Performance Analysis for Generalised Spatial Modulation

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Abstract—In this paper, the performance of generalised spatial modulation (GSM) using large scale multiple–input multiple–output (MIMO) systems is analysed from an information–theoretic point of view. The analysis is extended to account for the average bit error ratio (ABER). We introduce a simple model for the capacity which allows us to derive a closed–form expression for the capacity of GSM. This is achieved by modelling GSM as two independent sources of information, where we show that GSM has the potential to achieve the capacity of MIMO systems. Moreover, using Monte–Carlo simulation for link–level mutual information, it is shown that the theoretical limit can be reached with practical signal constellations in operationally important signal–to–noise–ratio (SNR) regions. Furthermore, the ABER performance of GSM is analysed in this paper, and a tight upper bound for correlated and uncorrelated, Rayleigh and Rician channels is derived. Finally, the performance of GSM is validated through Monte Carlo simulations and compared with the performance of spatial modulation (SM). It is shown that GSM outperforms SM by up to 3.5 dB while the same low complexity of SM is retained. Specifically, the complexity is equal to that of single–input multiple–output (SIMO) systems. Moreover, a further advantage of SM is maintained which is that only a single radio frequency (RF) chain is required. Thus, GSM is a perfect candidate for large scale MIMO systems.

Index Terms—Generalised Spatial Modulation (GSM), Spatial modulation (SM), multiple–input multiple–output (MIMO), large scale MIMO.

I. INTRODUCTION

Spatial modulation (SM) is a transmission technology proposed for multiple–input multiple–output (MIMO) wireless systems. It aims to increase the modulation rate, \( (r) \), of single–antenna systems while avoiding inter–channel interference (ICI) [1, 2]. This is attained through the adoption of a new modulation and coding scheme, which foresees: i) the activation, at each time instance, of a single antenna that transmits a given data symbol (constellation symbol), and ii) the exploitation of the spatial position (index) of the active antenna as an additional dimension for data transmission (spatial symbol) [3]. An overall increase by the base–two logarithm of the number of transmit–antennas of the modulation rate is achieved. The computational complexity of SM is equal to the complexity of single–antenna systems [4]. This is an important property in the context of large–scale MIMO systems.

In [5], generalised spatial modulation (GSM) was proposed to increase the number of bits sent over the spatial domain. This is achieved by transmitting the same symbol simultaneously from more than one transmit antenna. Transmitting the same data symbol from the active antennas retains the key advantages of SM, in particular:

• Complete avoidance of ICI at the receiver as all active antennas transmit the same symbol;

• Low computational complexity. It will be shown later that the complexity is equal to that of single–input multiple–output (SIMO) systems;

• Only one radio frequency (RF) chain is required. This is because all active antennas transmit the same symbol.

A maximum–likelihood (ML) receiver is considered, where an exhaustive search is performed over all possible antenna combinations and signal constellation points.

The analytical average bit error ratio (ABER) performance of GSM is studied in [5], and an upper bound is proposed. However, the bound is applicable only for uncorrelated Rayleigh fading channels. Motivated by this, in this paper we derive a general closed–form upper bound for the ABER performance of GSM in correlated and uncorrelated channel conditions. The proposed bound is applicable to generalised fading channels such as Rayleigh, Rice, and Nakagami–m channels.

There is a common perception that at least as many independent data streams as degrees of freedom (DoF) have to be transmitted into the channel in order to achieve maximum capacity, where from [6] the DoF is equal to the minimum between the number of transmit antennas and the number of receive antennas. In this paper we pursue this idea further and show that GSM has the potential to achieve maximum capacity with fewer number of independent data streams. This is approached by an information–theoretic capacity analysis of GSM. Unlike traditional MIMO links, the wireless channel itself carries information. Therefore, capacity analysis tends to be more challenging. In the literature the generally adopted approach towards the capacity of SM systems is to analyse separately the information capacity conveyed in the antenna index and signal constellation [7]. However, in this paper we use a different approach in which we model the system as two independent sources of information, \( i.e., \) signal constellation and antenna index, with multiplicative interaction. This model allows us to calculate the capacity of GSM in a more straightforward manner. Moreover, it enables us to derive a close–form solution which can be used for further optimization. Note, to the best of the authors’ knowledge the close–form expression for the capacity of GSM has not been derived in literature.

The remainder of this paper is organised as follows. In Section II, GSM, the channel model, and the ML receiver are introduced. In Section III, the capacity analysis for GSM is described. In Section IV, the generalised closed–form expression for the ABER performance of GSM over correlated and uncorrelated channels is derived. Finally, the results are presented in Section V, and the paper is concluded in Section VI.
II. SYSTEM MODEL

A. GSM Modulator

In GSM more than one transmit antenna is active and all active antennas send the same complex symbol. Hence, a set of antenna combinations can be formed, and used as spatial constellation points. The number of possible antenna combinations is \( \binom{N_t}{N_u} \), where \( N_t \) is the number of transmit antennas, \( N_u \) is the number of active antennas at each channel use, and \( \binom{}{} \) denotes the binomial operation. However, the number of antenna combinations that can be considered for transmission must be a power of two. Therefore, only \( 2^n \) combinations, can be used, where \( n = \left\lfloor \log_2 \binom{N_t}{N_u} \right\rfloor \), and \( \left\lfloor \cdot \right\rfloor \) is the floor operation.

In general, the maximum number of bits that can be transmitted using GSM is given by,

\[
\eta = \eta_t + \eta_s = \left\lfloor \log_2 \binom{N_t}{N_u} \right\rfloor + \log_2 M \tag{1}
\]

where \( M \) is the size of the signal constellation diagram

The GSM mapper divides the incoming bit stream into blocks of \( \eta_t \) bits. First, the initial \( \eta_t \) bits are used to select the set of active antennas. In this paper, the set of active transmit–antennas are denoted by \( \Upsilon_{\ell_t} \), with \( \Upsilon_{\ell_t} \in \{ \Upsilon_1, \Upsilon_2, \ldots, \Upsilon_{\ell_t} \} \). Second, the last \( \eta_s \) bits choose a symbol in the signal–constellation diagram. Without loss of generality, quadrature amplitude modulation (QAM) is considered in this paper. The transmitted complex symbol is denoted by \( s_t \), with \( s_t \in \{ s_1, s_2, \ldots, s_M \} \).

B. Channel Model

The transmitted vector, \((x_{\ell_t,s_t})\), is transmitted over a flat fading \( N_r \times N_t \) MIMO channel with a transfer function \( \hat{H} \), where \( N_r \) is the number of receive antennas. In this paper, perfect channel state information (CSI) at the receiver is assumed. Knowledge of CSI at the transmitter is not assumed. Thus, the \( N_r \times 1 \) dimensional receive vector can be written as follows,

\[
y = \hat{H}x_{\ell_t,s_t} + n = \hat{h}^{\ell_t} s_t + n \tag{2}
\]

where \( n \) is the \( N_r \)-dimensional additive white Gaussian noise (AWGN) with zero–mean and variance \( \sigma_n^2 \) per dimension at the receiver input, and the vector \( \hat{h}^{\ell_t} = \sum_{n \in \Upsilon_{\ell_t}} h_n \) contains the summation of the active antennas channel vectors, and \( h_{\ell_n} \) is the \( n \)-th column of the \( \hat{H} \). Note, the SNR \( E_s/N_0 = 1/\sigma_n^2 \), where \( E_s = E\left\{ \| \hat{H}x_{\ell_t,s_t} \|^2 \right\} = 1 \).

In this paper, frequency–flat fading MIMO channels are considered. In particular: non line–of–sight (NLOS) channels, line–of–sight (LOS) channels, and spatial correlation (SC) are considered in the simulation.

1) Non Line–of–Sight (Rayleigh Fading): The entries of \( \hat{H} \) are modelled as complex identical and independently distributed (i.i.d) Gaussian random variables with zero–mean and unit–variance.

2) Line–of–Sight (Rician Fading): The standard statistical model for a multipath fading channel with a LOS component follows a Rician distribution. Thus, the channel impulse response is modelled as,

\[
\hat{H} = \sqrt{\frac{K}{1+K}} + \sqrt{\frac{1}{1+K}} \mathbf{H}' \tag{3}
\]

where \( K \) is the Rician factor, \( K/(1+K) \) is the mean power of the LOS component, \( 1/(1+K) \) is the mean power of the random component, and \( \mathbf{H}' \) is a \( N_r \times N_t \) matrix whose entries are modelled as complex i.i.d. Gaussian random variables with zero–mean and unit–variance.

3) Spatial Correlation Model (SC): The correlation channel matrix is modelled using the Kronecker channel model for its straightforward mathematical description [8],

\[
\mathbf{H} = \mathbf{R}_r^{\frac{1}{2}} \mathbf{H}_\mathbf{R}^{\frac{1}{2}} \mathbf{R}_t^{\frac{1}{2}} \tag{4}
\]

where \( \mathbf{H} \) is the uncorrelated channel, which can be NLOS or LOS channels, \( \mathbf{R}_r^{\frac{1}{2}} \) is the transmitter correlation matrix, and \( \mathbf{R}_r^{\frac{1}{2}} \) is the receiver correlation matrix.

Moreover, the correlation matrices can be generated using an exponential decay model [9],

\[
\mathbf{R}_c = \begin{bmatrix}
1 & r_c & r_c^2 & \cdots & r_c^{n-1} \\
r_c & 1 & r_c & \cdots & r_c^{2} \\
r_c^2 & r_c & 1 & \cdots & r_c \\
\vdots & \vdots & \vdots & \ddots & \vdots \\
r_c^{n-1} & r_c^2 & r_c^1 & \cdots & 1
\end{bmatrix} \tag{5}
\]

where \( r_c = \exp(-\beta) \), \( \beta \) is the correlation decay coefficient, and \( n \) is the number of transmit or receive antennas.

C. ML–Optimum Detector

The ML–optimum detector for GSM can be written as follows,

\[
\hat{\ell}_t, \hat{s}_t = \arg \min_{s \in M_{QAM}} \{ \| y - \hat{h}^\ell s \|^2_F \} \tag{6}
\]

where \( \| \cdot \|_F \) is the Frobenius norm, and \( \hat{\ell}_t, \hat{s}_t \) denotes the estimated used antenna combination and symbol respectively.

From (6), the number of real multiplicative operations needed by GSM–ML receiver is equal to,

\[
C_{ML} = 6N_r 2^n \tag{7}
\]

where the ML detector searches through the whole transmit and receive search spaces, and evaluating the Euclidean distance \( \| y - \hat{h}^\ell s \|^2 \) requires \( 6N_r \) real multiplications.

From (7), the complexity of GSM does not depend on the number of transmit antennas, and it is equal to the complexity of SM and SIMO systems.
III. CAPACITY ANALYSIS

In this section, we study the performance of GSM in an information-theoretic sense. In particular, we study the capacity of GSM in a fading channel. GSM in many ways, poses some interesting opportunities for system designers. For instance, unlike in SM, the number of active transmit antennas could be varied according to the error rate and capacity requirements. In following sections, we show how these differences in configuration may be manifested in error and capacity performances. We assume that the set of active antennas (i.e., $\mathcal{T}_t$) is fixed and known. We first consider the case of $N_r = 1$ where the channel transfer function in (2) reduces to a scalar equation:

$$y = h_t^s x_t + n$$

where all the vector variables have now reduced to scalar variables. In this analysis, the constellation symbol, $s_t$ is assumed to be continuous with finite second order moment (i.e., $E_s/N_0$) to be in line with classical standard single-input-single-output (SISO) capacity analysis, but the fact that $h_t^s$ is discrete and finite has to be considered. First, we model this channel as two source (or user) multiplicative multiple access channel (MAC). The sources, $h_t^s$ and $s_t$ are independent. We further assume the source $h_t^s$ to have a finite alphabet given by $h_1^s, ..., h^s$ and $s_t$. We are interested in finding the sum capacity of the system. The mutual information is by definition, $I(s_t, h_t^s; y)$. From the chain rule for mutual information of MAC

$$I(s_t, h_t^s; y) = H(y) - H(y|s_t, h_t^s)$$

It can be shown that $H(y|s_t, h_t^s) = \log_2 p\sigma_2^2$ in AWGN channels. Since, $I(s_t, h_t^s; y)$ is maximum when $H(y)$ is maximum, we are left with maximizing the entropy of $y$. $H(y)$ can be upper bounded by the entropy of a complex Gaussian random variable with the same variance. Therefore,

$$I(s_t, h_t^s; y) \leq \log_2 (\pi e\sigma^2) - \log_2 p\sigma_2^2$$

$$= \log_2 \left( 1 + \frac{\sigma^2}{\sigma_n^2} \right)$$

where $\sigma^2 = E\{|z|^2\}$, $z = h_t^s x_t$, and $E\{\cdot\}$ is the expectation operator. It can be shown that $\sigma^2 = E\{|h_t^s|^2\} E_s/N_0$. Finding the upper bound now is equivalent to finding the second moment of the source $h_t^s$. Due to the fact that $h_t^s$ has a finite alphabet, it is clear that its constellation energy is given by $\sum_{t,s=1} E_{\text{total}}$. Note that this is due to the fact that in the absence of CSI at the transmitter, the active antenna sets are used equally likely. Let the channel vector $H = (h_1^s, ..., h^s)$. Then the capacity bound in a more compact form becomes:

$$I(s_t, h_t^s; y) \leq \log_2 \left( 1 + \frac{E_s}{N_0 \sigma_n^2} \|H\|^2 \right).$$

Furthermore, we will show using simulations that this capacity upper bound is indeed achievable even with practical signal constellations (i.e., QAM, etc.). Next, we extend our arguments to the case of $N_r > 1$ where the upper bound in (10) becomes:

$$I(s_t, H_t^s; y) \leq \log_2 \left| I + \frac{1}{\sigma_n^2} Q_s \right|$$

where $Q_s$ is the covariance matrix of the vector $z = h_t^s x_t$. From the independence of $s_t$ and $h_t^s$, we obtain

$$Q_s = E_h y^s \left\{ h_t^s (h_t^s)^H \right\} E_s/N_0,$$

where from the law of large number,

$$E_h y^s \left\{ h_t^s (h_t^s)^H \right\} = \frac{1}{\theta} \sum_{t,s} h_t^s (h_t^s)^H.$$

In order to be able to interpret the result for the covariance of $z$, we consider an experiment where one observes $H_t^s$ for $\tau$ number of trials. Then, the sample average is taken. If $\tau$ is sufficiently large, the sample covariance converges to the distribution covariance. Due to the fact that $h_t^s$ are used equally likely by the transmitter, for a large but finite number of trials of $\tau$, the observer should get $H_t^s$, $\tau/p$ times. Therefore, the sample covariance converges to $\frac{1}{\theta} \sum_{t,s} h_t^s (h_t^s)^H$. Let $H_{\{\}} = \{h_1^s, ..., h^s\}$. Substituting the new covariance result from (14) into (12), we obtain the final results as

$$I(s_t, H_t^s; y) \leq \log_2 \left| I + \frac{E_s}{N_0 \sigma_n^2} H_{\{\}} H_{\{\}}^H \right|$$

Due to the fact that upper bound is achievable, we claim that the upper bound is indeed the capacity. Then the ergodic capacity in fading becomes

$$C = E_h \left\{ \log_2 \left| I + \frac{E_s}{N_0 \sigma_n^2} H_{\{\}} H_{\{\}}^H \right| \right\}.$$

IV. ANALYTICAL MODELLING FOR THE ABER OF GSM

The ABER for the GSM–ML system can be approximated by using the union bound [10], which can be expressed as follows,

$$\text{ABER} \leq \frac{1}{2\eta} \sum_{s_t, s} \sum_{x_{t,s}, x_{t,s}} \frac{N}{\eta} = \frac{1}{\eta} E_{\text{H}} \left\{ \frac{P_t}{\text{error}} \right\}$$

where $N(x_{t,s}, x_{t,s})$ is the number of bits in error between $x_{t,s}$, and $x_{t,s}$. $E_{\text{H}} \{\cdot\}$ is the expectation across the channel $H$, and $P_t$ is the pairwise error probability (PEP) of deciding on $x_{t,s}$ given that $x_{t,s}$ is transmitted,

$$\text{Pr}_{\text{error}} = \text{Pr} \left( \left| \mathbf{y} - \mathbf{H} x_{t,s} \right| > \left| \mathbf{y} - \mathbf{H} x_{t,s} \right| \right)$$

$$= Q \left( \frac{\|\mathbf{H} \mathbf{y} - \mathbf{H} \mathbf{H} \mathbf{y} \|}{\|\mathbf{H} \mathbf{y} \|} = \frac{1}{\sqrt{4\sigma_n^2}} \exp \left( -\frac{\|\mathbf{H} \mathbf{y} \|^2}{4\sigma_n^2 \sin^2 \theta} \right) \right)$$

where $\Psi = (x_{t,s} - x_{t,s})$, and from [11], the alternative
\[ E_H \{ P_{r_{\text{error}}} \} = \frac{1}{\pi} \int_0^{\pi} \exp \left( -\frac{1}{2 \sin^2 \theta} \left( \mathbf{u}_H^H \mathbf{A} \left( \mathbf{I}_{N_r, N_t} + \frac{1}{4 \sigma_n^2} \mathbf{L}_{\tilde{H}} \mathbf{A} \right)^{-1} \mathbf{u}_{\tilde{H}} \right) \right) d\theta \]

\[ \leq \frac{1}{2} \exp \left( -\frac{1}{2 \sigma_n^2} \left( \mathbf{u}_H^H \mathbf{A} \left( \mathbf{I}_{N_r, N_t} + \frac{1}{4 \sigma_n^2} \mathbf{L}_{\tilde{H}} \mathbf{A} \right)^{-1} \mathbf{u}_{\tilde{H}} \right) \right) \]

\[ \left( \text{25} \right) \]

The integral expression of the \( Q \)-function is,

\[ Q(x) = \frac{1}{\pi} \int_0^{\pi} \exp \left( \frac{x^2}{2 \sin^2 \theta} \right) d\theta \]

\[ \text{(20)} \]

Taking the expectation of (19),

\[ E_H \{ P_{r_{\text{error}}} \} = \frac{1}{\pi} \int_0^{\pi} \Phi \left( \frac{x}{\sigma^2_n} \right) d\theta \]

\[ \text{(21)} \]

where \( \Phi (\cdot) \) is the moment-generation function (MGF) of the random variable \( \| \mathbf{H} \mathbf{\Psi} \|^2 \).

The argument of the MGF in (21) can be written as,

\[ \| \tilde{\mathbf{H}} \mathbf{\Psi} \|^2 = \text{tr} \left( \tilde{\mathbf{H}} \mathbf{\Psi} \mathbf{H} \tilde{\mathbf{H}}^H \right) = \tilde{\mathbf{H}}^H \left( \mathbf{I}_{N_c} \otimes \mathbf{\Psi} \mathbf{H} \right) \tilde{\mathbf{H}} \]

\[ \text{(22)} \]

where \( \mathbf{I}_n \) is an \( n \times n \) identity matrix, \( (\cdot)^H \) denotes the Hermitian, and,

\[ \tilde{\mathbf{H}} = \mathbf{R}_s \odot \text{vec} (\mathbf{H}^H) \]

\[ \text{(23)} \]

where \text{vec}(\mathbf{H}) is the vectorisation operator where the columns of the matrix \( \mathbf{H} \) are stacked in a column vector, and \( \mathbf{R}_s = \mathbf{R}_{Rx} \otimes \mathbf{R}_{Tx} \) with \( \odot \) being the Kronecker product.

Now from (12) and (22), the MGF in (21) is,

\[ \Phi(s) = \exp \left( s \mathbf{u}_H^H \mathbf{A} \left( \mathbf{I}_{N_r, N_t} - s \mathbf{L}_{\tilde{H}} \mathbf{A} \right)^{-1} \mathbf{u}_{\tilde{H}} \right) \left( \mathbf{I}_{N_r, N_t} - s \mathbf{L}_{\tilde{H}} \mathbf{A} \right) \]

\[ \text{(24)} \]

where \( \mathbf{A} = \mathbf{I}_{N_r} \otimes \mathbf{\Psi} \mathbf{H} \), \( \mathbf{u}_{\tilde{H}} = \mathbf{u}_H \mathbf{R}_s^H \odot \text{vec}(\mathbf{1}_{N_c, N_i}) \), and \( \mathbf{L}_{\tilde{H}} = \sigma_n^2 \mathbf{R}_s \), where \( \mathbf{1}_n \) is an \( n \times n \) all ones matrix, and for,

1) Rayleigh Fading:

\[ u_H = 0 \]

\[ \sigma_H^2 = 1 \]

\[ \text{(26)} \]

\[ \text{(27)} \]

2) Rician Fading:

\[ u_H = \sqrt{\frac{K}{1 + K}} \]

\[ \sigma_H^2 = \frac{1}{1 + K} \]

\[ \text{(28)} \]

\[ \text{(29)} \]

From (19) and (24), the general PEP of SM is as given in (25) shown at the top of this page. The PEP in (25) is applicable for any fading channel with equal mean and equal variance for the real and imaginary parts, such as Rayleigh, Rice, and Nakagami–m channels.

In the next section by comparing to Monte Carlo simulations, the bound is shown to be a tight upper bound for GSM.

V. RESULTS

In the following, Monte Carlo simulations for at least \( \text{10}^6 \) channel realisations are conducted and the ABER and capacity performance of GSM are compared with SM. Note, the correlation decay coefficients are chosen to model moderate correlation, with \( \beta = 0.7 \) at the transmitter side and \( \beta = 0.6 \) at the receiver side [13].

A. Analytical Performance of GSM

The ABER simulation results for GSM with \( N_u = 3 \) over correlated and uncorrelated, Rayleigh channels, and Rician channels with \( K = 10 \text{ dB} \), are depicted in Figs. 1 and 2. These results are compared with the analytical results obtained from the bound in Section IV, where \( \eta = 8 \), \( N_t = 12 \), and \( N_r = 3 \). Moreover, binary phase shift keying (BPSK) is used. Analytical and simulation results for GSM demonstrate close match for wide and pragmatic range of signal-to-noise-ratio (SNR) values and for different channel conditions, which validates the derived analysis in this paper.
B. Capacity results

We present here our simulation results pertaining to capacity performance of GSM systems in two representative diagrams which are given in Figs. 3 and 4. Fig. 3 shows the capacity and simulated mutual information performance for GSM systems in an uncorrelated Rayleigh fading while Fig. 4 is for an uncorrelated Rician fading channel. The GSM capacity varies with the selection of possible antenna sets as we pointed out in Section III. For instance, in GSM for $\eta = 12$ and $N_u = 3$, we select $2^{15}$ antenna sets while for $\eta = 20$, we choose $2^{15}$ antenna sets. It is clear from Fig. 3 that, theoretical capacity curves for both GSM and SM do not exhibit much differences in Rayleigh fading environments. Furthermore, both simulated capacity curves for $\eta = 12$ achieve comparable mutual information but can only follow theoretical capacity until 5 dB. However, the GSM system with $\eta = 20$ closely follows the theoretical capacity until SNR = 15 dB. Therefore, this confirms that GSM can achieve the theoretical capacity limit of underlying MIMO systems in the absence of channel state information at the transmitter (CSIT) which is in fact equivalent to the capacity of SM (blue curve) in Fig. 3.

Next in Fig. 4, the ergodic capacity for Rician fading in general is considerably lower than in the Rayleigh fading case, this is due to the degenerative effect for the randomness of the channel caused by the LOS component of the Rician fading channel. Furthermore, capacity curves exhibit considerable variations with the varying configurations of the GSM systems. This can also be theoretically proved by upper bounding the ergodic capacity expression in (16) using Jenson inequality and evaluating the expectation inside the determinant. The GSM with $\eta = 20$ has the highest theoretical capacity and SM which is a special case of GSM with $N_u = 1$ has the lowest capacity in the SNR region considered. Notably, it appears that all mutual information curves struggle to follow theoretical capacity beyond SNR = 5 dB. This is because, due to the LOS effect, mutual information transferred in the antenna index reduces drastically. Therefore, the mutual information curves exhibit poor performance in Rician fading when there are a large number of bits encoded in the antenna sets. For instance, the GSM system with $\eta = 12$ where 10 bits are modulated in the antenna index can only follow theoretical capacity curve until $-5$ dB. The GSM system with $\eta = 20$ follows theoretical capacity curves until SNR = 15 dB in a Rayleigh fading channel. However, the GSM system follows theoretical capacity until 5 dB in a Rician fading channel. This observation is due to the fact that mutual information achieved from modulating information in the antenna index reduces drastically in Rician fading. In general, GSM and SM systems can achieve higher mutual information in any fading channels when $\eta$ is increased. However, if the increase in $\eta$ is achieved through increasing the number of bits encoded in the antenna index, the capacity gains are higher in Rayleigh fading channels than in Rician fading channels.

Fig. 2. ABER versus the SNR for GSM over Rician fading channels with $K = 10$ dB for $\eta = 8$, $N_r = 12$, $N_u = 3$, and $N_e = 4$. (Marker) Analytical bound, (Solid line) Simulation.

![Fig. 2](image-url)

Fig. 3. Ergodic capacity and link-level simulated mutual information of $64 \times 3$ MIMO system with different configurations for SM, GSM in Rayleigh Fading.

![Fig. 3](image-url)

Fig. 4. Ergodic capacity and link-level simulated mutual information of $64 \times 3$ MIMO system with different configurations for SM, GSM in Rician fading channels with $K = 5$ dB.

![Fig. 4](image-url)
In this paper, the capacity of GSM along with the ABER performance is studied and compared with SM. Unlike traditional capacity analyses, the capacity of GSM is studied by modelling GSM as two independent sources of information, constellation symbol and spatial symbol. By doing so a closed-form expression for the capacity of GSM was derived. Using the derived expression it is shown that GSM has the potential to achieve the capacity of MIMO systems. Furthermore, a novel, simple and accurate analytical closed-form upper bound for the ABER of SM over Rayleigh, and Rician fading channels is proposed. The bound allows the calculation of the ABER of GSM in correlated and uncorrelated channels, and for small and large scale MIMO systems. The performance of GSM is also compared with the performance of SM, and it is shown that GSM, 1) offers up to 3.5 dB better performance than SM, 2) has the same complexity as SM and SIMO systems, and 3) uses only one RF chain. Thus, we can conclude that GSM is perfectly suitable for large scale MIMO systems.

### VI. Conclusion

In this section, we summarise our findings and discuss their implications. The main contributions of this paper are: 1) the closed-form expression for the capacity of GSM, 2) the ABER performance of GSM along with the SM, and 3) the capacity analysis of GSM. These results show that GSM is a promising candidate for future wireless communication systems.

### References


