Gallager's exponent for MIMO channels: A reliability-rate tradeoff

The MIT Faculty has made this article openly available. Please share how this access benefits you. Your story matters.

<table>
<thead>
<tr>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>As Published</td>
<td><a href="http://dx.doi.org/10.1109/TCOMM.2008.04.050673">http://dx.doi.org/10.1109/TCOMM.2008.04.050673</a></td>
</tr>
<tr>
<td>Publisher</td>
<td>Institute of Electrical and Electronics Engineers</td>
</tr>
<tr>
<td>Version</td>
<td>Final published version</td>
</tr>
<tr>
<td>Accessed</td>
<td>Sun Mar 10 18:41:58 EDT 2019</td>
</tr>
<tr>
<td>Citable Link</td>
<td><a href="http://hdl.handle.net/1721.1/66695">http://hdl.handle.net/1721.1/66695</a></td>
</tr>
<tr>
<td>Terms of Use</td>
<td>Article is made available in accordance with the publisher's policy and may be subject to US copyright law. Please refer to the publisher's site for terms of use.</td>
</tr>
<tr>
<td>Detailed Terms</td>
<td></td>
</tr>
</tbody>
</table>
Gallager’s Exponent for MIMO Channels: 
A Reliability-Rate Tradeoff

Hyundong Shin, Member, IEEE, and Moe Z. Win, Fellow, IEEE

Abstract—In this paper, we derive Gallager’s random coding error exponent for multiple-input multiple-output (MIMO) Rayleigh block-fading channels, assuming no channel-state information (CSI) at the transmitter and perfect CSI at the receiver. This measure gives insight into a fundamental tradeoff between the communication reliability and information rate of MIMO channels, enabling to determine the required codeword length to achieve a prescribed error probability at a given rate below the channel capacity. We quantify the effects of the number of antennas, channel coherence time, and spatial fading correlation on the MIMO exponent. In addition, the general formulae for the ergodic capacity and the cutoff rate in the presence of spatial correlation are deduced from the exponent expressions. These formulae are applicable to arbitrary structures of transmit and receive correlation, encompassing all the previously known results as special cases of our expressions.

Index Terms—Block fading, channel capacity, cutoff rate, multiple-input multiple-output (MIMO) system, random coding error exponent, spatial fading correlation.

I. INTRODUCTION

The channel capacity is a crucial information-theoretic perspective that determines the fundamental limit on achievable information rates over a communication channel [1]. However, since the channel capacity alone gives only the knowledge of the maximum achievable rate, a stronger form of the channel coding theorem has been pursued to determine the behavior of the error probability $P_e$ as a function of the codeword length $N$ and information rate $R$ [2]–[4]. The reliability function or the error exponent of a communication system is defined by [2]

$$E(R) \triangleq \limsup_{N \to \infty} - \frac{\ln P_e^{\text{opt}}(R, N)}{N}$$

where $P_e^{\text{opt}}(R, N)$ is the average block error probability for the optimal block code of length $N$ and rate $R$. The error exponent describes a decaying rate in the error probability as a function of the codeword length, and hence serves to indicate how difficult it may be to achieve a certain level of reliability in communication at a rate below the channel capacity. Although it is difficult to find the exact error exponent, its classical lower bound is available due to Gallager [3]. This lower bound is known as the random coding error exponent or Gallager’s exponent in honor of his discovery, and has been used to estimate the codeword length required to achieve a prescribed error probability [5]–[7].

The random coding exponent was extensively studied for single-input single-output (SISO) and single-input multiple-output (SIMO) flat-fading channels with average or peak power constraint [5], [6]. For SIMO block-fading channels, the random coding exponent was derived in [8] with perfect channel-state information (CSI) at the receiver, where it has been shown that although the capacity is independent of the channel coherence time (first asserted in [9] and also recently addressed in [10] and [11] for multiple-antenna communication), the error exponent suffers a considerable decrease due to a reduction in the effective codeword length as the coherence time increases.2 Therefore, this so-called channel-incorrect effect reduces the communication reliability. While there are numerous prior investigations (following the seminal work of [12]–[15]) on the capacity for multiple-input multiple-output (MIMO) channels [16]–[24], only limited results are available for error exponents. The random coding exponent was given implicitly in [16] (without final analytical expressions) for independent and identically distributed (i.i.d.) Rayleigh-fading MIMO channels with a single-symbol coherence time, perfect receive CSI, and Gaussian inputs subject to the average power constraint.3 Also, the random coding exponent was analyzed in [7] for MIMO i.i.d. block-fading channels with no CSI and isotropically unitary inputs subject to the average power constraint.

In this paper, taking into account spatial fading correlation, we derive Gallager’s exponent for MIMO channels. We consider a Rayleigh block-fading channel with Gaussian inputs subject to the average power constraint and perfect CSI at the receiver. Our results resort to the methodology

1In the following, we will use the term “error probability” to denote the average block error probability.

2This observation is parallel to the divergent behavior of the channel capacity and cutoff rate of a channel with block memory [9].

3As the number of transmit and receive antennas tends to infinity, the asymptotic error exponent was found in [17] using the Gaussian behavior of the random determinant.
developed in [23] and [24], which is based on the finite random matrix theory [25], [26]. The MIMO exponent obtained in the paper provides insight into a fundamental tradeoff between the communication reliability and information rate (below the channel capacity), enabling to determine the required codeword length for a prescribed error probability. It is interesting to note that as a special case of this reliability–rate tradeoff, one can obtain the diversity–multiplexing tradeoff of MIMO channels [27], which is a scaled version of the asymptotic one can obtain the diversity–multiplexing tradeoff of MIMO channels [27], which is a scaled version of the asymptotic reliability–rate tradeoff at high signal-to-noise ratio (SNR).4

We quantify the effects of the number of antennas, the channel coherence time, and the amount of spatial fading correlation on the MIMO exponent. Moreover, the general formulae for the ergodic capacity and cutoff rate are deduced from the exponent expressions. In particular, our capacity formula embraces all the previously known results for i.i.d. [16], [22], one-sided correlated [20], [21], and doubly correlated [23] channels.

This paper is organized as follows. In Section II, signal and channel models are presented. Section III derives the expression for the MIMO random coding exponent. Section IV gives proofs of the main results presented in Section III. In Section V, some numerical results are provided to illustrate the reliability–rate tradeoff in MIMO block-fading channels. Finally, Section VI concludes the paper.

Notation: Throughout the paper, we shall use the following notation. \( \mathbb{N}, \mathbb{R}, \) and \( \mathbb{C} \) denote the natural numbers and the fields of real and complex numbers, respectively. The superscripts \( T \) and \( \dagger \) stand for the transpose and transpose conjugate, respectively. \( I_n \) is the \( n \times n \) identity matrix and \( (A)_{ij} \) denotes the matrix with the \( (i,j) \)-th entry \( A_{ij} \). The trace operator of a square matrix \( A \) is denoted by \( \text{tr} (A) \) and \( \text{etr} (A) = e^{\text{tr}(A)} \). The Kronecker product of matrices is denoted by \( \otimes \). By \( A > 0 \), we denote \( A \) is positive definite. For a Hermitian matrix \( A \in \mathbb{C}^{n \times n} \), \( \lambda_1 (A) \geq \lambda_2 (A) \geq \ldots \geq \lambda_n (A) \) denotes the eigenvalues of \( A \) in decreasing order and \( \lambda (A) \in \mathbb{R}^n \) denote the vector of the ordered eigenvalues, whose \( i \)-th element is \( \lambda_i (A) \). Also, \( \varrho (A) \) denotes the number of distinct eigenvalues of \( A \), and \( \chi_k (A) \), \( k = 1, 2, \ldots, \varrho (A) \), denote the distinct eigenvalues of \( A \) in decreasing order and its multiplicity, respectively, that is, \( \lambda_{(1)} (A) > \lambda_{(2)} (A) > \ldots > \lambda_{(k)} (A) \) and \( \sum_{k=1}^{\varrho (A)} \chi_k (A) = n \). Finally, we shall use the notation \( X \in \mathbb{C}^{m \times n} \sim \mathcal{N}_{m,n} (M, \Sigma, \Psi) \) to denote that a random matrix \( X \) is (matrix-variate) Gaussian distributed with the probability density function (pdf)

\[ p_X (X) = \frac{1}{\pi^{mn} \det (\Sigma)^{n/2} \det (\Psi)^{m/2}} \exp \left\{ - \frac{1}{2} (X - M)^\dagger (\Sigma^{-1} (X - M)) \right\} \]

where \( M \in \mathbb{C}^{m \times n} \), \( \Sigma \in \mathbb{C}^{m \times m} > 0 \), and \( \Psi \in \mathbb{C}^{n \times n} > 0 \).

II. SIGNAL AND CHANNEL MODELS

We consider a MIMO system with \( n_T \) transmit and \( n_R \) receive antennas, where the channel remains constant for \( N_c \) symbol periods and changes independently to a new value for each coherence time, i.e., every \( N_c \) symbols. Since the propagation coefficients independently acquire new values for every coherence interval, the channel is memoryless when considering a block length of \( N_c \) symbols as one channel use with input and output signals of dimension \( n_T \times N_c \) and \( n_R \times N_c \), respectively.

For an observation interval of \( N_b N_c \) symbol periods, the received signal is a sequence \( \{ Y_1, Y_2, \ldots, Y_{N_b} \} \), each \( Y_k \in \mathbb{C}^{n_T \times N_c} \), \( k = 1, 2, \ldots, N_b \), is given by

\[ Y_k = H_k X_k + W_k \]

where \( X_k \in \mathbb{C}^{n_T \times N_c} \) are the transmitted signal matrices, \( H_k \in \mathbb{C}^{n_R \times n_T} \) are the channel matrices, and \( W_k \sim \mathcal{N}_{n_R, N_R} (0, N_0 I_{n_R \times n_R}) \) are the additive white Gaussian noise (AWGN) matrices. Fig. 1 shows a communication link with \( n_T \) transmit and \( n_R \) receive antennas to communicate at a rate \( R \) over \( N_b \) independent \( N_c \)-symbol coherence intervals.

Let \( p_X (X) \) be the input probability assignment for \( X \) with the covariance \( \text{Cov} \{ \text{vec}(X^\dagger) \} = Q^T \otimes I_{N_c} \) subject to the average power constraint of the form

\[ \mathbb{E} \{ \text{tr}(XX^\dagger) \} = \text{tr}(Q^T \otimes I_{N_c}) \leq N_c P \]

(3)
where $Q$ is the $n_T \times n_T$ positive semidefinite matrix and $P$ is the total transmit power over $n_T$ transmit antennas. Taking into account spatial fading correlation at both the transmitter and the receiver, we consider the channel matrix $H$ is given by [18], [19]

$$H = \Phi_R^{1/2} H_0 \Phi_T^{1/2}$$

(4)

where $\Phi_T \in \mathbb{C}^{n_T \times n_T} > 0$ and $\Phi_R \in \mathbb{C}^{n_R \times n_R} > 0$ are the transmit and receive correlation matrices, respectively, and $H_0 \sim \mathcal{N}_{n_R \times n_T} (0, I_{n_R}, I_{n_T})$ is a matrix with i.i.d., zero-mean, unit-variance, complex Gaussian entries. The $(i,j)$-th entry $H_{ij}$, $i = 1, 2, \ldots, n_R$, $j = 1, 2, \ldots, n_T$, of $H$ is a complex propagation coefficient between the $j$-th transmit antenna and the $i$-th receive antenna with $E \{|H_{ij}|^2\} = 1$. Note that $H \sim \mathcal{N}_{n_R \times n_T} (0, \Phi_R, \Phi_T)$ [22]. With perfect CSI at the receiver, we have the transition pdf

$$p(Y|X,H) = \frac{\text{etr} \left\{ -\frac{1}{N_0} (Y - HX)(Y - HX)^\dagger \right\}}{(\pi N_0)^{n_R N_c}}$$

(5)

which completely characterizes MIMO block-fading channels.

In what follows, we define the random matrix $\Theta \in \mathbb{C}^{m \times m} > 0$ as

$$\Theta \triangleq \begin{cases} HH^\dagger, & \text{if } n_R \leq n_T \\ H^\dagger H, & \text{otherwise} \end{cases}$$

(6)

which is a matrix quadratic form in complex Gaussian matrices, denoted by $\Theta \sim \mathcal{Q}_m$ [22], where $m = \min \{n_T, n_R\}$, $n = \max \{n_T, n_R\}$, and

$$\mathcal{Q}_m \triangleq \left\{ \Phi_1 \in \mathbb{C}^{m \times m}, \Phi_2 \in \mathbb{C}^{n \times n} : (\Phi_1, \Phi_2) \right\}$$

(7)

The pdf of $\Theta \sim \mathcal{Q}_m$ is given by [23]

$$p(\Theta) = \frac{1}{\Gamma_m(n)} \frac{\det (\Phi_1)^{-n} \det (\Phi_2)^{-m}}{\Gamma(n)} \times \det (\Theta)^{n-m} \tilde{f}_q^{(n)} (\Phi_1^{-1} \Theta, \Phi_2^{-1})$$

(8)

where $\Gamma_m(n) = \pi^{m(m-1)/2} I_{m-1} \Gamma (\alpha - i), \Re (\alpha) > m - 1$, is the complex multivariate gamma function, $\Gamma (\cdot)$ is the Euler gamma function, and $\tilde{f}_q^{(n)} (\cdot)$ is the hypergeometric function of two Hermitian matrices [25, eq. (88)].

III. MIMO EXPONENT: RELIABILITY–RATE TRADEOFF

This section is based on Gallager’s random coding bound on the error probability of maximum-likelihood (ML) decoding for a channel with continuous inputs and outputs [3]. Notably, the bound determines the behavior of the error probability as a function of the rate and codeword length. Hence, by determining Gallager’s exponent, we can obtain significant insight into the reliability–rate tradeoff in communication over MIMO channels and the required codeword length to achieve a certain level of reliable communication. In particular, the diversity–multiplexing tradeoff of MIMO channels [27] is a special case of the reliability–rate tradeoff as the SNR goes to infinity.

A. Random Coding Exponent

Using the formulation developed in [3, ch. 7], we obtain the random coding bound on the error probability of ML decoding over MIMO block-fading channels as

$$P_e \leq \left( \frac{2 \sigma^2}{\xi} \right)^{2 - \frac{N_0 N_c}{r}}$$

(9)

where $r, \delta \geq 0$ and

$$\xi \approx \frac{2 \sigma^2}{\sqrt{2 \pi N_0 \sigma^2}}$$

(10)

$$\sigma^2 = \int_{X} \left( \text{tr} (XX^\dagger) - n_c \mathbb{P} \right)^2 p_X(X) dX.$$  

(11)

The random coding exponent $E_r(p_X(X), R, N_c)$ in (9) is given by

$$E_r(p_X(X), R, N_c) = \max_{0 \leq p \leq 1} \left\{ \max_{r \geq 0} E_0 (p_X(X), \rho, r, N_c) - \rho R \right\}$$

(12)

with $E_0 (p_X(X), \rho, r, N_c)$ in (13) shown at the bottom of the page. The parameter $r$ to be optimized may be viewed as a Lagrange multiplier corresponding to the input power constraint [7].

1) Capacity-Achieving Input Distribution: In general, optimization of the input distribution $p_X(X)$ to maximize the error exponent (i.e., to minimize the upper bound) is a difficult task. As in [3]–[8], we choose the capacity-achieving distribution for $p_X(X)$ satisfying the power constraint (3),

$$P_X (X) = \frac{1}{N_c} \ln \left\{ \int_H p_H(H) \int_Y \left( \int_X p_X (X) e^{[\text{tr} (XX^\dagger) - N_c \mathbb{P}] p (Y|X,H)^{1/(1+p)}} dX \right)^{1+p} dY dH \right\}$$

(13)

where $X = (X_{ij})$ is an $m \times n$ matrix of complex variables that do not depend functionally on each other.

$$dX = \prod_{i=1}^m \prod_{j=1}^n d\Re X_{ij} d\Im X_{ij}.$$  

If $X \in \mathbb{C}^{m \times n}$ is Hermitian, then

$$dX = \prod_{i=1}^m dX_{ii} \prod_{i<j}^m d\Re X_{ij} d\Im X_{ij}.$$
namely,

\[ p_X(X) = \pi^{-nT N_c} \det(Q)^{-N_c} \text{etr} \left( -Q^{-1}XX^\dagger \right) \]  

(14)

with \( \text{tr}(Q) \leq P \). Although this choice of the Gaussian input distribution for the error exponent calculation is optimal only if the rate \( R \) approaches the channel capacity, it makes the problem analytically tractable [3].

**Proposition 1:** Let \( E_{0,G}(Q, \rho, r, N_c) \) be \( E_0(p_X, \rho, r, N_c) \) in (13) for the Gaussian input distribution \( p_X(X) \) of (14). Then, we have

\[
E_{0,G}(Q, \rho, r, N_c) = rP (1 + \rho) + (1 + \rho) \ln \det \left( I_{nT} - rQ \right) - \frac{1}{N_c} \ln \mathbb{E} \left\{ \det \left( I_{nR} + \frac{H(Q^{-1} - rI_{nT})^{-1} H^\dagger}{N_0 (1 + \rho)} \right)^{-N_c \rho} \right\}.
\]

(15)

**Proof:** See Appendix A.

For the case of equal power allocation to each transmit antenna, i.e., \( Q = \frac{P}{nT} I_{nT} \) (because the transmitter has no channel knowledge), (15) becomes

\[
E_{0,G} \left( \frac{P}{nT} I_{nT}, \rho, r, N_c \right) = rP (1 + \rho) + nT (1 + \rho) \ln \left( \frac{nT - rP}{nT} \right) - \frac{1}{N_c} \ln \mathbb{E} \left\{ \det \left( I_{nR} + \frac{(nT - rP)}{(nT - rP) (1 + \rho)} \right)^{-N_c \rho} \right\}
\]

(16)

where \( \gamma = P/N_0 \) is the average SNR at each receive antenna. Let us introduce a new variable \( \beta = nT - rP \) where \( \beta \) is restricted to the range \( 0 \leq \beta \leq nT \) to have a meaningful result in (16). Then, we have

\[
\hat{E}_0 (\rho, \beta, N_c) \triangleq E_{0,G} \left( \frac{P}{nT} I_{nT}, \rho, r, N_c \right)|_{\beta = nT - rP} = \left( 1 + \rho \right) (nT - \beta) + nT (1 + \rho) \ln \left( \beta/nT \right) - \frac{1}{N_c} \ln \mathcal{L}_0 (\rho, \beta, N_c)
\]

(17)

where

\[
\mathcal{L}_0 (\rho, \beta, N_c) = \mathbb{E} \left\{ \det \left( I_m + \frac{\gamma \Theta}{\beta (1 + \rho)} \right)^{-N_c \rho} \right\}.
\]

(18)

With maximization over \( \beta \) in \([0, nT]\) and \( \rho \) in \([0, 1]\) to obtain the tightest bound, we have the random coding exponent for Gaussian codebooks and equal power allocation as follows:

\[
E_t (R, N_c) \Delta \equiv E_t (p_X(X), R, N_c) \bigg|_{X \sim N_{nT, N_c} \left( 0, \frac{P}{nT} I_{nT}, I_{nT} \right)} = \max_{0 \leq \rho \leq 1} \left\{ \max_{0 \leq \beta \leq nT} \hat{E}_0 (\rho, \beta, N_c) - \rho R \right\}.
\]

(19)

**Proposition 2:** Let \( \beta^* (\rho) \) be the value of \( \beta \) that maximizes \( \hat{E}_0 (\rho, \beta, N_c) \) defined in (17) for each \( \rho \in [0, 1] \). Then, \( \beta^* (\rho) \) is the solution of \( \partial \hat{E}_0 (\rho, \beta, N_c)/\partial \beta = 0 \) and is always in the range \( 0 \leq \beta \leq nT \).

**Proof:** See Appendix B.

It can be shown using (65) and (70) in Appendix B that as \( \gamma \to \infty \) or \( \gamma \to 0 \), the optimal value of \( \beta \) does not depend on \( N_c \), that is,

\[
\lim_{\gamma \to \infty} \beta^* (\rho) = nT - \frac{m_p}{1 + \rho}, \quad \lim_{\gamma \to 0} \beta^* (\rho) = nT.
\]

(20)

According to Proposition 2 and using the general relation \( dE_t (R, N_c)/dR = -\rho \), the maximization of the exponent in (19) over \( \beta \in [0, nT] \) and \( \rho \in [0, 1] \) can be performed by the following parametric equations:

\[
E_t (R, N_c) = \hat{E}_0 (\rho, \beta^* (\rho), N_c) - \rho R
\]

(21)

with

\[
\frac{\partial \hat{E}_0 (\rho, \beta, N_c)}{\partial \rho} = (nT - \beta) + nT \ln \left( \beta/nT \right) \frac{\partial K^0 (\rho, \beta)}{\partial \rho} - \frac{1}{N_c} \mathcal{L}_0^{-1} (\rho, \beta, N_c) \frac{\partial \mathcal{L}_0 (\rho, \beta, N_c)}{\partial \rho}
\]

(22)

where

\[
\frac{\partial \mathcal{L}_0 (\rho, \beta, N_c)}{\partial \rho} = \mathbb{E} \left\{ N_c \det (\frac{1}{\beta} \Omega_{\rho, \beta})^{-N_c \rho} \right\}
\]

(23)

\[
\frac{\partial K^0 (\rho, \beta)}{\partial \rho} = \frac{\rho \gamma}{\left( 1 + \rho \right)^2} \left\{ \Theta (\frac{1}{\beta} \Omega_{\rho, \beta})^{-1} - \ln \det \left( \frac{1}{\beta} \Omega_{\rho, \beta} \right) \right\}.
\]

\[
\mathcal{L}_0 (\rho, \beta, N_c) = \mathbb{E} \left\{ \det \left( I_m + \frac{\gamma \Theta}{\beta (1 + \rho)} \right)^{-N_c \rho} \right\},
\]

(18)

\[
\mathcal{L}_0 (\rho, \beta, N_c) = \mathbb{E} \left\{ \det \left( I_m + \frac{\gamma \Theta}{\beta (1 + \rho)} \right)^{-N_c \rho} \right\}.
\]

(24)

\[
\mathcal{L}_0 (\rho, \beta, N_c) = \mathbb{E} \left\{ \det \left( I_m + \frac{\gamma \Theta}{\beta (1 + \rho)} \right)^{-N_c \rho} \right\}.
\]

(25)

We remark that the capacity expression (25) obtained from the exponent is independent of the channel coherence time \( N_c \) and is in agreement with the previous result [14]–[16]. Also, the quantity \( E_t \) is defined as the value of the exponent \( E_t (R, N_c) \) at \( R = 0 \), referred to as the exponential error-bound parameter [4], [5], and is given by \( \hat{E}_0 (1, \beta^* (1), N_c) \). This quantity is
equal to the value of \( R \) at which the exponent becomes zero by setting \( \rho = 1 \) and \( \beta = \beta^* (1) \). If setting \( r = 0 \) or equivalently \( \beta = n_T \) (i.e., without the constraint on the minimum energy of the codewords) in (13), \( E_0 \) becomes equal to the cutoff rate \( R_0 \) of the channel

\[
R_0 = \tilde{E}_0(1, n_T, N_c)
\]

(26)

This is an important parameter, as it determines both the magnitude of the zero-rate exponent and the rate regime in which the error probability can be made arbitrarily small by increasing the codeword length.

3) Effect of Channel Coherence—Channel-Incurable Effect: Using Jensen’s inequality, it is easy to show

\[
\frac{1}{N_c} \ln \mathcal{L}_0 (\rho, \beta, N_c) \geq \frac{1}{N_c - 1} \ln \mathcal{L}_0 (\rho, \beta, N_c - 1)
\]

(28)

yielding

\[
\tilde{E}_0 (\rho, \beta, N_c) \leq \tilde{E}_0 (\rho, \beta, N_c - 1).
\]

(29)

Therefore, for fixed \( R \), the random coding exponent decreases with \( N_c \), while the channel capacity is independent of \( N_c \). This reliability reduction is due to the fact that the increase in \( N_c \) results in a decrease in the number of independent channel realizations across the code and hence, reduces the effectiveness of channel coding to mitigate unfavorable fading. We call this effect of the channel coherence time on communication reliability “a channel-incurable effect”. In particular, since \( \lim_{N_c \to \infty} \frac{1}{N_c} \ln \mathcal{L}_0 (\rho, \beta, N_c) = 0 \), we have

\[
\lim_{N_c \to \infty} \tilde{E}_0 (\rho, \beta, N_c) = \mathcal{K} (\rho, \beta)
\]

(30)

leading to \( \lim_{N_c \to \infty} \beta^* (\rho) = n_T \) and \( \lim_{N_c \to \infty} E_t (R, N_c) = 0 \). Therefore, if \( N_c \to \infty \), it is impossible to transmit information at any positive rate with arbitrary reliability even with the use of multiple antennas. In fact, \( n_T \) must also increase without limit so that the so-called space–time autocoding effect takes place, which makes arbitrarily reliable communications possible [11].

B. Evaluation of \( \tilde{E}_0 (\rho, \beta, N_c) \), \( \partial \tilde{E}_0 (\rho, \beta, N_c) / \partial \beta \), and \( \partial \tilde{E}_0 (\rho, \beta, N_c) / \partial \rho \)

To calculate the random coding exponent, the quantities \( E_0 (\rho, \beta, N_c) \), \( \partial E_0 (\rho, \beta, N_c) / \partial \beta \), and \( \partial E_0 (\rho, \beta, N_c) / \partial \rho \) need to be determined. We now evaluate them in the following theorem which will be proven in the next section (see Table I for some quantities and matrices involved in this theorem).

**Theorem 1:** Let \( \mathbf{H} \sim \mathcal{N}_{n_{HR}, n_T} (0, \Phi_R, \Phi_T) \) or \( \Theta \sim \mathcal{Q}_{m,n} (I_m, \Phi_1, \Phi_2) \). Then,

1) \( \tilde{E}_0 (\rho, \beta, N_c) \) is given by (31) shown at the bottom of the page. If \( \Phi_T = I_{n_T} \) and \( \Phi_R = I_{n_R} \) (i.i.d. MIMO channel), then (31) reduces to

\[
\tilde{E}_0 (\rho, \beta, N_c) = \mathcal{K} (\rho, \beta) - \frac{1}{N_c} \ln \left[ K^{-1}_\text{id} \det \mathbf{Y}_\text{id} (\rho, \beta) \right].
\]

(32)

2) \( \partial \tilde{E}_0 (\rho, \beta, N_c) / \partial \beta \) is given by

\[
\frac{\partial \tilde{E}_0 (\rho, \beta, N_c)}{\partial \beta} = \mathcal{K} (\rho, \beta) - \frac{T_A}{N_c \beta} - \frac{1}{N_c} \operatorname{tr} \left[ \mathbf{G}_{(n-m)} (\Phi_{2})^{-1} \mathbf{Y}^\beta (\rho, \beta) \right].
\]

(33)

3) \( \partial \tilde{E}_0 (\rho, \beta, N_c) / \partial \rho \) is given by (35) shown at the bottom of the next page. If \( \Phi_T = I_{n_T} \) and \( \Phi_R = I_{n_R} \), then

\[
\frac{\partial \tilde{E}_0 (\rho, \beta, N_c)}{\partial \rho} = \mathcal{K} (\rho, \beta) - \frac{1}{N_c} \operatorname{tr} \left[ \mathbf{Y}_\text{id}^\beta (\rho, \beta) \mathbf{Y}_\text{id} (\rho, \beta) \right].
\]

(34)

**Corollary 1 (Ergodic Capacity):** Let

\[
\mathbf{H} \sim \mathcal{N}_{n_{HR}, n_T} (0, \Phi_R, \Phi_T).
\]

Then, the ergodic capacity \( \langle C \rangle \) is given by

\[
\langle C \rangle = \mathbf{tr} \left[ \begin{bmatrix} \mathbf{G}_{(n-m)} (\Phi_{2})^{-1} & 0 \\ \mathbf{Y} (0, n_T) & \mathbf{A} \end{bmatrix} \right] - (m - 1) + \sum_{i=1}^{\phi (\Phi_1)} \sum_{j=1}^{\chi (\Phi_1)} \left( \frac{j}{m - \chi_i (\Phi_1)} + j \right)
\]

(37)
with $\Lambda \in \mathbb{R}^{m \times n}$ given by

$$
\Lambda = \begin{bmatrix}
\Lambda_{1,1} & \cdots & \Lambda_{1,\rho(\Phi_2)} \\
\vdots & \ddots & \vdots \\
\Lambda_{\rho(\Phi_1),1} & \cdots & \Lambda_{\rho(\Phi_1),\rho(\Phi_2)}
\end{bmatrix}
$$

(38)

where the $(i,j)$-th entry $\Lambda_{p,q,ij}$ of $\Lambda_{p,q} \in \mathbb{R}^{\chi_p(\Phi_1) \times \chi_q(\Phi_2)}$, $p = 1, 2, \ldots, g(\Phi_1)$, $q = 1, 2, \ldots, g(\Phi_2)$, is

$$
\Lambda_{p,q,ij} = G_{i+1,2} \left( \frac{\gamma}{n_T} \lambda_{(p)}(\Phi_1), \lambda_{(q)}(\Phi_2), m - i + 1 \right).
$$

(39)

Proof: Note that

$$
\begin{align*}
\text{tr} \left\{ \left[ G_{(n-m)}(\Phi_2) \right]^{-1} \left[ \Upsilon_{(\rho)}(0,n_T) \right] \right\} \\
= -N_c \text{tr} \left\{ \left[ G_{(n-m)}(\Phi_2) \right]^{-1} \left[ 0 \Lambda \right] \right\} \\
+ n_T \text{tr} \left\{ \left[ G_{(n-m)}(\Phi_2) \right]^{-1} \left[ \Upsilon(\beta)(0,n_T) \right] \right\}
\end{align*}
$$

(40)

The proof follows immediately from (24), Theorem 1.3 with $\rho = 0$ and $\beta = n_T$, and (40).

Note that the expression (37) for the ergodic capacity $\langle C \rangle$ is sufficiently general and applicable to arbitrary structures of correlation matrices $\Phi_T$ and $\Phi_R$, and hence, embraces all the previously known results for i.i.d. channels ($\Phi_T = \mathbf{I}_{n_T}$, $\Phi_R = \mathbf{I}_{n_R}$) [16, 22], one-sided correlated channels ($\Phi_1 = \mathbf{I}_{m}$ [20] or $\Phi_2 = \mathbf{I}_{n}$ [21]), and doubly correlated channels [23] (where all the eigenvalues of $\Phi_T$ and $\Phi_R$ are assumed to be distinct) as special cases of (37).

Corollary 2 (Cutoff Rate): If $\mathbf{H} \sim \mathcal{N}_{n_R,n_T}(0, \Phi_R, \Phi_T)$, then the cutoff rate $R_0$ is given by (41) shown at the bottom of the page. In particular, if $\Phi_T = \mathbf{I}_{n_T}$ and $\Phi_R = \mathbf{I}_{n_R}$, then we have

$$
R_0 = -\frac{1}{N_c} \ln \left\{ K_{\text{cor}}^{-1} \det \left[ \Upsilon_{\text{id}}(1, n_T) \right] \right\}
$$

(42)

Proof: It follows immediately from (26) and Theorem 1.1 with $\rho = 1$ and $\beta = n_T$.

C. Coding Requirement

As in [6], we can approximate the required codeword length to achieve a prescribed error probability $P_e$ at a rate $R$ by solving for $N_b$ in the following equation:

$$
P_e = \left( \frac{2e^{r\delta}}{\xi} \right)^{2} e^{-N_b N_c E_e(R,N_c)}.
$$

(43)

Using (10), it is easy to see that the factor $(2e^{r\delta}/\xi)^2$ in (43) is minimized over $\delta \geq 0$, for large $N_b$, by choosing $\delta = 1/r$ [3]. This yields

$$
\min_{\delta \geq 0} \left( \frac{2e^{r\delta}}{\xi} \right)^{2} \approx 8\pi e^2 \sigma_\xi^2 r^2 N_b \quad \text{for large } N_b.
$$

(44)

Also, from (11) and [24, Lemma 5], we have

$$
\sigma_\xi^2 = \frac{N_c P^2}{n_T}.
$$

(45)

Combining (44) and (45) together with the fact that $\beta = n_T - rP$, (43) can be written as

$$
P_e = \frac{8\pi}{n_T} [n_T - \beta^* (\rho)]^2 N_b N_c e^{-N_b N_c E_e(R,N_c) + 2}.
$$

(46)

After solving for $N_b$ in (46), we take $L = N_c \cdot [N_b]$ as our estimate of the codeword length (in symbol) required to achieve $P_e$ at the rate $R$, where $[\cdot]$ denotes the smallest integer larger than or equal an enclosed quantity.

IV. PROOF OF THE MAIN THEOREM

In this section, we provide proofs of the main results stated in Theorem 1. The methodology recently developed in [23]
TABLE I
SOME QUANTITIES AND MATRICES INVOLVED IN THEOREM 1

<table>
<thead>
<tr>
<th>( G_{(m)}(\Psi) )</th>
<th>( \Theta_{(m)}(\Phi_1) )</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \mathbf{A}_k )</td>
<td>( \mathbf{\bar{A}}_k )</td>
</tr>
</tbody>
</table>

where \( A_k = (A_{k,i,j}) \in \mathbb{R}^{r \times k}(\Phi), \ k = 1, 2, \ldots, g(\Phi) \)

\( \mathbf{\bar{A}}_k = (\mathbf{\bar{A}}_{k,i,j}) \in \mathbb{R}^{u \times k}(\Phi), \ k = 1, 2, \ldots, g(\Phi) \)

\( K_{\text{id}} = \det (\Phi_1)^{N_{c}\rho} \det \{ G_{(m)}(\Phi_1) \} \prod_{k=1}^{N_{c}\rho} (k-1)! \), \( N_{c}\rho \in \{1, 2, \ldots, m\} \)

\( K_{\text{ad}} = \frac{m}{2} m (m+1) - \frac{1}{2} \sum_{i=1}^{g(\Phi_1)} \sum_{j=1}^{g(\Phi_1)} \chi_i(\Phi_1) [\chi_i(\Phi_1) + 1] \)

\( \Xi(\rho, N_{c}) = \det (\Phi_2)^{m-1} \det \{ \Theta_{(m)}(\Phi_1) \}^{-1} \det \{ G_{(m)}(\Phi_2) \}^{-1} \prod_{k=1}^{N_{c}\rho} (N_{c}\rho - m + 1)_{k-1} \)

The \( G_{\kappa,\nu}(a, b, \mu) \) is defined as the integral

\( G_{\kappa,\nu}(a, b, \mu) = \int_0^{\infty} (1 + ax)^{\nu-1}(1 + ax)^{\nu-1} - \kappa e^{-x/b} dx \), \( a, b > 0 \), \( \kappa, \nu \in \mathbb{N} \), \( \mu \in \mathbb{C} \)

where \( \rho \Phi_q(t) \) is the \((p, q)\)-th characteristic coefficient of \( \Phi_2 \) (see for details [24, Definition 6]).

The integral identity can be found in [23, Appendix A].

and [24] for dealing with random matrices paves a way to prove the theorem.

A. Proof of Theorem 1.1

Using the same steps leading to [23, Theorem 1], we get

\[ \mathcal{L}_0(\rho, \beta, N_c) = \int_{\Theta_{\Theta_0}} \det (I_{m} + \eta(\Theta))^{-N_{c}\rho} \rho_{\Theta}(\Theta) d\Theta \]

For \( A \in \mathbb{C}^{p \times p} > 0 \), we denote

\[ \int_{\lambda(A)} d\lambda(A) = \int_{0 < \lambda_p(A) \leq \cdots \leq \lambda_1(A) < \infty} d\lambda_1(A) \cdots d\lambda_p(A). \]

If the integrand is symmetric in \( \lambda_1(A), \lambda_2(A), \ldots, \lambda_p(A) \), then

\[ \int_{\lambda(A)} d\lambda(A) = \frac{1}{p!} \int_0^\infty \cdots \int_0^\infty d\lambda_1(A)d\lambda_2(A) \cdots d\lambda_p(A). \]

and [24] for dealing with random matrices paves a way to prove the theorem.

A. Proof of Theorem 1.1

Using the same steps leading to [23, Theorem 1], we get

\[ \mathcal{L}_0(\rho, \beta, N_c) = \int_{\Theta_{\Theta_0}} \det (I_{m} + \eta(\Theta))^{-N_{c}\rho} \rho_{\Theta}(\Theta) d\Theta \]

where \( \eta = \frac{1}{\beta (1 + \rho)} \) and \( D = \text{diag} (\lambda_1(\Theta), \ldots, \lambda_m(\Theta)) \).

Successively applying the generic determinantal formula for hypergeometric functions of matrix arguments [24, Lemma 4] and the generalized Cauchy–Binet formula [23, Lemma 2], the integral in (47) can be evaluated, after some algebra, as

\[ \mathcal{L}_0(\rho, \beta, N_c) = \eta^{-T_0} T_0(\rho, N_c) \det \left[ G_{(m-\rho)\rho}(\Phi_2) \right], \]

where \( N_c \rho \neq 1, 2, \ldots, m - 1 \). (48)
SHIN and WIN: GALLAGER’S EXPONENT FOR MIMO CHANNELS: A RELIABILITY-RATE TRADEOFF

TABLE 1
(Continued.) SOME QUANTITIES AND MATRICES INVOLVED IN THEOREM 1

In Theorem 1

\[
Y_p(q) = (\gamma / q)_{i,j} (\rho, \beta) = \begin{bmatrix}
Y_{1,1} (\rho, \beta) & \cdots & Y_{1,q} (\rho, \beta) \\
\vdots & \ddots & \vdots \\
Y_{p,q} (\rho, \beta) & \cdots & Y_{p,q} (\rho, \beta)
\end{bmatrix} \in \mathbb{R}^{m \times n}
\]

where
\[
Y_{p,q} (\rho, \beta) = (\gamma / q)_{i,j} (\rho, \beta), \quad p = 1, \ldots, g (\Phi_1), \quad q = 1, \ldots, g (\Phi_2)
\]

(i, j)th element: \(Y_{p,q} (\rho, \beta) = G_{i+j-1,1} \left( \gamma / q (\rho, \beta), \lambda (\Phi_1), \lambda (\Phi_2), -N_c \rho + m - i + 1 \right)\).

\[
\mathcal{Y}^{(3)} (\rho, \beta) = \frac{\partial}{\partial \rho} Y(\rho, \beta) = \begin{bmatrix}
\mathcal{Y}_{1,1}^{(3)} (\rho, \beta) & \cdots & \mathcal{Y}_{1,q}^{(3)} (\rho, \beta) \\
\vdots & \ddots & \vdots \\
\mathcal{Y}_{p,q}^{(3)} (\rho, \beta) & \cdots & \mathcal{Y}_{p,q}^{(3)} (\rho, \beta)
\end{bmatrix}
\]

where
\[
\mathcal{Y}_{p,q}^{(3)} (\rho, \beta) = (\partial / \partial \rho)_{p,q,ij} (\rho, \beta), \quad p = 1, \ldots, g (\Phi_1), \quad q = 1, \ldots, g (\Phi_2)
\]

(i, j)th element: \(\mathcal{Y}_{p,q}^{(3)} (\rho, \beta) = \frac{\partial}{\partial \rho} G_{i+j-1,1} \left( \gamma / q (\rho, \beta), \lambda (\Phi_1), \lambda (\Phi_2), -N_c \rho + m - i + 1 \right)\).

Substituting (48) into (17) gives the second case of (31). It should be noted that the formula in the second case of (31) has singular points at \(N_c \rho = 1, 2, \ldots, m - 1\) for each \(p \in \{0, 1\}\).

These singularities stem from the quantity \(T_B (\rho, N_c)\), which can be alleviated using the following analysis.

Suppose that \(N_c \rho\) is a positive integer. Then, using [24, Lemma 1], we have

\[
\mathcal{L}_0 (\rho, \beta, N_c) = \mathbb{E}_\Theta \left\{ \mathbb{E}_S \left\{ \exp (-\eta \Theta S^\dagger S) \right\} \right\}
\]

\[
= \mathbb{E}_S \left\{ \det (I_{m,n} + \eta S^\dagger \Phi_1 \otimes \Phi_2)^{-1} \right\}
\]

\[
= \mathbb{E}_S \left\{ \det (I_{m,n} + \eta S^\dagger \Phi_1 S \otimes \Phi_2)^{-1} \right\}
\]

(49)

where \(S \sim \mathcal{N}_{m,n,\rho} (0, I_m, I_{N_c \rho})\) is a complex Gaussian matrix statistically independent of \(\Theta\), and the last equality follows from the fact that \(S S^\dagger \Phi_1\) and \(S^\dagger \Phi_1 S\) have the same nonzero eigenvalues.

If \(N_c \rho \in \{1, 2, \ldots, m\}\), then

\[
Z = S^\dagger \Phi_1 S \sim \mathcal{N}_{N_c \rho, m} (I_m, I_{N_c \rho}, \Phi_1).
\]

Hence, using [24, Theorem 9], (49) for the case of \(N_c \rho \in \{1, 2, \ldots, m\}\) can be written as

\[
\mathcal{L}_0 (\rho, \beta, N_c) = \mathbb{E}_\Theta \left\{ \prod_{k=1}^{N_c \rho} \det \left\{ I_{m,n} + \eta \lambda_k (Z) \Phi_2 \right\}^{-1} \right\}
\]

\[
= K_{-1} \int_{\lambda (Z)} \prod_{k=1}^{N_c \rho} \det \left\{ I_{m,n} + \eta \lambda_k (Z) \Phi_2 \right\}^{-1}
\]

\[
\times \det \left\{ G(m-N_c \rho, \Theta_1) \right\}
\]

\[
\times \det \left\{ G(N_c \rho, \Phi_2) \right\}
\]

(50)

where \(\hat{\Xi} = [\hat{\xi}_1, \hat{\xi}_2, \ldots, \hat{\xi}_{g(\Phi_1)}]\) and the \((i, j)\)-th entry \(\hat{\xi}_{k,ij}\) of \(\hat{\Xi}_k \in \mathbb{R}^{N_c \rho \times N_c \rho}\), \(k = 1, 2, \ldots, g(\Phi_1)\), is given by

\[
\hat{\xi}_{k,ij} = \lambda_k^{-1} (Z) e^{-\lambda_k(Z)/(\lambda_k(k) \Phi_1)}.
\]

(51)
Now, applying [23, Lemma 2] to (50) yields
\[ L_0(\rho, \beta, N_c) = K_{\text{co}}^{-1} \det \left( \begin{bmatrix} \mathbf{G}_{(m-N_c,\rho)}(\Phi_1) \end{bmatrix} \right) \] (52)
where the \((i, j)\)-th entry \(\Xi_{k,ij}(\rho, \beta)\) of the \(k\)-th constituent matrix \(\Xi_k(\rho, \beta)\) is given by
\[ \Xi_{k,ij}(\rho, \beta) = \int_0^{\infty} \det \left( \mathbf{I}_n + \eta z \Phi_2 \right)^{-1} z^{i+j-2} e^{-z/\lambda(k_i)(\Phi_1)} \, dz. \] (53)

Using the characteristic coefficients [24, Definition 6], (53) can be written as
\[ \Xi_{k,ij}(\rho, \beta) = \sum_{p=1}^{5} \sum_{q=1}^{5} \left\{ \chi_{p,q}(\Phi_2) \right. \times \int_0^{\infty} \left( 1 + \eta \lambda(p) \langle \Phi_2 \rangle z \right)^{-q} z^{i+j-2} e^{-z/\lambda(k_i)(\Phi_1)} \, dz \left. \right\} \] (54)
where \(\chi_{p,q}(\Phi_2)\) is the \((p, q)\)-th characteristic coefficient of \(\Phi_2\). Finally, substituting (52) into (17) gives the first case of (31) and hence, we complete the proof of the first part.

**B. Proofs of Theorem 1.2 and 1.3**

The second and third parts can be obtained by differentiating \(E_0(\rho, \beta, N_c)\) in Theorem 1.1 with respect to \(\beta\) and \(\rho\), respectively, with the help of the logarithmic derivative of a determinant [30, Theorem 9.4] (or more generally [23, Lemma 1]).

**V. NUMERICAL RESULTS AND DISCUSSION**

In this section, we provide some numerical results to illustrate the reliability–rate tradeoff in MIMO block-fading channels. For spatial fading correlation, we consider an exponential correlation model with \(\Phi_T = (\zeta_T^{i-j})\) and \(\Phi_R = (\zeta_R^{i-j})\), \(\zeta_T, \zeta_R \in [0, 1]\), in all examples.

To ascertain the effect of the channel coherence on the error exponent, Figs. 2 and 3, respectively, show the random coding exponent \(E_r(R, N_c)\) as a function of a rate \(R\) for i.i.d. (\(\zeta_T = 0, \zeta_R = 0\)) and exponentially correlated (\(\zeta_T = 0.5, \zeta_R = 0.7\)) MIMO channels at \(\gamma = 15\) dB, where \(n_T = n_R = 3\) and \(N_c = 1, 2, 3, 4, 5, 6, 7, 8, 9, 10\). We see from the figures that the exponent at a rate \(R\) below the ergodic capacity decreases with \(N_c\), while the ergodic capacity remains constant for all \(N_c\) (i.e., 8.48 and 7.19 nats/symbol for Figs. 2 and 3, respectively). For example, the error exponents at rates \(R \leq R_{\text{er}}\) for \(N_c = 10\) are reduced by roughly 3.46 and 2.86 for i.i.d. and exponentially correlated cases, respectively, compared with those for \(N_c = 1\). This reduction in the exponent, consequently, requires using a longer code to achieve the same error probability. Hence, we see that unlike the capacity (with perfect receive CSI), the channel coherence time plays a fundamental role in the error exponent or the reliability of communications.

Fig. 4 demonstrates the effect of spatial fading correlation on the random coding exponent, where \(\zeta_T = \zeta_R = \zeta, \gamma = 15\) dB, \(n_T = n_R = 3, N_c = 5\), and \(\zeta\) ranges from 0 (i.i.d.) to 0.9. As seen from the figure, there exists a remarkable reduction in the exponent at the same rate due to correlation, especially for \(\zeta \geq 0.5\). The amount of reduction in the exponent at rates \(R \leq R_{\text{er}}\), relative to the i.i.d. MIMO exponent, ranges from 0.07 for \(\zeta = 0.2\) to 2.17 for \(\zeta = 0.9\), indicating that a longer code is required to achieve the same level of reliable communications. Equivalently, a decrease in the information rate is required for more correlated channels to achieve the same value of the exponent. For example, the exponent at a rate 3 nats/symbol are 1.94 and 1.53 for the i.i.d. and correlated (\(\zeta_T = \zeta_R = 0.5\)) channels, respectively. This implies that 27% increase in the codeword length, due to spatial fading correlation, is required to achieve a rate 3 nats/symbol with the same communication reliability.

To get more insight into the influences of the number of antennas, channel coherence time, and fading correlation on a
coding requirement for MIMO channels, the codeword length required to achieve $P_e \leq 10^{-6}$ at a rate 8.0 bits/symbol (5.55 nats/symbol) are investigated in Tables II–IV. The codeword lengths in the tables are calculated in such a manner as described in Section III-C. Table II serves to demonstrate the effect of increasing the number of antennas on the coding requirement, in which the required codeword length $L$ is shown for i.i.d. MIMO channels with $N_c = 5$. Note that it is impossible to reliably communicate at a rate 8.0 bits/symbol below the SNR $\gamma$ of 14.55 dB, 9.68 dB, and 6.79 dB for $n_T = n_R = 2$, 3, and 4, respectively, since these SNR’s are required to attain the ergodic capacity $\langle C \rangle$ of 8.0 bits/symbol in each of the cases. As seen from the table, with increasing the number of antennas at both transmit and receive sides, the required codeword lengths are remarkably reduced. This is due to the advantages of the use of multiple antennas, e.g., spatial multiplexing and diversity gains [22]. For example, at $\gamma = 16$ dB, increasing the number of antennas at both sides from 2 to 3 and 4 reduces the corresponding codeword length to almost 2.8% and 0.9% of the amount required for two transmit and receive antennas, respectively, which is a tremendous reduction in the codeword length.

Table III shows the required codeword length $L$ for i.i.d. and exponentially correlated ($\zeta_T = 0.5$, $\zeta_R = 0.7$) MIMO channels with $n_T = n_R = 3$ at $\gamma = 15$ dB when $N_c$ varies from 1 to 10. It is clear from Table III that for each value of $N_c$, the codeword lengths for correlated channels are much longer than those for i.i.d. channels. For example, the increase in the required codeword length, due to exponential correlation ($\zeta_T = 0.5$, $\zeta_R = 0.7$), ranges from 364% for $N_c = 1$ to 138% for $N_c = 10$, which is a significant increase in required codeword length. Also, when going $N_c$ from 1 to 10, there is a considerable increase in the required codeword length, relative to that for the single-symbol coherence time, which ranges from 33% to 344% for the i.i.d. case and from 28% to 258% for the correlated case, respectively.

Table IV demonstrates the effect of correlation on the required code length $L$, where $n_T = n_R = 3$, $\zeta_T = \zeta_R = \zeta$, $N_c = 5$, and $\gamma = 15$ dB. The table contains the corresponding codeword lengths for $\zeta$ from 0 to 0.9. As seen from the table, the required codeword length for the case of exponential correlation $\zeta = 0.7$ is equal to 4.5 times as long as for the i.i.d. channel ($\zeta = 0$). Particularly, when $\zeta \geq 0.5$, there exists a large amount of increase in required codeword length due to a stronger correlation. Also, since the ergodic capacity is 7.36 bits/symbol for $\zeta_T = \zeta_R = 0.9$ at $\gamma = 15$ dB, it is impossible to achieve reliable communications at a rate 8.0 bits/symbol (regardless of the codeword length), when $\zeta_T = \zeta_R = 0.9$.

Finally, Fig. 5 shows the cutoff rate $R_0$ in nats/symbol as a function of a correlation coefficient $\zeta$ for exponentially correlated MIMO channels with $\zeta_T = \zeta_R = \zeta$ at $\gamma = 15$ dB, where $n_T = n_R = 3$ and $N_c$ varies from 1 to 10. We see that the cutoff rate $R_0$ decreases with $N_c$ for all $\zeta \in [0, 1]$. While $\langle C \rangle$ remains constant, $R_0$ monotonically decreases with $N_c$, going to 0 as $N_c \to \infty$. Hence, these two measures diverge as $N_c$ increases and eventually $\lim_{N_c \to \infty} \langle C \rangle = \infty$, which coincides with the divergent behavior of the capacity and cutoff rate of a channel with block memory [9]. This observation reveals that $R_0$ is more pertinent than $\langle C \rangle$ as a figure of merit that reflects the quality of block-fading channels.

**VI. CONCLUSIONS**

In this paper, we derived Gallager’s random coding error exponent to investigate a fundamental tradeoff between the communication reliability and information rate in spatially correlated MIMO channels. We considered a block-fading channel with perfect receive CSI and Gaussian codebooks. The required codeword lengths for a prescribed error probability were calculated from the random coding bound to aid in the assessment of the coding requirement on such MIMO channels, taking into account the effects of the number of antennas, the channel coherence time, and the amount of spatial fading correlation. In addition, we obtained the general formulae for the ergodic capacity and cutoff rate, which encompass all the previous capacity results as special cases of our expressions. In parallel to the capacity–cutoff rate divergence in a block-memory channel, we observed the channel-incurable effect:
the monotonically decreasing property of the MIMO exponent (i.e., communication reliability) with the channel coherence time.

APPENDIX A
PROOF OF PROPOSITION 1

Lemma 1: Let \( S \sim N_{m,n}(M, \Sigma, I_n) \) and \( A \in \mathbb{C}^{m \times m} > 0 \) be Hermitian. Then, we have

\[
E \left\{ \text{etr}(-A S S^\dagger) \right\} = \frac{\det(\Sigma)^{-n}}{\pi^{mn}} \text{etr} \left\{ - (A^{-1} + \Sigma)^{-1} M M^\dagger \right\}
\]

\[
\times \int_S \text{etr} \left\{ - (A + \Sigma^{-1}) \left[ S - (I_m + \Sigma A)^{-1} M \right] \left[ S - (I_m + \Sigma A)^{-1} M \right]^\dagger \right\} dS
\]

(58)

\[
\int_X p_X(X) e^{\left[ \text{tr}(X X^\dagger) - N_c p \right]} p(Y|X,H)^{1/(1+p)} dX = e^{-r N_c (\pi N_0)^{-n N_c/(1+p)} \det(I_{n_T} - r Q)^{-N_c}}
\]

\[
\times \det \left( I_{n_R} + \frac{H (Q^{-1} - r I_{n_T})^{-1} H^\dagger}{N_0 (1 + \rho)} \right)^{-N_c} \left[ \text{etr} \left\{ - \frac{1}{N_0 (1 + \rho)} \left( I_{n_R} + \frac{H (Q^{-1} - r I_{n_T})^{-1} H^\dagger}{N_0 (1 + \rho)} \right) Y Y^\dagger \right\} \right]^{-1}
\]

(59)

\[
\int_Y \left\{ \int_X p_X(X) e^{\left[ \text{tr}(X X^\dagger) - N_c p \right]} p(Y|X,H)^{1/(1+p)} dX \right\}^{1+p} dY
\]

\[
e^{-r N_c p (1+p)} \det(I_{n_T} - r Q)^{-N_c (1+p)} \det \left( I_{n_R} + \frac{H (Q^{-1} - r I_{n_T})^{-1} H^\dagger}{N_0 (1 + \rho)} \right)^{-N_c p}
\]

(60)

**TABLE III**
REQUIRED CODEWORD LENGTH \( L \) AS A FUNCTION OF CHANNEL COHERENCE TIME \( N_c \) FOR I.I.D. AND EXPONENTIALLY CORRELATED MIMO CHANNELS AT A RATE 8.0 bits/symbol for \( P_e \leq 10^{-6} \), \( n_T = n_R = 3 \), AND \( \gamma = 15 \) dB

<table>
<thead>
<tr>
<th>Coherence time ( N_c )</th>
<th>Codeword length ( L )</th>
</tr>
</thead>
<tbody>
<tr>
<td>i.i.d. ( \xi_T = 0.5 ), ( \xi_R = 0.7 )</td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>18</td>
</tr>
<tr>
<td>2</td>
<td>24</td>
</tr>
<tr>
<td>3</td>
<td>30</td>
</tr>
<tr>
<td>4</td>
<td>36</td>
</tr>
<tr>
<td>5</td>
<td>45</td>
</tr>
<tr>
<td>6</td>
<td>48</td>
</tr>
<tr>
<td>7</td>
<td>56</td>
</tr>
<tr>
<td>8</td>
<td>64</td>
</tr>
<tr>
<td>9</td>
<td>72</td>
</tr>
<tr>
<td>10</td>
<td>80</td>
</tr>
</tbody>
</table>

**TABLE IV**
REQUIRED CODEWORD LENGTH \( L \) AS A FUNCTION OF CORRELATION COEFFICIENT \( \xi \) FOR EXPONENTIALLY CORRELATED MIMO CHANNELS WITH \( \xi_T = \xi_R = \xi \) AT A RATE 8.0 bits/symbol for \( P_e \leq 10^{-6} \), \( n_T = n_R = 3 \), \( N_c = 5 \), AND \( \gamma = 15 \) dB

<table>
<thead>
<tr>
<th>Correlation coefficient ( \xi )</th>
<th>Codeword length ( L )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.0</td>
<td>45</td>
</tr>
<tr>
<td>0.1</td>
<td>45</td>
</tr>
<tr>
<td>0.2</td>
<td>45</td>
</tr>
<tr>
<td>0.3</td>
<td>50</td>
</tr>
<tr>
<td>0.4</td>
<td>60</td>
</tr>
<tr>
<td>0.5</td>
<td>75</td>
</tr>
<tr>
<td>0.6</td>
<td>105</td>
</tr>
<tr>
<td>0.7</td>
<td>200</td>
</tr>
<tr>
<td>0.8</td>
<td>1015</td>
</tr>
<tr>
<td>0.9</td>
<td>-</td>
</tr>
</tbody>
</table>

*Note: For \( \xi_T = \xi_R = 0.9 \), the ergodic capacity \( C \) is 7.36 bits/symbol at \( \gamma = 15 \) dB.*

**Proof:** Note that

\[
E \left\{ \text{etr}(-A S S^\dagger) \right\} = \frac{\det(\Sigma)^{-n}}{\pi^{mn}} \text{etr} \left\{ - (A^{-1} + \Sigma)^{-1} M M^\dagger \right\}
\]

\[
\times \int_S \text{etr} \left\{ - A S S^\dagger - \Sigma^{-1} (S - M) (S - M)^\dagger \right\} dS
\]

(56)

By writing the trace of the quadratic form in the exponent of (56) as

\[
\text{tr} \left\{ A S S^\dagger + \Sigma^{-1} (S - M) (S - M)^\dagger \right\}
\]

\[
= \text{tr} \left\{ (A + \Sigma^{-1} - I_m + \Sigma A)^{-1} M \right\}
\]
It is easy to show that

\[
\frac{\partial \det (\mathbf{\Omega}_{p,\beta})}{\partial \beta} = \det (\mathbf{\Omega}_{p,\beta}) \text{tr} (\mathbf{\Omega}_{p,\beta}^{-1}) \tag{63}
\]

and hence,

\[
\frac{\partial^2 \mathcal{L}_1 (\rho, \beta, N_c)}{\partial \beta^2} = \mathbb{E} \left\{ \mathcal{H} \left( \mathbf{\Omega}_{p,\beta}^{-1} \right) \right\} \tag{66}
\]

Let us now define the random variables

\[
X^2 = \det (\mathbf{\Omega}_{p,\beta})^{-N_c \rho} \tag{67}
\]

\[
Y^2 = (N_c \rho)^2 \det (\mathbf{\Omega}_{p,\beta})^{-N_c \rho} \text{tr} (\mathbf{\Omega}_{p,\beta}^{-1}) \tag{68}
\]

From Schwarz’s inequality, we have

\[
\left( \frac{\partial \mathcal{L}_1 (\rho, \beta, N_c)}{\partial \beta} \right)^2 = \mathbb{E} \{XY\}
\]

\[
\leq \mathbb{E} \{X^2\} \cdot \mathbb{E} \{Y^2\}
\]

\[
\leq \mathbb{E} \{X^2\} \cdot \mathbb{E} \left\{ Y^2 + N_c \rho X^2 \text{tr} (\mathbf{\Omega}_{p,\beta}^{-2}) \right\}
\]

\[
= \mathcal{L}_1 (\rho, \beta, N_c) \frac{\partial^2 \mathcal{L}_1 (\rho, \beta, N_c)}{\partial \beta^2} \tag{69}
\]

From (69), we see that \( \tilde{E}_0 (\rho, \beta, N_c) \) is a concave function of \( \beta \) for all \( \rho \in [0, 1] \). Hence, the maximum over \( \beta \) occurs at \( \beta^* (\rho) \) for which \( \left| \frac{\partial \tilde{E}_0 (\rho, \beta, N_c) / \partial \beta}{\beta^* (\rho)} \right|_{\beta = \beta^* (\rho)} = 0 \) and it is sufficient to show that \( \left| \frac{\partial \tilde{E}_0 (\rho, \beta, N_c) / \partial \beta}{\beta^* (\rho)} \right|_{\beta = \beta^* (\rho)} \geq 0 \) and \( \left| \frac{\partial \tilde{E}_0 (\rho, \beta, N_c) / \partial \beta}{\beta^* (\rho)} \right|_{\beta = \beta^* (\rho)} \leq 0 \) for all \( \rho \in [0, 1] \) in order to prove \( 0 < \beta^* (\rho) \leq n_T \). Since

\[
\frac{\partial \tilde{E}_0 (\rho, \beta, N_c)}{\partial \beta} = \left( 1 + \rho \right) \left( n_T - \beta \right) - \frac{m \rho \ln (\beta)}{N_c} - \frac{1}{N_c} \ln \mathcal{L}_1 (\rho, \beta, N_c)
\]

\[
\triangleq \mathcal{K}^{(\rho)} (\rho, \beta) = \mathcal{K}^{(\rho)} (\rho, \beta) \tag{60}
\]

\[
= \left. \frac{\mathcal{L}_1 (\rho, \beta, N_c) \frac{\partial \mathcal{L}_1 (\rho, \beta, N_c)}{\partial \beta}}{\partial \beta^2} \right|_{\beta = n_T^{-1}} \tag{70}
\]

it is clear that \( \lim_{\beta \to 0} \tilde{E}_0 (\rho, \beta, N_c) \to \infty \). Also,

\[
\left. \left| \frac{\partial \tilde{E}_0 (\rho, \beta, N_c) / \partial \beta}{\beta^* (\rho)} \right|_{\beta = n_T^{-1}} \right|_{\beta = n_T^{-1}} \tag{71}
\]

Since \( \mathcal{L}_1 (\rho, \beta, N_c) \geq 0 \) and

\[
\mathcal{L}_1 (\rho, n_T, N_c) = \mathbb{E} \left\{ \det (\mathbf{\Omega}_{\rho, n_T})^{-N_c \rho} \right\}
\]
Thus, proof of the proposition.

\[ E \left\{ \det (\Omega_{\rho,n})^{-N_\rho} \frac{\text{tr} (\Omega^{-1}_{\rho,n})}{\text{tr} (\Omega_{\rho,n})} \right\} \geq \frac{m}{m^*} E \left\{ \det (\Omega_{\rho,n})^{-N_\rho} \frac{\text{tr} (\Omega^{-1}_{\rho,n})}{\text{tr} (\Omega_{\rho,n})} \right\} \]

\[ = \frac{-m \rho}{m^* N} \cdot \frac{1}{N \epsilon} \left[ \left. \frac{\partial \mathcal{L}_1 (\rho, \beta, N)}{\partial \beta} \right|_{\beta=m^* N} \right] \leq 0. \]

Thus, \( \left[ \frac{\partial \mathcal{E}_0 (\rho, \beta, N)}{\partial \beta} \right|_{\beta=m^* N} \leq 0 \) and we complete the proof of the proposition.

**ACKNOWLEDGEMENT**

The authors wish to thank E. Abbe, R. G. Gallager, and S. M. Moser for their comments and careful reading of the manuscript.

**REFERENCES**


Hyundong Shin (S’01–M’04) received the B.S. degree in Electronics Engineering from Kyung Hee University, Korea, in 1999, and the M.S. and Ph.D. degrees in Electrical Engineering from Seoul National University, Seoul, Korea, in 2001 and 2004, respectively.

From September 2004 to February 2006, Dr. Shin was a Postdoctoral Associate at the Laboratory for Information and Decision Systems (LIDS), Massachusetts Institute of Technology (MIT), Cambridge, MA, USA. In March 2006, he joined the faculty of the School of Electronics and Information, Kyung Hee University, Korea, where he is now an Assistant Professor at the Department of Electronics and Radio Engineering. His research interests include wireless communications, information and coding theory, cooperative/collaborative communications, and multiple-antenna wireless communication systems and networks.

Professor Shin served as a member of the Technical Program Committee in the IEEE International Conference on Communications (2006), the IEEE International Conference on Ultra Wideband (2006), the IEEE International Conference on Communications (2009), the IEEE Global Communications Conference (2009), and the IEEE Vehicular Technology Conference (2009, Fall). He served as a Technical Program co-chair for the IEEE Wireless Communications and Networking Conference PHY Track (2009). Dr. Shin is currently an Editor for the IEEE TRANSACTIONS ON WIRELESS COMMUNICATIONS. He was a Guest Editor for the 2008 EURASIP Journal on Advances in Signal Processing (Special Issue on Wireless Cooperative Networks).

Professor Shin received the IEEE Communications Society’s Guglielmo Marconi Best Paper Award (2008) and the IEEE Vehicular Technology Conference Best Paper Award (2008 Spring).
Moe Z. Win (S’85-M’87-SM’97-F’04) received both the Ph.D. in Electrical Engineering and the M.S. in Applied Mathematics as a Presidential Fellow at the University of Southern California (USC) in 1998. He received an M.S. in Electrical Engineering from USC in 1989, and a B.S. (magna cum laude) in Electrical Engineering from Texas A&M University in 1987.

Dr. Win is an Associate Professor at the Massachusetts Institute of Technology (MIT). Prior to joining MIT, he was at AT&T Research Laboratories for five years and at the Jet Propulsion Laboratory for seven years. His research encompasses developing fundamental theory, designing algorithms, and conducting experimentation for a broad range of real-world problems. His current research topics include location-aware networks, time-varying channels, multiple antenna systems, ultra-wide bandwidth systems, optical transmission systems, and space communications systems.

Professor Win is an IEEE Distinguished Lecturer and was elected Fellow of the IEEE, cited for “contributions to wideband wireless transmission.” He was honored with the IEEE Eric E. Sumner Award (2006), an IEEE Technical Field Award for “pioneering contributions to ultra-wide band communications science and technology.” Together with students and colleagues, his papers have received several awards including the IEEE Communications Society’s Guglielmo Marconi Best Paper Award (2008) and the IEEE Antennas and Propagation Society’s Sergei A. Schelkunoff Transactions Prize Paper Award (2003). His other recognitions include the Laurea Honoris Causa from the University of Ferrara, Italy (2008), the Technical Recognition Award of the IEEE ComSoc Radio Communications Committee (2008), Wireless Educator of the Year Award (2007), the Fulbright Foundation Senior Scholar Lecturing and Research Fellowship (2004), the U.S. Presidential Early Career Award for Scientists and Engineers (2004), the AIAA Young Aerospace Engineer of the Year (2004), and the Office of Naval Research Young Investigator Award (2003).

Professor Win has been actively involved in organizing and chairing a number of international conferences. He served as the Technical Program Chair for the IEEE Wireless Communications and Networking Conference in 2009, the IEEE Conference on Ultra Wideband in 2006, the IEEE Communication Theory Symposia of ICC-2004 and Globecom-2000, and the IEEE Conference on Ultra Wideband Systems and Technologies in 2002; Technical Program Vice-Chair for the IEEE International Conference on Communications in 2002; and the Tutorial Chair for ICC-2009 and the IEEE Semiannual International Vehicular Technology Conference in Fall 2001. He was the chair (2004-2006) and secretary (2002-2004) for the Radio Communications Committee of the IEEE Communications Society. Dr. Win is currently an Editor for the IEEE TRANSACTIONS ON WIRELESS COMMUNICATIONS. He served as Area Editor for Modulation and Signal Design (2003-2006), Editor for Wideband Wireless and Diversity (2003-2006), and Editor for Equalization and Diversity (1998-2003), all for the IEEE TRANSACTIONS ON COMMUNICATIONS. He was Guest-Editor for the PROCEEDINGS OF THE IEEE (Special Issue on UWB Technology & Emerging Applications) in 2009 and the IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS (Special Issue on Ultra-Wideband Radio in Multiaccess Wireless Communications) in 2002.