proposition 2 in [7]
\[
P_{c} = \frac{(2N)^N H_0}{2t(2N)^\Omega} \int_0^\infty P_c(\gamma) \gamma^{N-1} d\gamma + o \left( \frac{1}{\Omega^N} \right)
= \frac{[2N(K+1)]^N \exp\{B\}}{2[1+(N-1)\rho(1-\rho)]^{N-1}(2N)^\Omega N} + o \left( \frac{1}{\Omega^N} \right) \quad (15)
\]
where \( P_c(\gamma) \) is the conditional symbol error rate of a coherent modulation.

When coherent binary phase shift keying (BPSK) signaling is considered, we have
\[
P_c(\gamma) = Q(\sqrt{2\gamma}). \quad (16)
\]

Using a very similar procedure to that applied in the development of proposition 1 in [9], and an integral identity [9, eq. (4)]
\[
\int_0^\infty Q(\sqrt{2\gamma}) \gamma^{N-1} d\gamma = \frac{(2N-1)!!}{N \cdot 2^{N+1}} \quad (17)
\]
one obtains
\[
P_c = \frac{[N(K + 1)]^N \exp\{B\}}{4N[1 + (N-1)\rho(1-\rho)]^{N-1}(2N-2)!! \Omega^N} + o \left( \frac{1}{\Omega^N} \right) \quad (18)
\]
where \( !! \) denotes the double factorial [10].

In the special case of dual-branch reception \( (N = 2) \), we have
\[
P_{c} = \frac{(K + 1)^2}{4\Omega^2(1-\rho^2)} \exp \left\{ - \frac{2K}{1-\rho^2} [1 - \rho \cos(\theta_{L1} - \theta_{L2})] \right\}
+ o \left( \frac{1}{\Omega^2} \right) \quad (19)
\]
which agrees with the result in [6], [7].

IV. ASYMPTOTIC ERROR RATE ANALYSIS FOR SC

The received signal for SC can be expressed as \( r_S = \max(r_1, r_2, \ldots, r_N) \) [1, p. 169]. Thus, the instantaneous SC output SNR is given by
\[
\Omega_S = \max(a_1^2, a_2^2, \ldots, a_N^2). \quad (20)
\]

To find a PDF expression for \( a_S = \sqrt{\Omega_S} = \max(a_1, a_2, \ldots, a_N) \), we start with the cumulative distribution function (CDF) of \( a_S \)
\[
F(a_S) = \int_{0}^{a_S} \cdots \int_{0}^{a_S} \int_{0}^{2\pi} \cdots \int_{0}^{2\pi} f(a_1, \theta_1, \ldots, a_N, \theta_N) \ d\theta_1 \cdots d\theta_N da_1 \cdots da_N. \quad (21)
\]

From [11] we know there is a coefficient \( \frac{N^N}{(k-N)!(N-k)!} \) for the \( k \)th order statistic in \( N \) RVs when differentiating the CDF. For
Applying the approximation to (22), we obtain
\[ f(a_s) = dF(a_s)_{\mathrm{ds}} = N \int_0^{a_s} \cdots \int_0^{a_s} f(a_1, \theta_1, \ldots, a_N, \theta_N) \]
\[ = N \int_0^{a_s} \int_0^{a_s} \cdots \int_0^{a_s} f(a_1, \theta_1, \ldots, a_N, \theta_N) \]
\[ = N \cdot a_s H_0 \frac{a^2_s}{\pi N \Omega^N} \int_0^{a_s} \cdots \int_0^{a_s} H_N(a_s, \theta_N) \]
\[ \times \int_0^{a_s} H_{N-1}(a_{N-1}, \theta_{N-1}, a_s, \theta_N) \cdots \]
\[ \times \int_0^{a_s} H_2(a_2, \theta_2, \ldots, a_s, \theta_N) \]
\[ \times \left( \int_0^{a_s} H_1(a_1, \theta_1, \ldots, a_s, \theta_N) a_1 \mathrm{d}a_1 \right) \]
\[ \times a_2 \mathrm{d}a_2 \ldots a_{N-1} \mathrm{d}a_{N-1} \mathrm{d}\theta_1 \ldots \mathrm{d}\theta_N. \] (22)

Applying the approximation
\[ \int_0^{a_s} H_1(a_1, \theta_1, \ldots, a_s, \theta_N) a_1 \mathrm{d}a_1 = \frac{a^2_s}{2} + o(a^2_s), \text{for } a_s \to 0 \] (23)
to (22), we obtain
\[ f(a_s) = N \cdot a_s H_0 \frac{a^2_s}{\pi N \Omega^N} \left[ \frac{a^2_s}{2} + o(a^2_s) \right] \]
\[ \times \int_0^{a_s} \cdots \int_0^{a_s} H_N(a_s, \theta_N) \]
\[ \times \int_0^{a_s} H_{N-1}(a_{N-1}, \theta_{N-1}, a_s, \theta_N) \cdots \]
\[ \times \left( \int_0^{a_s} H_2(a_2, \theta_2, \ldots, a_s, \theta_N) a_2 \mathrm{d}a_2 \right) \]
\[ \ldots a_{N-1} \mathrm{d}a_{N-1} \mathrm{d}\theta_1 \ldots \mathrm{d}\theta_N. \] (24)

We can repeat the process for \( a_2, a_3, \ldots, a_{N-1} \) and obtain
\[ f(a_s) = N \cdot a_s H_0 \frac{a^2_s}{\pi N \Omega^N} \left[ \frac{a^2_s}{2} + o(a^2_s) \right] \]
\[ \times \int_0^{a_s} \cdots \int_0^{a_s} \left[ 1 + o(1) \right] \mathrm{d}\theta_1 \ldots \mathrm{d}\theta_N \] (25)
\[ = 2^{p_t} \frac{(a^2_s)^{2t+1}}{\Omega^{2t+1}} + o \left( \frac{(a^2_s)^{2t+1}}{\Omega^{2t+1}} \right) \]

where \( t = N - 1 \) and \( p_t = N \cdot H_0 \). Now we apply proposition 1 in [7]

\[ P_e^S = N \cdot H_0 \frac{a^2_s}{\pi N \Omega^N} \int_0^{\infty} P_e(\gamma) \gamma^{N-1} \mathrm{d}\gamma + o \left( \frac{1}{\Omega^N} \right) \]
\[ = N(K+1)^N \exp \{ B \} \int_0^{\infty} P_e(\gamma) \gamma^{N-1} \mathrm{d}\gamma + o \left( \frac{1}{\Omega^N} \right). \] (26)

When coherent BPSK is considered, one obtains
\[ P_e^S = \frac{(2N - 1)!(K+1)^N \exp \{ B \}}{2^{N+1} [1 + (N - 1)\rho] [1 - \rho(N-1)\Omega^N] + o \left( \frac{1}{\Omega^N} \right)}. \] (27)

In the dual-branch case (\( N = 2 \)), we have
\[ P_e^S = \frac{3(K+1)^2}{8\Omega^2 (1 - \rho^2)} \exp \left\{ -2K \rho [1 - \rho \cos (\theta_1 - \theta_2)] \right\} \]
\[ + o \left( \frac{1}{\Omega^2} \right). \] (28)

which also agrees with the result in [6], [7].

V. NUMERICAL RESULTS AND DISCUSSION

We simulate the EGC and SC models for the 3-branch and 4-branch cases as shown in Fig. 1 and Fig. 2, respectively, for different correlation parameters. Observe that the simulated curves agree well with our asymptotic BER curve at high SNRs. Our numerical results suggest that the asymptotic error rate estimations are accurate when the SNR is larger than 25 dB.

A. Asymptotic Performance Comparison Between EGC and SC

Observe from (15) and (26) that the two analytical expressions both contain multiplication of a constant with the common term \( \frac{H_0}{\Omega^N} \int_0^{\infty} P_e(\gamma) \gamma^{N-1} \mathrm{d}\gamma \). Some insights can be obtained from this observation. First, according to the expression of the average error rate in [9], the factor \( \frac{1}{\Omega^N} \) indicates that the diversity order of \( N \)-branch diversity combining is \( N \), as expected. Second, we notice that the factor \( H_0 \) is determined by the parameters \( K \), \( \rho \) and the phase differences, and these parameters will contribute to the coding gain. Finally, taking the ratio of the asymptotic BERs for EGC and SC, we obtain \( \frac{P_e^S}{P_e^S \text{ asympt}} = (2N)^{N-1} \), which can be shown to be less than unity. This indicates that the error rate performance of EGC is better than SC, as expected. This fact has been confirmed by computer simulation.

B. Impact of \( K \) on Error Rate Performance

We observe that the asymptotic error rate is only determined by \( H_0 \) for a fixed \( N \) and SNR, so it is sufficient to study the impact of \( K \) on \( H_0 \). We take the partial derivative of \( H_0 \) with respect to \( K \) and obtain
\[ \frac{\partial H_0}{\partial K} = \frac{N \cdot H_0}{K + 1} D_K \] (29a)
where
\[ D_K = 1 - \frac{K + 1}{(1 - \rho)[1 + (N - 1)\rho]} \]
\[ \times \left\{ \left( \frac{(N - 2)\rho + 1}{N} \right) \sum_{k=1}^{N} \sum_{j=k+1}^{N} \cos (\theta_{Lk} - \theta_{Lj}) \right\}. \] (29b)
We observe that the sign of $\frac{\partial H_0}{\partial K}$ is determined by the sign of $D_K$. If $D_K > 0$, an increase in the Rice factor $K$ will increase the error rate. If $D_K < 0$, an increase of Rice factor $K$ will lower the error rate. We take 3-branch diversity combining with $\rho = 0.9$ and all phase differences of 0 rad as an example. Set $D_K = 1 - \frac{k+1}{k+1} \geq 0$, then $K < 1.8$. This means that the error rate is higher with a higher $K$ if $K \in [0, 1.8)$. Thus, the BER performance of the case with $K = 2$ may be poorer than the one with $K = 0$ (see our simulation). This also explains why a Rayleigh fading channel has better performance than a Rician fading channel as observed in [4].

C. Impact of $\rho$ on Error Rate Performance

To study the impact of $\rho$ on error rate performance we can also differentiate $H_0$ with respect to $\rho$ and get

$$
\frac{\partial H_0}{\partial \rho} = \frac{N H_0}{(1-\rho)(1 + (N-1)\rho)} D_\rho \tag{30}
$$

where

$$
D_\rho = (N-1)\rho - K[(N-1)(2\rho - 1) + 1]
- \frac{(1-\rho)(1 + (N-1)\rho)}{N-2} \sum_{k=1}^{N-1} \sum_{j=k+1}^{N} \cos(\theta_{L_k} - \theta_{L_j})
- \frac{(1-\rho)(1 + (N-1)\rho)}{K[N-2 - \frac{2}{N} \sum_{k=1}^{N-1} \sum_{j=k+1}^{N} \cos(\theta_{L_k} - \theta_{L_j})]} \tag{31}
$$

where it can be shown that $D_\rho$ can take either positive or negative values. If $D_\rho > 0$, an increase in correlation coefficient $\rho$ will increase the error rate. If $D_\rho < 0$, an increase in $\rho$ will lower the error rate. This explains the behavior observed by Ma et al. in [5].

VI. CONCLUSION

Closed-form asymptotic error rate expressions have been derived for EGC and SC diversity operating on multi-branch equally correlated Rician channels. These analytical expressions can be used to predict error rate performances at large SNRs accurately. By analyzing our closed-form solutions we have obtained additional insights into the transmission characteristics of EGC and SC on equally correlated Rician fading channels.

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