Performance Analysis of MIMO Channel Inversion in Rayleigh Fading

Damith Senaratne, *Student Member, IEEE*, Chintha Tellambura, *Fellow, IEEE*, and Himal A. Suraweera, *Member, IEEE*

***Abstract*—The performance of eigenmode transmission over a multiple-input–multiple-output (MIMO) Rayleigh channel under the channel inversion (CI) power allocation scheme is investigated. The moment-generating function of the reciprocal of the received signal-to-noise ratio (SNR) is derived. For the special case when the minimum number of antennas at the transmitter and the receiver is two, the exact closed-form expressions for the probability density function (pdf) and cumulative distribution function (cdf) are also derived. The average symbol error rate (SER) is derived as an application of the results. The extension to Rician and semicorrelated Rayleigh fading scenarios is also outlined.**

***Index Terms*—Channel inversion (CI), eigenmode transmission, multiple-input multiple-output (MIMO), Wishart distribution.**

I. Introduction

MULTIPLE-ANTENNA wireless terminals, which are used along with special signal-processing techniques to achieve diversity and multiplexing benefits, characterize multiple-input–multiple-output (MIMO) wireless technology [3]. MIMO technology exploits the *space dimension*, in addition to the time and frequency dimensions, to deliver data rates and a quality of service unmatched otherwise with comparable spectral resources. MIMO is widely investigated and being deployed (e.g., IEEE 802.11n [4] and Third-Generation Partnership Project (3GPP) Long-Term Evolution (LTE) [5] standards).

A MIMO channel is represented by a *channel matrix*, whose elements are channel gains between transmit–receive antenna pairs. Thus, mathematical tools such as the random matrix theory [6] help the analysis. A MIMO channel can be reduced to a set of noninterfering spatial channels (*virtual channels*) with appropriate transmitter and receiver signal-processing techniques. The ability to independently code, modulate, and power allocate for [7], [8] these virtual channels facilitates the use of single-antenna signal-processing techniques in MIMO. Our focus is on applying the channel inversion (CI) power allocation scheme [9] across the virtual channels realized through eigenmode transmission.

In a multichannel communication system with channel state information (CSI) at the transmitter, CI allocates power among the channels such that the total instantaneous transmit power is held constant and the instantaneous received signal-to-noise ratio (SNR) is identical across the channels. Although CI yields worse capacity than optimal water-filling power allocation, it may be suitable for applications with tight delay constraints [10, Ch. 5]. A *temporal* variant of CI can be used in single-carrier single-antenna systems [9]. Its use with MIMO has also been investigated in literature [11]. MIMO CI [12] considered here, by contrast, is *spatial*, i.e., achieved across spatially multiplexed virtual channels.

Hereafter, CI refers to CI power allocation on the virtual channels produced by eigenmode transmission. References [12]–[14] examine the basics of MIMO CI, whereas [15]–[17] investigate certain variants exhibiting improved capacity. Notably, signal processing for CI in certain MIMO configurations does not even require singular value decomposition of the channel matrix.

Another noteworthy fact is the similarity that CI has with zero-forcing (ZF) beamforming [18], [19]; thus, CI might find use in MIMO and multiuser MIMO scenarios [20]. ZF simply inverts the channel at the transmitter or the receiver. Transmit ZF thus causes the instantaneous transmit power to fluctuate unbounded, making its practical realization challenging. CI is immune to this issue. ZF reception employs nonunitary signal processing, which is susceptible to noise enhancement and correlated noise. CI, with only unitary receiver processing (if at all) [12], [14], is free from these disadvantages. However, CI requires nonunitary transmit processing, and the achievable error rate is the same as that of receive ZF but is worse than that of transmit ZF. These distinctions can be observed in terms of error performance (see Fig. 1).

Our main contribution is the development of a mathematical framework for an $N_r \times N_t$ MIMO system to accurately characterize the per-virtual-channel received SNR Λ under CI. More specifically, we obtain the following:

1) for arbitrary N_t and N_r: the exact moment-generating function (mgf) of Λ^{-1};
2) for the case $\min(N_t, N_r) = 2$: the exact probability density function (pdf) of Λ, the cumulative density function (cdf) of Λ, and the average symbol error rate (SER) for a class of modulation schemes (presented in [1]).

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D. Senaratne and C. Tellambura are with the Department of Electrical and Computer Engineering, University of Alberta, Edmonton, AB T6G 2V4, Canada (e-mail: damith@ece.ualberta.ca; chintha@ece.ualberta.ca).

H. A. Suraweera is with Singapore University of Technology and Design, Singapore 138682 (e-mail: himalsuraweera@sutd.edu.sg).

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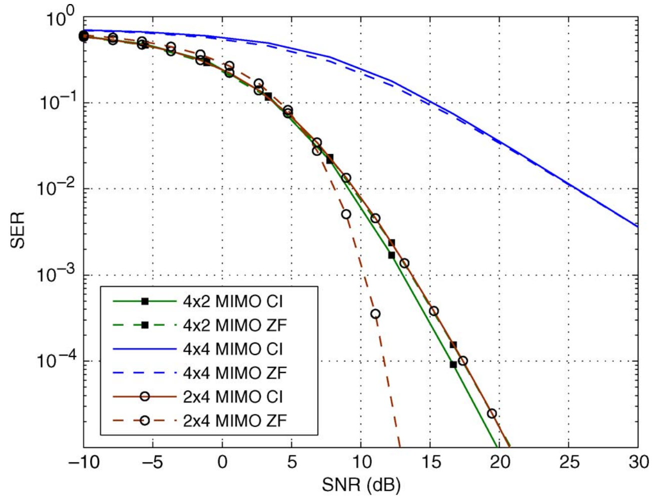

Fig. 1. Average SER for quadratic phase-shift keying (QPSK) on $(N_r \times N_t)$ MIMO systems: ZF versus CI. Transmit ZF is used for the 2×4 configuration, whereas ZF reception is used with the others. 10^6 channel realizations, with 100 QPSK symbols per virtual channel for each, have been simulated.

Although independent Rayleigh fading is assumed for the main results, the extension of some of the results for Rician fading (presented in [2]) and semicorrelated Rayleigh fading is briefly developed.

This paper is organized as follows: The system model is given in Section II, and the mathematical formulation and the numerical results for independent and identically distributed (i.i.d.) Rayleigh fading follow in Sections III and IV, respectively. Section V extends the analysis for Rician fading and semicorrelated Rayleigh fading. Section VI concludes this paper. Proof of the results is annexed.

Notation: $P[\mathcal{A}]$ is the probability of event $\mathcal{A}$. The pdf, cdf, complimentary cdf (ccdf), and the mgf of a random variable X are given by $f_X(x)$, $F_X(x)$, $\bar{F}_X(x)$, and $\mathcal{M}_X(s)$, respectively. The expected value is $\mathcal{E}_X\{\cdot\}$. $\mathcal{I}_\nu(\cdot)$ and $\mathcal{K}_\nu(\cdot)$ are the Modified Bessel functions [21, Sec. 9.6] of the first kind and second kind of order ν, respectively. $\mathcal{G}_{p\ q}^{m\ n}(\cdot|_{b_1,\ldots,b_m,b_{m+1},\ldots,b_q}^{a_1,\ldots,a_n,a_{n+1},\ldots,a_p})$ is the Meijer G function [22, Sec. 9.3]. $\mathcal{Q}(\cdot)$ is the Gaussian Q-function [21, eq. (26.2.3)]. $\log_2(\cdot)$ represents the logarithm to base 2, whereas $k!$ is the factorial of k. $\mathbb{C}^{m\times n}$ is the set of $m \times n$ complex matrices.

$\mathbf{B}^H$, $\|\mathbf{B}\|_F$, trace$(\mathbf{B})$, and eig$(\mathbf{B})$ represent the conjugate transpose, Frobenius norm, trace, and eigenvalues of a matrix $\mathbf{B}$, respectively. The remainder, in a Maclaurin series of a function of x after the x^n term, is denoted by $o(x^n)$.

II. System Model

We consider a MIMO system with N_t and N_r transmit and receive antennas. Let $m = \min(N_t, N_r)$ and $n = \max(N_t, N_r)$. The channel matrix is $\mathbf{H} \in \mathbb{C}^{N_r \times N_t}$. We also assume perfect transmit CSI and the usual additive i.i.d. complex Gaussian noise at the receiver antennas.

Define $\mathcal{W}$ as either $\mathbf{HH}^H$ $(N_t > N_r)$ or $\mathbf{H}^H\mathbf{H}$ $(N_t \le N_r)$. The eigenvalues $\{\lambda_1, \ldots, \lambda_m\}$ of $\mathcal{W}$ sufficiently characterize the MIMO channel. For instance, they relate to the received SNR along the m virtual channels under eigenmode transmission. The total transmit power P is allocated as p_i for each ith virtual channel $i \in \{1, \ldots, m\}$, using CI, causing $\lambda_i p_i = K$ to be held identical for each of them at any given time instant. Thus, we have

$$P = \sum_{i=1}^{m} p_i = K \sum_{i=1}^{m} \lambda_i^{-1}.$$

Let $\Lambda = K/P$. Then, we get

$$\Lambda^{-1} = \sum_{i=1}^{m} \lambda_i^{-1} = \text{trace}(\mathcal{W}^{-1}). \tag{1}$$

Note that (1) holds only under block fading assumptions, which permit averaging out the randomness of additive noise and transmitted data for each channel realization.

III. Mathematical Formulation

This section analyzes MIMO CI systems. The links between the transmit–receive antenna pairs are assumed to undergo i.i.d. Rayleigh fading. Thus, $\mathbf{H} \in \mathbb{C}^{N_r \times N_t}$ becomes a complex Gaussian matrix; $\mathcal{W}$ becomes a rank-m complex central Wishart matrix [7], [23] having n degrees of freedom; and Λ^{-1} becomes the trace of an inverse-Wishart matrix.

A. Arbitrary $m \le n$

The joint pdf of $\lambda_i, i \in \{1, \ldots, m\}$, which are the unordered eigenvalues of $\mathcal{W}$, is given [23], [24] by

$$f_{\lambda_1,\ldots\lambda_m}(\lambda_1, \ldots, \lambda_m) = \frac{e^{-\sum_{i=1}^{m}\lambda_i}}{m!\mathcal{K}_{m,n}} \prod_{i=1}^{m} \lambda_i^{n-m} \prod_{1\le i<j\le m} (\lambda_i - \lambda_j)^2 \tag{2}$$

where $\mathcal{K}_{m,n} = \prod_{k=1}^{m}(m-k)!(n-k)!$. The joint pdf of unordered eigenvalues differs only by a factorial term from that of the ordered eigenvalues. All the unordered eigenvalues have the range $[0, \infty)$, and therefore, using their joint pdf simplifies further derivations.

The factor $\prod_{1\le i<j\le m}(\lambda_i - \lambda_j)^2$ in (2) may be expanded to obtain the more manipulable following form:

$$f_{\lambda_1,\ldots\lambda_m}(\lambda_1, \ldots, \lambda_m) = \frac{e^{-\sum_{i=1}^{m}\lambda_i}}{m!\mathcal{K}_{m,n}} \left(\prod_{i=1}^{m} \lambda_i^{n-m}\right) \times \sum_{\substack{k_1,\ldots,k_m \in \{0,\ldots,2(m-1)\} \\ \sum k_i = m(m-1)}} b(k_1, \ldots, k_m) \lambda_1^{k_1} \cdots \lambda_m^{k_m}. \tag{3}$$

Coefficients $b(k_1, \ldots, k_m)$, corresponding to variables $\lambda_1, \ldots, \lambda_m$ raised to respective powers $k_1, \ldots, k_m$, can be obtained by expanding the factor as a multivariate polynomial. The equality $\sum_{i=1}^{m} k_i = m(m-1)$ is seen to hold for each term of this expansion.

Theorem 1—MGF of Λ^{-1} (for Arbitrary $m \le n$): Let $\lambda_1, \ldots, \lambda_m$ be the unordered eigenvalues of an $m \times m$ complex central Wishart matrix having n degrees of freedom. The mgf of Λ^{-1} in (1) is given by

$$\mathcal{M}_{1/\Lambda}(s) = \frac{2^m s^{-\frac{mn}{2}}}{m!\mathcal{K}_{m,n}} \sum_{\substack{k_1,\ldots,k_m \in \{0,\ldots,2(m-1)\} \\ \sum k_i = m(m-1)}} b(k_1,\ldots,k_m) \times \prod_{i=1}^{m} \mathcal{K}_{k_i+n-m+1}(2\sqrt{s}). \quad (4)$$

Proof: See the Appendix. ■

B. Special Case: $m = 2$

This scenario occurs in any MIMO channel having two antennas at one end and at least two antennas in the other. The MIMO downlink from a multiantenna base station to two-antenna mobile station (as in 4 × 2 3GPP LTE downlink configuration [5]) is an example. Another is the multiuser MIMO downlink [20] from a multiantenna base station to two single-antenna mobile stations. Owing to antenna spacing constraints, equipping a mobile terminal operating in cellular frequency bands (currently, below 4 GHz) with more antennas, is technically challenging. Therefore, the case $m = 2$ is realistic.

For the case $m = 2$, (2) reduces to

$$f_{\lambda_1,\lambda_2}(\lambda_1,\lambda_2) = \frac{1}{2\mathcal{K}_{2,n}} e^{-(\lambda_1+\lambda_2)}(\lambda_1-\lambda_2)^2 \lambda_1^{n-2}\lambda_2^{n-2} \quad (5)$$

where $\mathcal{K}_{2,n}$ simplifies to $(n-1)!(n-2)!$. The distribution of Λ can be found from (1) and (5).

Theorem 2—PDF of Λ (for $m = 2$): Let λ_1 and λ_2 be the unordered eigenvalues of a 2 × 2 complex central Wishart matrix having n degrees of freedom. The pdf of Λ in (1) is given by

$$f_\Lambda(x) = \frac{1}{\mathcal{K}_{2,n}} x^{2(n-1)} e^{-2x} \sum_{k=0}^{2n} \binom{2n}{k} \times \left((n-k-2x)\mathcal{K}_{k-n}(2x) + 2x\mathcal{K}_{k+1-n}(2x)\right). \quad (6)$$

Proof: See the Appendix. ■

Corollary 1—CDF of Λ (for $m = 2$): Let λ_1 and λ_2 be the unordered eigenvalues of a 2 × 2 complex central Wishart matrix having n degrees of freedom. The cdf of Λ in (1) is given by

$$F_\Lambda(x) = \frac{2\sqrt{\pi}}{4^{2n}\mathcal{K}_{2,n}} \sum_{k=0}^{2n} \binom{2n}{k} \times \Big[2(n-k)\mathcal{G}_{2\,3}^{2\,1}\left(4x\big|^{1,2n-0.5}_{3n-k-1,n+k-1,0}\right) - \mathcal{G}_{2\,3}^{2\,1}\left(4x\big|^{1,2n+0.5}_{3n-k,n+k,0}\right) + \mathcal{G}_{2\,3}^{2\,1}\left(4x\big|^{1,2n+0.5}_{3n-k-1,n+k+1,0}\right)\Big]. \quad (7)$$

Proof: See the Appendix. ■

Although relative uncommonly used in the wireless literature, the Meijer G function is well characterized [22, Sec. 9.3].

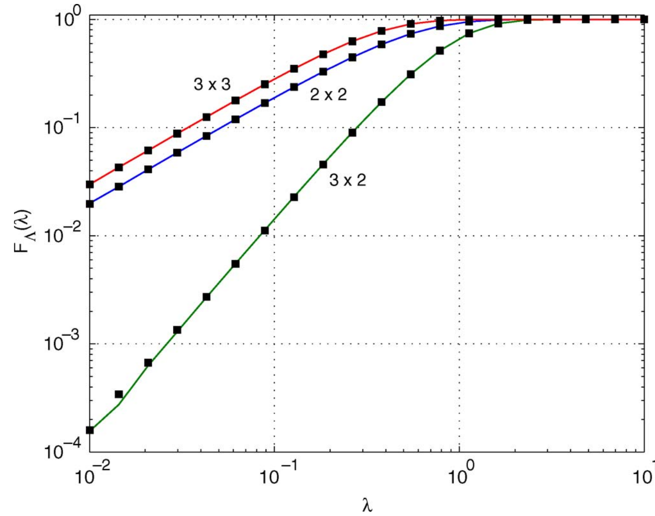

Fig. 2. CDF of Λ over $\{2\times2, 3\times2, 3\times3\}$ MIMO systems using CI; numerical versus simulated (■).

Moreover, it is directly available in the common computational environments including Mathematica, Maple, and MATLAB. Hence, the results can be easily evaluated at high precision.

IV. Numerical Results

This section highlights some applications of the characterization of Λ made in Section III. Numerical results on different performance metrics are presented to establish its validity.

A. Arbitrary $m \le n$

Further derivations based on the result (4) likely require the use of hypergeometric functions of multiple variables and are not examined here.[1] Having an exact expression for $\mathcal{M}_{\Lambda^{-1}}(s)$ is more appealing when it comes to numerical evaluation of the performance metrics for it reduces the number of folded integrals that one may have to evaluate. The ccdf of Λ, which relates to the probability of outage, for instance, is given by [25]

$$\bar{F}_\Lambda(x) = F_{\Lambda^{-1}}\left(\frac{1}{x}\right) = \frac{2}{\pi}\int_0^\infty \frac{\Re\left(\mathcal{M}_{\Lambda^{-1}}(j\omega)\right)}{\omega} \sin\left(\frac{\omega}{x}\right) d\omega \quad (8)$$

whose single integral can be evaluated using a simple quadrature technique. Likewise, evaluating a single integral suffices to obtain the SER [26].

Fig. 2 compares the cdf of Λ: numerical versus simulated (10^6-point semianalytic Monte Carlo). Numerical values have been computed using adaptive quadrature routine `quadl` in MATLAB after applying the variable transformation $\omega = (1+t)^2/(1-t)^2$ to adjust the range of integration. Since CI holds reciprocity and performs similar to a ZF receiver, the diversity order of CI can be deduced (from [27]) to be $|N_t - N_r| + 1$. The slope of cdf curves as $\lambda \to 0$ agrees with this observation.

[1]Native support for special functions of an arbitrary number of variables is not currently available in standard computational environments such as MATLAB and Mathematica. They are nevertheless implementable as cascaded infinite series.

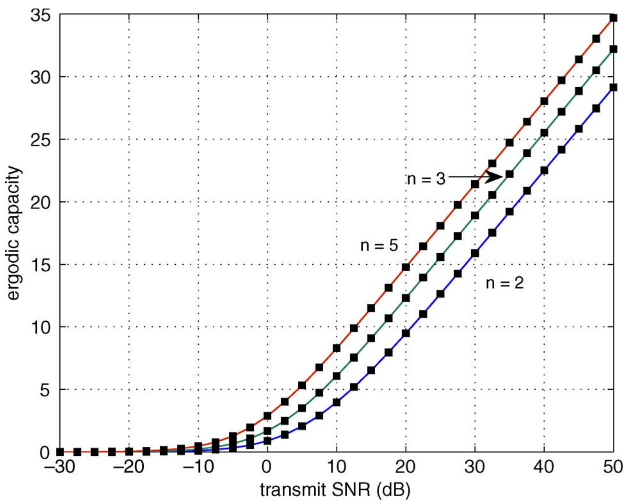

Fig. 3. Ergodic capacity (in bits per second per hertz) for a 2 × n MIMO system using CI; numerical versus simulated (■).

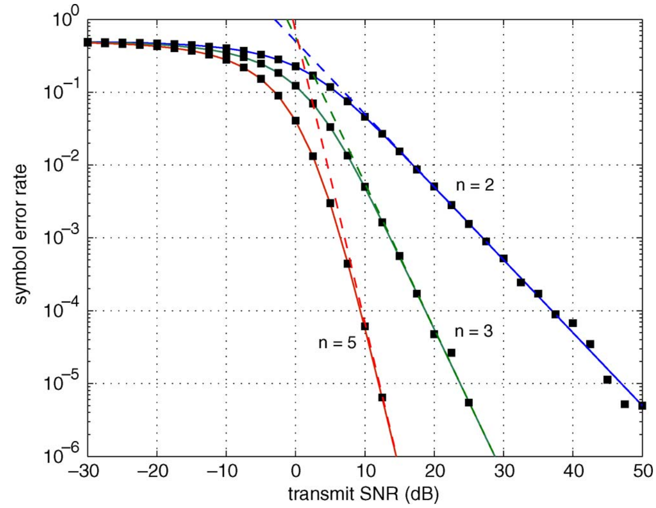

Fig. 4. Average SER for BPSK (i.e., $\mu = \nu = 1$) over a $2 \times n$ MIMO system using CI; analytical versus simulated (■). (Dashed lines) Asymptotes to curves.

B. Special Case $m = 2$

This scenario is more tractable. Three applications of the results are examined next, indirectly verifying (6) and (7).

1) Ergodic Capacity: The Shannon capacity of the MIMO system in concern is given by a random variable $C = \sum_{k=1}^{2} \log_2(1 + p_i\lambda_i/N_0) = 2\log_2(1 + P\Lambda)$, where P denotes the transmit SNR (transmit power normalized w.r.t. the noise variance). Its average, i.e., *ergodic capacity* $\mathcal{E}_C\{C\}$, can be numerically computed as

$$\mathcal{E}_C\{C\} = \int_0^\infty c f_C(c) dc = 2\int_0^\infty \log_2(1 + Px) f_\Lambda(x)\, dx$$

for any given P. Fig. 3 verifies the numerical values thus obtained for $\mathcal{E}_C\{C\}$ for the cases $n \in \{2, 3, 5\}$, against the simulation (10^5-point semianalytic Monte Carlo) results. As expected, the ergodic capacity logarithmically increases with the transmit SNR (i.e., appears as a straight line at high-transmit SNR, when SNR is given in decibels) and increases (although not linearly) with n.

2) Average SER: Since the receiver processing for CI leaves the distribution of additive Gaussian noise unaltered, the average SER P_s under many modulation schemes [28, Ch. 5] can be given by

$$P_s = \mu\mathcal{E}_\Lambda\left\{\mathcal{Q}(\sqrt{2\nu\Lambda P})\right\} \tag{9}$$

where P is the transmit SNR, and μ, ν are constants dependent on the modulation scheme. For example, $\mu = 1$, $\nu = 1$ exactly give the SER for binary phase-shift keying (BPSK), whereas $\mu = 2(M-1)/M$, $\nu = 3/(M^2-1)$ approximate those for other M-ary pulse amplitude modulation schemes. The average SER for such systems can be analytically derived using (9) and the distribution of Λ.

Corollary 2—Average SER (for $m = 2$): Let λ_1 and λ_2 be the unordered eigenvalues of a 2×2 complex central Wishart matrix having n degrees of freedom. The average SER (9) is given by

$$\begin{aligned}P_s = &\frac{\mu}{(4^{2n})\mathcal{K}_{2,n}} \sum_{k=0}^{2n} \binom{2n}{k} \\ &\times \left[2(n-k)\mathcal{G}_{3\,3}^{2\,2}\left(\frac{4}{\nu P}\Big|^{0.5,1,2n-0.5}_{3n-k-1,n+k-1,0}\right)\right. \\ &- \mathcal{G}_{3\,3}^{2\,2}\left(\frac{4}{\nu P}\Big|^{0.5,1,2n+0.5}_{3n-k,n+k,0}\right) \\ &\left.+ \mathcal{G}_{3\,3}^{2\,2}\left(\frac{4}{\nu P}\Big|^{0.5,1,2n+0.5}_{3n-k-1,n+k+1,0}\right)\right].\end{aligned} \tag{10}$$

Proof: See the Appendix. ■

3) High SNR Analysis: The diversity and coding gains of the system can be easily deduced [29, Prop. I] from the least order approximation of x on the pdf of Λ. For the MIMO system in concern, it may be obtained, after some manipulations, using [21, eq. (9.6.9)] on (6).

Corollary 3: Let λ_1 and λ_2 be the unordered eigenvalues of a 2×2 complex central Wishart matrix having n degrees of freedom. The least order approximation of the pdf of Λ in (1) is given by

$$f_\Lambda(x) = \frac{n}{2(n-2)!}x^{n-2} + o(x^{n-2}). \tag{11}$$

Proof: Evidently, only the $k = 0$ term of the summation in (6) contributes to this approximation. Using the first term of each series expansion [21, eq. (4.2.1)] and [21, eq. (9.6.9)], followed by the selection of the least order term of x, yields the result. ■

Fig. 4 shows for the cases $n \in \{2, 3, 5\}$ how the analytical result (10) for the SER of the BPSK compares with the simulation (10^5-point semianalytic Monte Carlo) results. It also shows the asymptotes for the curves computed using (11) based on [29]. A reduction of SER with increased n and a diversity order of $(n-1)$ are observed.

V. Extension to Rician and Correlated Rayleigh Cases

The joint pdf of the unordered eigenvalues of a complex central Wishart distribution resembles those of the noncentral and semicorrelated central Wishart distributions [30]. Therefore, certain results presented in Section III can be generalized for these scenarios. Only pdf results are presented here since the derivation of other results from them (for case $m = 2$) requires no different approach from the central-Wishart case.

A. Rician Fading for the Case $\min(N_t, N_r) = 2$

Without further loss of generality, let $N_t = n \geq 2$ and $N_r = 2$. Suppose that the resulting $2 \times n$ channel matrix $\mathbf{H}$ is of the form $\mathbf{H} = a\mathbf{H}_{\rm sp} + b\mathbf{H}_{\rm sc}$, where $\mathbf{H}_{\rm sp}$ represents the deterministic specular (line-of-sight) component, $\mathbf{H}_{\rm sc} \in \mathbb{C}^{2\times n}$ represents the random scatter component, and $a^2 + b^2 = 1$. The specular component is governed by the directional gains of the antennas, presence of dominant multipaths, etc. $K = a^2\|\mathbf{H}_{\rm sp}\|_F^2/b^2\|\mathbf{H}_{\rm sc}\|_F^2$ is the Rician factor [31]. $\mathbf{\Omega} = (a^2/b^2)\mathbf{H}_{\rm sp}\mathbf{H}_{\rm sp}^H$ is the *noncentrality matrix*.

Let $\{\lambda_1, \lambda_2\} = \text{eig}(\mathbf{H}\mathbf{H}^H)$ and $\{\omega_1, \omega_2 | \omega_1 > \omega_2\} = \text{eig}(\mathbf{\Omega})$. The joint distribution of unordered eigenvalues λ_1, λ_2 is given by [31, eq. (15)]

$$\begin{aligned} f_{\lambda_1,\lambda_2}(\lambda_1, \lambda_2) &= \frac{e^{-(\omega_1+\omega_2)}}{2} \frac{|\lambda_1 - \lambda_2|(\lambda_1\lambda_2)^{\frac{n-2}{2}}}{(\omega_1 - \omega_2)(\omega_1\omega_2)^{\frac{n-2}{2}}} \\ &\quad \times e^{-(\lambda_1+\lambda_2)} \Big(\mathcal{I}_{n-2}(2\sqrt{\omega_1\lambda_1})\mathcal{I}_{n-2}(2\sqrt{\omega_2\lambda_2}) \\ &\quad - \mathcal{I}_{n-2}(2\sqrt{\omega_2\lambda_1})\mathcal{I}_{n-2}(2\sqrt{\omega_1\lambda_2})\Big). \quad (12) \end{aligned}$$

The pdf result corresponding to case $\omega_1 = \omega_2$ can be obtained through the limiting operation $\omega_1 \to \omega_2$ on (12).

Assume perfect transmit CSI and the CI scheme. Given transmit SNR P, the per-channel received SNR $(b^2/2)P\Lambda$ relates to λ_1 and λ_2 through (1).

Theorem 3—PDF of Λ: Let λ_1 and λ_2 be the eigenvalues of a rank-2 complex noncentral Wishart matrix having n degrees of freedom and noncentrality matrix Ω whose eigenvalues are $\{\omega_1, \omega_2 | \omega_1 > \omega_2 > 0\}$. The pdf of Λ in (1) is given by

$$\begin{aligned} f_\Lambda(x) &= \sum_{i=0}^{\infty}\sum_{j=0}^{i} \frac{g_{i,j}(\omega_1, \omega_2)}{(j+n-2)!(i-j+n-2)!j!(i-j)!} \\ &\quad \times \sum_{p=0}^{i+2n-1} \binom{i+2n-1}{p} e^{-2x}x^{i+2n-2} \\ &\quad \times \left(\mathcal{K}_{n+j-p}(2x) - \mathcal{K}_{n+j-p-1}(2x)\right) \quad (13) \end{aligned}$$

where

$$g_{i,j}(\omega_1, \omega_2) = \frac{e^{-(\omega_1+\omega_2)}}{(\omega_1 - \omega_2)} \left(\omega_1^j\omega_2^{(i-j)} - \omega_1^{(i-j)}\omega_2^j\right). \quad (14)$$

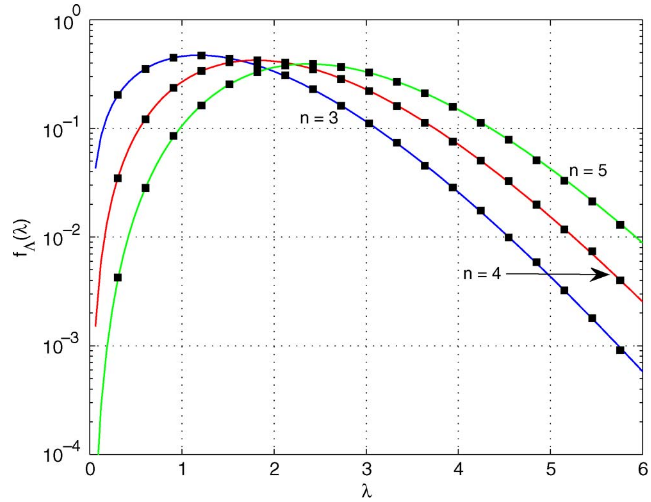

Fig. 5. PDF of Λ over $2 \times n$ MIMO systems using CI; analytical versus simulated (■). Rician fading modeled by a rank-2 noncentrality matrix having eigenvalues [4, 1] assumed.

Proof: The identity [21, eq. (9.6.10)] is used to expand each $\mathcal{I}_n(\cdot)$ as an infinite series. Each term of the resulting cascaded infinite series is of a form similar to (6). Hence, the rest of the proof is similar to that of Theorem 2. ■

Fig. 5 verifies the analytical pdf results for $2 \times n$ MIMO systems using CI under Rician fading against the simulated (10^6-point semianalytic Monte Carlo) results. The noncentrality matrix has eigenvalues $\omega_1 = 4$ and $\omega_2 = 1$.

Equation (13) gets simplified further for the case $\omega_1 \neq 0$, $\omega_2 = 0$ and the limiting case $\omega_1 \to \omega_2$. Corresponding pdf expressions, along with the cdf and average SER results, have been presented and verified in [2]. The least order approximation of the pdf therein is reproduced here. It shows that the diversity order is the same for both Rayleigh and Rician fading.

Corollary 4: Let λ_1 and λ_2 be the eigenvalues of a rank-2 noncentral complex Wishart matrix having n degrees of freedom and noncentrality matrix Ω, whose eigenvalues are $\{\omega_1, \omega_2 | \omega_1 > \omega_2 > 0\}$. The first-order approximation of the pdf of Λ in (1) is given by $f_\Lambda(x) = ax^{n-2} + o(x^{n-2})$, where

$$a = \frac{(n+\omega_1-1)e^{-\omega_2} - (n+\omega_2-1)e^{-\omega_1}}{2(\omega_1 - \omega_2)(n-2)!}. \quad (15)$$

Proof: The proof is similar to that of Corollary 3. ■

Note that the limit of a in (15) as $\omega_1 \to \omega_2$ and $\omega_2 \to 0$ is $n/2(n-2)!$, which corresponds to Rayleigh fading.

B. Min Semicorrelated Rayleigh Fading for the Case $\min(N_t, N_r) = 2$

Semicorrelated MIMO channel models [30] represent the scenarios where only one of the set of transmit or receive antennas is correlated. Under Rayleigh fading, the joint distributions of corresponding unordered eigenvalues hold certain similarities to that of the complex central Wishart matrix.

Let us extend the analysis of $\min(N_t, N_r) = 2$ case in Section III for min semicorrelated Rayleigh fading, where correlation exists only at the terminal having a smaller number of antennas. Without loss of generality, let $N_t = n \geq 2$ and

$N_r = 2$. The channel matrix becomes $\mathbf{H} = \mathbf{\Sigma}^{1/2}\mathbf{H}_w$, where $\mathbf{H}_w \in \mathbb{C}^{N_r \times N_t}$ is complex Gaussian distributed, and $\mathbf{\Sigma}$ is the 2×2 receive correlation matrix whose ordered eigenvalues are σ_1 and σ_2, such that $\sigma_2 > \sigma_1$. The joint distribution of the ordered eigenvalues of $\mathbf{HH}^H$ is given by [32, eq. (17)]. Thus, we get the joint distribution of the unordered eigenvalues $\{\lambda_1, \lambda_2\} = \text{eig}(\mathbf{HH}^H)$ as

$$f_{\lambda_1,\lambda_2}(\lambda_1,\lambda_2) = \frac{K}{2}|\lambda_1 - \lambda_2|(\lambda_1\lambda_2)^{n-2} \times \left(\mathrm{e}^{-(\lambda_1/\sigma_1+\lambda_2/\sigma_2)} - e^{-(\lambda_1/\sigma_2+\lambda_2/\sigma_1)}\right) \quad (16)$$

where

$$K = \frac{(\sigma_1\sigma_2)^{1-n}}{(n-1)!(n-2)!(\sigma_2-\sigma_1)}. \quad (17)$$

The result for the case $\sigma_1 = \sigma_2$ would be given by the limiting operation $\sigma_2 \to \sigma_1$. The distribution of Λ in (1) can be characterized as follows:

Theorem 4—PDF of Λ: Let λ_1 and λ_2 be the eigenvalues of a rank-2 complex central Wishart matrix having n degrees of freedom and a correlation matrix having eigenvalues $\{\sigma_1, \sigma_2 | \sigma_2 > \sigma_1 > 0\}$. The pdf of Λ in (1) is given by

$$f_\lambda(x) = Kx^{2n-2}e^{-(1/\sigma_1+1/\sigma_2)}\left(\mathbb{I}_x(\sigma_1,\sigma_2,0) - x\mathbb{I}_x(\sigma_1,\sigma_2,1) - \mathbb{I}_x(\sigma_2,\sigma_1,0) + x\mathbb{I}_x(\sigma_2,\sigma_1,1)\right)$$

where

$$\mathbb{I}_x(\mu_1,\mu_2,a) = \sum_{k=0}^{2n-1}\binom{2n-1}{k}\left(\frac{\mu_1}{\mu_2}\right)^{\frac{k-n-a+1}{2}} \times \frac{\mathcal{K}_{k-n-a+1}\left(\frac{2x}{\sqrt{\mu_1\mu_2}}\right)}{x^a}.$$

Proof: The proof is omitted, given the similarity to that of Theorem 2. ■

Fig. 6 verifies the analytical pdf results for $2 \times n$ MIMO systems using CI against the simulated results. The receive correlation matrix has eigenvalues $\sigma_1 = 0.1$ and $\sigma_2 = 0.3$. Special cases such as $\sigma_1 = \sigma_2$, as well as the cdf and average SER results, are mathematically tractable but not attempted here.

VI. Conclusion

The performance of MIMO eigenmode transmission under the CI power allocation scheme has been examined. A mathematical framework has been developed to characterize received signal power Λ in CI. The exact mgf of Λ^{-1} (for arbitrary N_t and N_r) and the exact pdf of Λ and the cdf of Λ (for the special case $\min(N_t, N_r) = 2$) have been derived assuming Rayleigh fading. Some extensions have been derived for Rician fading and semicorrelated Rayleigh fading. Numerical results, including that of the average SER, have been provided and verified through simulation to highlight possible applications of the framework developed.

The results confirm the intuition that CI has the diversity order of the weakest eigenmode, which is $|N_t - N_r| + 1$ under both Rayleigh and Rician fading. This is because CI allocates power so that the received SNR is the same over all eigenmodes. Note that ZF reception performs similar to CI, without requiring transmit CSI. Hence, CI is not attractive when the receiver has more antennas than the transmitter. When the transmitter has more antennas, just like transmit ZF, CI does not require spatial processing at the receiver. Therefore, in such point-to-point and point-to-multipoint MIMO configurations, CI is an easier to implement alternative to transmit ZF.

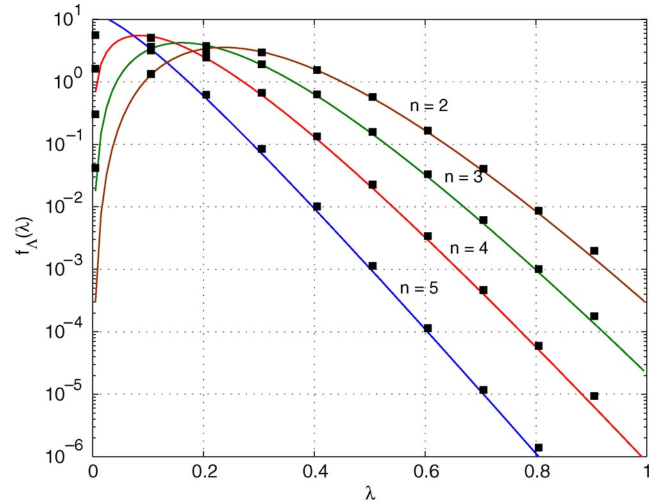

Fig. 6. PDF of Λ over $2 \times n$ MIMO systems using CI; analytical versus simulated (■). Receive correlation matrix assumed to have [0.1, 0.3] as the eigenvalues.

Appendix

This Appendix presents the proofs of Theorems 1 and 2 and Corollaries 1 and 2.

Proof—Theorem 1: MGF of Λ^{-1} (Arbitrary $m \in \mathcal{Z}^+$): The mgf of Λ^{-1} is given by the m-folded integral, i.e.,

$$\mathcal{M}_{1/\Lambda}(s) = \int_{\lambda_1=0}^{\infty} \cdots \int_{\lambda_m=0}^{\infty} e^{-\sum_{i=1}^m \frac{1}{\lambda_i}s} \times f_{\lambda_1,\ldots\lambda_m}(\lambda_1,\ldots,\lambda_m)\, d\lambda_1 \cdots d\lambda_m. \quad (18)$$

Substituting (3) into (18) and simplifying with [33, 4.5.1.(9)], we get

$$\begin{aligned}\mathcal{M}_{1/\Lambda}(s) &= \frac{1}{m!\mathcal{K}_{m,n}} \sum_{\substack{k_1,\ldots,k_m \in \{0,\ldots,2(m-1)\} \\ \sum_{k_i} = m(m-1)}} b(k_1,\ldots,k_m) \times \prod_{i=1}^{m} \int_{\lambda_1=0}^{\infty} \lambda_i^{k_i+n-m} e^{-\left(\lambda_i + \frac{s}{\lambda_i}\right)} d\lambda_i \\ &= \frac{1}{m!\mathcal{K}_{m,n}} \sum_{\substack{k_1,\ldots,k_m \in \{0,\ldots,2(m-1)\} \\ \sum_{k_i} = m(m-1)}} b(k_1,\ldots,k_m) \times \prod_{i=1}^{m} \left(2s^{-\frac{k_i+nm+1}{2}} \mathcal{K}_{k_i+n-m+1}(2\sqrt{s})\right) \end{aligned} \quad (19)$$

and, hence, (4). ■

Proof—Theorem 2: PDF of Λ (Case: $m = 2$): From the definition of the cdf

$$F_\Lambda(x) = \mathcal{P}\left[\frac{\lambda_1\lambda_2}{\lambda_1+\lambda_2} \leq x\right] = \mathcal{P}\left[\lambda_1(\lambda_2 - x) \leq \lambda_2 x\right] = \int_0^x \underbrace{\bar{F}_{\lambda_1|\lambda_2}\left(\frac{\lambda_2 x}{\lambda_2 - x}\bigg|\lambda_2\right)}_{\doteq 1} f_{\lambda_2}(\lambda_2)\, d\lambda_2 + \int_x^\infty F_{\lambda_1|\lambda_2}\left(\frac{\lambda_2 x}{\lambda_2 - x}\bigg|\lambda_2\right) f_{\lambda_2}(\lambda_2)\, d\lambda_2$$

$$\bar{F}_\Lambda(x) = \int_x^\infty \bar{F}_{\lambda_1|\lambda_2}\left(\frac{\lambda_2 x}{\lambda_2 - x}\bigg|\lambda_2\right) f_{\lambda_2}(\lambda_2)\, d\lambda_2.$$

Differentiating it with respect to x, we get

$$f_\Lambda(x) = \int_x^\infty \frac{\lambda_2^2}{(\lambda_2 - x)^2} f_{\lambda_1|\lambda_2}\left(\frac{\lambda_2 x}{\lambda_2 - x}\bigg|\lambda_2\right) f_{\lambda_2}(\lambda_2) d\lambda_2 = \int_x^\infty \frac{\lambda_2^2}{(\lambda_2 - x)^2} f_{\lambda_1,\lambda_2}\left(\frac{\lambda_2 x}{\lambda_2 - x}, \lambda_2\right) d\lambda_2 = \int_0^\infty \left(\frac{t+x}{t}\right)^2 f_{\lambda_1,\lambda_2}\left(\frac{x(t+x)}{t}, t+x\right) dt. \quad (20)$$

Substituting (5) into (20) and using the binomial expansion and [22, eqs. (3.471.9, 8.471.1)], we get

$$f_\Lambda(x) = \frac{1}{2\mathcal{K}_{2,n}} \int_0^\infty \left(\frac{t+x}{t}\right)^2 e^{-\frac{(t+x)^2}{t}} \times \left(\frac{(t+x)^2(t-x)^2}{t^2}\right)\left(\frac{x(t+x)^2}{t}\right)^{n-2} dt$$
$$= \frac{1}{2\mathcal{K}_{2,n}} x^{n-2} e^{-2x} \sum_{k=0}^{2n} \binom{2n}{k} x^{2n-k} \times \int_0^\infty \frac{(t-x)^2}{t^{n+2-k}} e^{-\left(t+\frac{x^2}{t}\right)} dt$$
$$= \frac{1}{\mathcal{K}_{2,n}} x^{2(n-1)} e^{-2x} \sum_{k=0}^{2n} \binom{2n}{k} \times \left((n-k-2x)\mathcal{K}_{k-n}(2x) + 2x\mathcal{K}_{k+1-n}(2x)\right). \quad (21)$$

■

Proof—Corollary 1: CDF of Λ (Case: $m = 2$): Consider the following integral, which can be simplified into a single *Meijer G function* [34, p. 419] first by using [22, eq. (9.34.4)] and then by applying [35, eq. (7.34.21.2.1)] with a substitution $u \doteq 2t$:

$$\int_0^x t^\mu \mathcal{K}_\nu(t) e^{-t} dt = \sqrt{\pi} \int_0^x t^\mu \mathcal{G}_{1\,2}^{2\,0}\left(2t\big|^{0.5}_{-\nu,\nu}\right) dt = \frac{\sqrt{\pi}}{2^{\mu+1}} \int_0^{2x} u^\mu \mathcal{G}_{1\,2}^{2\,0}\left(u\big|^{0.5}_{-\nu,\nu}\right) du = \frac{\sqrt{\pi}}{2^{\mu+1}} \mathcal{G}_{2\,3}^{2\,1}\left(2x\big|^{1,\mu+1.5}_{\mu-\nu+1,\mu+\nu+1,0}\right). \quad (22)$$

Now, let us consider the cdf of Λ, i.e.,

$$F_\Lambda(x) = \int_0^x f_\Lambda(t)\, dt = \frac{1}{\mathcal{K}_{2,n}} \sum_{k=0}^{2n} \binom{2n}{k} \times \left((n-k)\int_0^x t^{2n-2} e^{-2t} \mathcal{K}_{k-n}(2t)\, dt - 2\int_0^x t^{2n-1} e^{-2t} \mathcal{K}_{k-n}(2t)\, dt + 2\int_0^x t^{2n-1} e^{-2t} \mathcal{K}_{k+1-n}(2t)\, dt\right). \quad (23)$$

Applying the result of (22) in (23), we get (7). ■

Proof—Corollary 2: Average SER (Case: $m = 2$): Equation (9) can be simplified as follows using integration by parts and the *Leibniz's rule for differentiation* [36, eq. (32)], [37]:

$$P_s = \int_0^\infty \mu \mathcal{Q}(\sqrt{2\nu x P})\, dF_\Lambda(x) = \mu \int_0^\infty \frac{e^{-\nu P x}}{\sqrt{2\pi}} \cdot \frac{\sqrt{2\nu P}}{2\sqrt{x}} F_\Lambda(x)\, dx. \quad (24)$$

Consider the integral

$$\mathbb{I}(q,\alpha,\beta,\gamma) = \int_0^\infty x^{-0.5} e^{-qx} \mathcal{G}_{2\,3}^{2\,1}\left(4x\big|^{1,\alpha}_{\beta,\gamma,0}\right) dx = \mathcal{L}\left\{x^{-0.5}\mathcal{G}_{2\,3}^{2\,1}\left(4x\big|^{1,\alpha}_{\beta,\gamma,0}\right)\right\}\Big|_{s\doteq q}$$

where q, α, β, and γ are positive reals, and $\mathcal{L}\{.\}$ denotes the Laplace transform. It can be solved using [33, eq. (4.23.34)] to get

$$\mathbb{I}(q,\alpha,\beta,\gamma) = q^{-0.5} \mathcal{G}_{3\,3}^{2\,2}\left(\frac{4}{q}\big|^{0.5,1,\alpha}_{\beta,\gamma,0}\right). \quad (25)$$

Substituting (7) into (24) and using (25) complete the proof. ■

References

[1] D. Senaratne and C. Tellambura, “Performance analysis of channel inversion over MIMO channels,” in *Proc. IEEE Globecom*, Honolulu, HI, Dec. 2009.

[2] D. Senaratne, C. Tellambura, and H. A. Suraweera, “Channel inversion in MIMO systems over Rician fading,” in *Proc. IEEE Globecom*, Miami, FL, Dec. 2010.

[3] G. J. Foschini, “Layered space-time architecture for wireless communication in a fading environment when using multi-element antennas,” *Bell Labs Tech. J.*, vol. 1, no. 2, pp. 41–59, Summer 1996.

[4] *Wireless LAN Medium Access Control (MAC) and Physical Layer (PHY) Specifications Amendment 5: Enhancements for Higher Throughput*, IEEE Std. 802.11n, Oct. 2009.

[5] *Technical Specification Group Radio Access Network; Evolved Universal Terrestrial Radio Access (E-UTRA); LTE Physical Layer; General Description (Rel. 9)*, 3GPP TS 36.201 V9.1, Mar. 2010.

[6] A. M. Tulino and S. Verdu, *Random Matrix Theory and Wireless Communications*, 1st ed. Hanover, MA: Now, 2004, ser. Foundations and Trends in Communications and Information Theory.

[7] I. E. Telatar, “Capacity of multi-antenna Gaussian channels,” Bell Lab., Lucent Technol., Murray Hill, NJ, Tech. Rep., Oct. 1995.

[8] A. Zanella and M. Chiani, “Analytical comparison of power allocation methods in MIMO systems with singular value decomposition,” in *Proc. IEEE Globecom*, Honolulu, HI, Dec. 2009.

[9] A. J. Goldsmith and P. P. Varaiya, “Capacity of fading channels with channel side information,” *IEEE Trans. Inf. Theory*, vol. 43, no. 6, pp. 1986–1992, Nov. 1997.

[10] D. Tse and P. Viswanath, *Fundamentals of Wireless Communication*. Cambridge, U.K.: Cambridge Univ. Press, Jun. 2005.

[11] A. Maaref and S. Aïssa, “Impact of spatial fading correlation and keyhole on the capacity of MIMO systems with transmitter and receiver CSI,” *IEEE Trans. Wireless Commun.*, vol. 7, no. 8, pp. 3218–3229, Aug. 2008.

[12] T. Haustein, C. von Helmolt, E. Jorswieck, V. Jungnickel, and V. Pohl, “Performance of MIMO systems with channel inversion,” in *Proc. VTC*, Birmingham, AL, May 2002, vol. 1, pp. 35–39.

[13] E. Jorswieck, G. Wunder, V. Jungnickel, and T. Haustein, “Inverse eigenvalue statistics for Rayleigh and Rician MIMO channels,” in *Proc. Inst. Elect. Eng. Semin. MIMO, Commun. Syst. Concept Implement.*, London, U.K., Dec. 2001, pp. 3/1–3/6.

[14] V. Jungnickel, T. Haustein, E. Jorswieck, and C. von Helmolt, “A MIMO WLAN based on linear channel inversion,” in *Proc. Inst. Elect. Eng. Semin. MIMO Commun. Syst. Concept Implement.*, London, U.K., Dec. 2001, pp. 20/1–20/6.

[15] V. Jungnickel, T. Haustein, V. Pohl, and C. von Helmolt, “Link adaptation in a multi-antenna system,” in *Proc. VTC*, Jeju, Korea, Apr. 2003, vol. 2, pp. 862–866.

[16] C. B. Peel, B. M. Hochwald, and A. L. Swindlehurst, “A vector-perturbation technique for near-capacity multiantenna multiuser communication—Part I: Channel inversion and regularization,” *IEEE Trans. Commun.*, vol. 53, no. 1, pp. 195–202, Jan. 2005.

[17] C. Masouros and E. Alsusa, “Dynamic linear precoding for the exploitation of known interference in MIMO broadcast systems,” *IEEE Trans. Wireless Commun.*, vol. 8, no. 3, pp. 1396–1404, Mar. 2009.

[18] R. U. Nabar, O. Oyman, H. Bölcskei, and A. J. Paulraj, “Capacity scaling laws in MIMO wireless networks,” in *Proc. Allerton Conf. Commun., Control, Comput.*, Monticello, IL, Oct. 2003, pp. 378–389.

[19] M. Matthaiou, C. Zhong, and T. Ratnarajah, “Novel generic bounds on the sum rate of MIMO ZF receivers,” *IEEE Trans. Signal Process.*, vol. 59, no. 9, pp. 4341–4353, Sep. 2011.

[20] G. Caire and S. Shamai, “On the achievable throughput of a multiantenna Gaussian broadcast channel,” *IEEE Trans. Inf. Theory*, vol. 49, no. 7, pp. 1691–1706, Jul. 2003.

[21] M. Abramowitz and I. Stegun, *Handbook of Mathematical Functions*. New York: Dover, 1970.

[22] I. Gradshteyn and I. Ryzhik, *Table of Integrals, Series, and Products*, 7th ed. New York: Academic, 2000.

[23] A. T. James, “Distributions of matrix variates and latent roots derived from normal samples,” *Ann. Math. Stat.*, vol. 35, no. 2, pp. 475–501, Jun. 1964.

[24] M. Chiani and A. Zanella, “Joint distribution of an arbitrary subset of the ordered eigenvalues of Wishart matrices,” in *Proc. IEEE Int. Symp. PIMRC*, Cannes, France, Sep. 2008.

[25] A. H. Nuttall, “Alternate forms for numerical evaluation of cumulative probability distributions directly from characteristic functions,” *Proc. IEEE*, vol. 58, no. 11, pp. 1872–1873, Nov. 1970.

[26] C. Tellambura, M. Soysa, and D. Senaratne, “Performance analysis of wireless systems from the mgf of the reciprocal of the signal-to-noise ratio,” *IEEE Commun. Lett.*, vol. 15, no. 1, pp. 55–57, Jan. 2011.

[27] D. Gore, R. W. Heath, Jr., and A. Paulraj, “On performance of the zero forcing receiver in presence of transmit correlation,” in *Proc. IEEE ISIT*, Lausanne, Switzerland, Jun. 2002, p. 159.

[28] J. G. Proakis, *Digital Communications*, 4th ed. New York: McGraw-Hill, 2000.

[29] Z. Wang and G. Giannakis, “A simple and general parameterization quantifying performance in fading channels,” *IEEE Trans. Commun.*, vol. 51, no. 8, pp. 1389–1398, Aug. 2003.

[30] L. Ordoez, D. Palomar, and J. Fonollosa, “Ordered eigenvalues of a general class of Hermitian random matrices with application to the performance analysis of MIMO systems,” *IEEE Trans. Signal Process.*, vol. 57, no. 2, pp. 672–689, Feb. 2009.

[31] P. J. Smith and L. M. Garth, “Exact capacity distribution for dual MIMO systems in Ricean fading,” *IEEE Commun. Lett.*, vol. 8, no. 1, pp. 18–20, Jan. 2004.

[32] M. Chiani, M. Z. Win, and A. Zanella, “On the capacity of spatially correlated MIMO Rayleigh-fading channels,” *IEEE Trans. Inf. Theory*, vol. 49, no. 10, pp. 2363–2371, Oct. 2003.

[33] A. Erdelyi, W. Magnus, F. Oberhettinger, and F. G. Tricomi, *Tables of Integral Transforms*. New York: McGraw-Hill, 1954.

[34] L. C. Andrews, *Special Functions of Mathematics for Engineers*, 2nd ed. New York: McGraw-Hill, 1992.

[35] The Wolfram Functions Site. [Online]. Available: http://functions.wolfram.com/07.34.21.0002.01

[36] Y. Chen and C. Tellambura, “Distribution functions of selection combiner output in equally correlated Rayleigh, Rician, and Nakagami-m fading channels,” *IEEE Trans. Commun.*, vol. 52, no. 11, pp. 1948–1956, Nov. 2004.

[37] A. Zanella, M. Chiani, and M. Z. Win, “Performance of MIMO MRC in correlated Rayleigh fading environments,” in *Proc. VTC*, Stockholm, Sweden, May 2005, pp. 1633–1637.

Damith Senaratne (S’06) received the B.Sc. degree (first-class honors) and the M.Sc. degree in telecommunications from the University of Moratuwa, Moratuwa, Sri Lanka, in 2005 and 2008, respectively. He is currently working toward the Ph.D. degree with the Department of Electrical and Computer Engineering, University of Alberta, Edmonton, AB, Canada.

His research interests include multiple-input–multiple-output signal processing and performance analysis of wireless communication systems.

Chintha Tellambura (F’11) received the B.Sc. degree (first-class honors) from the University of Moratuwa, Moratuwa, Sri Lanka, in 1986, the M.Sc. degree in electronics from the University of London, London, U.K., in 1988, and the Ph.D. degree in electrical engineering from the University of Victoria, Victoria, BC, Canada, in 1993.

He was a Postdoctoral Research Fellow with the University of Victoria (1993–1994) and the University of Bradford, Bradford, U.K. (1995–1996). He was with Monash University, Melbourne, Australia, from 1997 to 2002. He is currently a Professor with the Department of Electrical and Computer Engineering, University of Alberta, Edmonton, AB, Canada. His research interests include communication theory dealing with the wireless physical layer.

Dr. Tellambura is an Associate Editor for the IEEE Transactions on Communications and the Area Editor for Wireless Communications Systems and Theory for the IEEE Transactions on Wireless Communications. He was the Chair for the Communication Theory Symposium at Globecom 2005, which was held in St. Louis, MO.

Himal A. Suraweera (S'04–M'07) was born in Kurunegala, Sri Lanka. He received the B.Sc. degree (first-class honors) in electrical and electronics engineering from Peradeniya University, Peradeniya, Sri Lanka, in 2001 and the Ph.D. degree from Monash University, Melbourne, Australia, in 2007.

From 2001 to 2002, he was with the Department of Electrical and Electronics Engineering, Peradeniya University, as an Instructor. From October 2006 to January 2007, he was with Monash University as a Research Associate. From February 2007 to June 2009, he was with the Center for Telecommunications and Microelectronics, Victoria University, Melbourne, Australia, as a Research Fellow. From July 2009 to January 2011, he was with the Department of Electrical and Computer Engineering, National University of Singapore, Singapore, as a Research Fellow. He is currently a Research Fellow with Singapore University of Technology and Design, Singapore. His research interests include relay networks, cognitive radio, multiple-input multiple-output, and orthogonal frequency-division multiplexing.

Dr. Suraweera is an Associate Editor of the IEEE COMMUNICATIONS LETTERS. He has served or is serving as a Technical Program Committee Member for international conferences such as the 2012 International Conference on Communications, the 2011 Global communications Conference, and the 2011 Wireless Communications Networking Conference. He received an International Postgraduate Research Scholarship from the Australian Commonwealth during 2003–2006, the 2007 Mollie Holman Doctoral and 2007 Kenneth Hunt Medals for his doctoral thesis upon graduating from Monash University, and an IEEE COMMUNICATIONS LETTERS exemplary reviewer certificate for 2009.