GENERALISED SPATIAL MODULATION FOR LARGE–SCALE MIMO

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ABSTRACT

In this paper, the performance of generalised spatial modulation (GSM) and spatial modulation (SM) is studied assuming channel estimation errors (CSEs) and correlated Rayleigh and Rician fading channels. A new, simple, accurate and general analytical closed-form upper bound for the average bit error ratio (ABER) performance of both systems is derived. The analytical bound is shown to be applicable to correlated and uncorrelated channels, as well as to small and large scale multiple-input multiple-output (MIMO) systems. The results demonstrate that GSM is more suitable for large-scale MIMO systems than SM. The performance gain of GSM over SM is about 5 dB. The results also show that SM is very robust to CSEs. Specifically, the performance degradation of SM in the presence of CSEs are 0.7 dB and 0.3 dB for Rayleigh and Rician fading channels respectively. Lastly, the findings in this paper underpin the suitability of both GSM and SM for future large-scale MIMO systems.

***Index Terms*—** Generalised spatial modulation (GSM), spatial modulation (SM), multiple-input multiple-output (MIMO), large scale MIMO, channel estimation errors (CSEs).

1. INTRODUCTION

Spatial Modulation (SM) [1] is a hybrid multiple-input multiple-output (MIMO) and digital modulation technique that uses the multiple antennas at the transmitter in a unique way to achieve spatial multiplexing gains. The index of each transmit-antenna represents a spatial constellation point that is used to carry additional information bits. A spatial multiplexing gain of base-two logarithm of the overall number of transmit-antennas can be achieved. Also, activating single transmit-antenna at a time eliminates inter-channel interference (ICI), removes the need for inter-antenna synchronization requirements, reduces receiver complexity, and allows the use of a single radio frequency (RF) chain at the transmitter, which offers significant energy saving potential [2]. Furthermore, it is demonstrated that SM is more robust to channel imperfections, such as spatial channel correlation and channel estimation errors, as compared with other MIMO techniques. The main reason for this is that the probability of error in SM is not only determined by the actual channel realization, but also by the difference between channels associated with the different transmit antennas [3, and references therein].

A limitation of SM is that the data rate enhancement is proportional to the base-two logarithm of the number of transmit-antennas, and therefore, the number of antennas must be a power of two – although this limitation is mitigated by using fractional bit encoding techniques [4]. This is unlike other spatial multiplexing techniques where data rate increases linearly with the number of transmit-antennas and any number of transmit-antennas can be used. Another way to overcome this limitation is to consider generalised spatial modulation (GSM) as proposed in [5, 6], where a combination of transmit antennas is activated at each time instant to transmit a unique data symbol. It is shown that activating more than one antenna at a time increases the achievable spectral efficiency while maintaining all advantages of SM, including single RF chain transmission.

The performance of large-scale GSM and SM systems over spatially correlated Rayleigh and Rician fading channels in the presence of channel estimation errors (CSEs) at the receiver are addressed in this paper. A general expression for the pairwise error probability (PEP) of such systems is computed in closed-form without resorting to Monte Carlo simulations. In addition to being computationally complex, the latter only gives limited insight into the effects of different channel parameters on the overall system performance. Moreover, the derived PEP is used to calculate a tight upper bound on the average bit error ratio (ABER) of such systems. The results demonstrate that GSM outperforms SM for the same spectral efficiency and system configurations.

The remainder of this paper is organised as follows. In Section 2, GSM and SM system models along with the channel model are introduced. In Section 3, the generalised closed-form expression for the ABER performance is derived in the presence of CSEs and assuming correlated and uncorrelated fading channels. Finally, the results are presented in Section 4, and the paper is concluded in Section 5.

2. SYSTEM MODEL

2.1. Space Modulator

In space modulation, all active transmit antennas send the same complex symbol. Hence, a set of antenna combinations can be formed, and used as spatial constellation points. In GSM the number of active antennas N_u is constant. Therefore, the number of possible antenna combinations is $\binom{N_t}{N_u}$, where N_t is the number of transmit antennas, and $\binom{\cdot}{\cdot}$ denotes the binomial operation. However, the number of antenna combinations that can be considered for transmission must be a power of two. Therefore, only $\varrho = 2^{\eta_\ell^{\mathrm{GSM}}}$ combinations, can be used, where $\eta_\ell^{\mathrm{GSM}} = \lfloor \log_2 \binom{N_t}{N_u} \rfloor$, and $\lfloor \cdot \rfloor$ is the floor operation. Thus, the maximum number of bits that can be transmitted using GSM is given by,

$$\eta^{\mathrm{GSM}} = \eta_\ell^{\mathrm{GSM}} + \eta_s = \left\lfloor \log_2 \binom{N_t}{N_u} \right\rfloor + \log_2 M, \quad (1)$$

where M is the size of the signal constellation diagram.

In SM only one transmit antenna is active at any time instant, *i.e.*, $N_u = 1$. Hence, from (1) the maximum number of bits that can be transmitted in the spatial domain is $\eta_\ell^{\mathrm{SM}} = \log_2 N_t$. The space modulation mapper divides the incoming bitstream into blocks of $\eta = \eta_\ell + \eta_s$ bits. The first group of bits, η_ℓ, is used to select the set of active antennas. The set of active transmit–antennas is denoted by Υ_{ℓ_t}, with $\Upsilon_{\ell_t} \in \{\Upsilon_1, \Upsilon_2, \ldots, \Upsilon_\varrho\}$. The second group of bits, η_s, are used to map a data symbol signal from the signal–constellation diagram. Without loss of generality, M-quadrature amplitude modulation (QAM) is considered in this paper. The transmitted complex symbol is denoted by s_t, with $s_t \in \{s_1, s_2, \ldots, s_M\}$.

2.2. Channel Model

The spatially modulated $N_t \times 1$ vector, $(\mathbf{x}_t)$, is transmitted over a flat fading $N_r \times N_t$ MIMO channel with a transfer function $\mathbf{H}$, with N_r denoting the number of receive antennas. Thus, the $N_r \times 1$ dimensional received vector can be written as,

$$\mathbf{y} = \mathbf{H}\mathbf{x_t} + \mathbf{n} = \mathbf{h}^{\ell_t} s_t + \mathbf{n}, \quad (2)$$

where $\mathbf{n}$ is the N_r–dimensional additive white Gaussian noise (AWGN) with zero–mean and variance σ_n^2 per dimension at the receiver input, and $\mathbf{h}^{\ell_t} = \sum_{n \in \Upsilon_{\ell_t}} \mathbf{h}_n$, where $\mathbf{h}_n$ denotes the n–th column of $\mathbf{H}$. Note that, the signal-to-noise-ratio (SNR)= $E_s/N_o = 1/\sigma_n^2$, where $E_s = \mathrm{E}\left[\|\mathbf{H}\mathbf{x}_t\|^2\right] = 1$, where $\mathrm{E}\{\cdot\}$ is the expectation operator.

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

2.2.1. Non Line-of-Sight (Rayleigh Fading)

The entries of $\mathbf{H}$ are modelled as complex identical and independently distributed (i.i.d.) Gaussian random variables with mean $u_{\mathbf{H}} = 0$ and variance $\sigma_{\mathbf{H}}^2 = 1$.

2.2.2. Line-of-Sight (Rician Fading)

The standard statistical model for a multipath fading channel with a LOS component follows a Rician distribution. Thus, the entries of $\mathbf{H}$ are modelled as complex i.i.d. Gaussian random variables with mean $u_{\mathbf{H}} = \sqrt{\frac{K}{1+K}}$ and variance $\sigma_{\mathbf{H}}^2 = \frac{1}{1+K}$, where K is the Rician factor.

2.2.3. Spatial Correlation (SC) Model

Due to its straightforward mathematical description and acceptable accuracy, the Kronecker channel model [7] is used to model SC among antennas at either the transmitter or the receiver,

$$\mathbf{H} = \mathbf{R}_{\mathrm{Rx}}^{\frac{1}{2}} \acute{\mathbf{H}} \mathbf{R}_{\mathrm{Tx}}^{\frac{1}{2}}, \quad (3)$$

where $\acute{\mathbf{H}}$ is the uncorrelated channel, which can be a NLOS or a LOS channels, $\mathbf{R}_{\mathrm{Tx}}$ is the transmitter correlation matrix, and $\mathbf{R}_{\mathrm{Rx}}$ is the receiver correlation matrix.

The correlation matrices, $\mathbf{R}_{\mathrm{Tx}}$ and $\mathbf{R}_{\mathrm{Rx}}$, are generated using the exponential decay model as follows [7],

$$\mathbf{R} = \begin{bmatrix} 1 & r & r^2 & \cdots & r^{n-1} \\ \vdots & \ddots & \ddots & \ddots & r \\ r^{n-1} & \cdots & r^2 & r & 1 \end{bmatrix}, \quad (4)$$

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

2.2.4. Channel Estimation Error

In practical systems, the channel should be estimated at the receiver side and the channel estimate is used to decode the transmitted information. Let the estimate of $\mathbf{H}$ be $\tilde{\mathbf{H}}$. We assume that $\mathbf{H}$ and $\tilde{\mathbf{H}}$ are jointly ergodic and stationary processes. Furthermore, assuming orthogonality between the channel estimate and the estimation error due to the use least squares (LSs) estimation, the estimated channel can be modelled as [8]:

$$\tilde{\mathbf{H}} = \mathbf{H} + \mathbf{e}, \quad (5)$$

where $\mathbf{e}$ is an $N_r \times N_t$ error matrix to model the CSEs. The error matrix is a complex Gaussian random variable (RV) with zero mean and a variance equal to σ_e^2. Note that σ_e^2 is a parameter that captures the quality of the channel estimation and can be appropriately chosen depending on the channel dynamics and estimation schemes. It is assumed in this paper that for orthogonal pilot designs the estimation error reduces linearly with increasing the number of pilots [8], *i.e.*, $\sigma_e^2 = \sigma_n^2$.

2.3. ML-Optimum Detector

The maximum-likelihood (ML) optimum detector for GSM and SM can be written as,

$$\left[\hat{\ell}_t, \hat{s}_t\right] = \underset{\substack{s\in M\text{–QAM} \\ \ell\in\{1,2,\ldots,\varrho\}}}{\arg\min} \left\{\left\|\mathbf{y} - \tilde{\mathbf{h}}^{\ell} s\right\|_{\mathrm{F}}^2\right\}, \tag{6}$$

where $\tilde{\mathbf{h}}^{\ell} = \sum_{n\in\Upsilon_\ell} \tilde{\mathbf{h}}_n$, where $\tilde{\mathbf{h}}_n$ denotes the n–th column of $\tilde{\mathbf{H}}$, $\|\cdot\|_{\mathrm{F}}$ is the Frobenius norm, and $\hat{\ell}_t, \hat{s}_t$ denotes the joint estimation of the antenna index and data symbol combination, respectively.

From (6), the number of real valued multiplications required by the ML receiver for GSM and SM is, $\mathcal{C}_{\mathrm{ML}} = 6N_r 2^{\eta}$. The ML detector searches through the whole transmit and receive search spaces, and evaluating the Euclidean distance $\left(\left\|\mathbf{y} - \tilde{\mathbf{h}}^{\ell} s\right\|^2\right)$ requires $6N_r$ real multiplications.

An important observation is that the computational complexity of GSM and SM at the receiver does not directly depend on the number of transmit antennas, and it is equal to the computational complexity of single-input multiple-output (SIMO) systems. In other words, for a fixed η value, the complexity can be fixed even if the number of transmit antennas increases.

3. ANALYTICAL MODELLING OF THE ABER OF SPACE MODULATION

The ABER performance of GSM and SM is derived for NLOS and LOS channels assuming imperfect channel state information (CSI).

The ABER for space modulation systems can be computed using the union bound technique [9], and can be expressed as follows,

$$\mathrm{ABER} \leq \frac{1}{2^{\eta}} \sum_{\ell_t, s_t} \sum_{\ell, s} \frac{N(\mathbf{x}_t, \mathbf{x})}{\eta} \mathrm{E}_{\mathbf{H}}\left\{\Pr_{\text{error}}\right\}, \tag{7}$$

where $N(\mathbf{x}_t, \mathbf{x})$ is the number of bits in error between $\mathbf{x}_t$ and $\mathbf{x}$, $\mathrm{E}_{\mathbf{H}}\{\cdot\}$ is the expectation across the channel $\mathbf{H}$, and $\Pr_{\text{error}}$ is the conditional PEP of deciding on $\mathbf{x}$ given that $\mathbf{x}_t$ is transmitted.

The PEP of a space modulation system is given as follows,

$$\begin{aligned}
\Pr_{\text{error}} &= \Pr\left(\left\|\mathbf{y} - \tilde{\mathbf{H}}\mathbf{x}_t\right\|_{\mathrm{F}}^2 > \left\|\mathbf{y} - \tilde{\mathbf{H}}\mathbf{x}\right\|_{\mathrm{F}}^2\right) \\
&= Q\left(\sqrt{\frac{\left\|\tilde{\mathbf{H}}(\mathbf{x} - \mathbf{x}_t)\right\|_{\mathrm{F}}^2}{\sigma_\Gamma^2}}\right) \\
&= \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \exp\left(-\frac{\left\|\tilde{\mathbf{H}}\Psi\right\|_{\mathrm{F}}^2}{2\sigma_\Gamma^2 \sin^2\theta}\right) d\theta,
\end{aligned} \tag{8}$$

where $\sigma_\Gamma^2 = 2\left(\sigma_e^2 |s_t|^2 + \sigma_n^2\right)$, $\Psi = (\mathbf{x} - \mathbf{x}_t)$, and from [9], the alternative integral expression of the Q–function is,

$$Q(x) = \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \exp\left(-\frac{x^2}{2\sin^2\theta}\right) d\theta. \tag{9}$$

Taking the expectation of (8) yields,

$$\mathrm{E}_{\mathbf{H}}\left\{\Pr_{\text{error}}\right\} = \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \Phi\left(-\frac{1}{2\sigma_\Gamma^2 \sin^2\theta}\right) d\theta, \tag{10}$$

where $\Phi(\cdot)$ is the moment-generation function (MGF) of the random variable $\left\|\tilde{\mathbf{H}}\Psi\right\|_{\mathrm{F}}^2$. Let,

$$\begin{aligned}
\left\|\tilde{\mathbf{H}}\Psi\right\|_{\mathrm{F}}^2 &= \mathrm{tr}\left(\tilde{\mathbf{H}}\Psi\Psi^H\tilde{\mathbf{H}}^H\right) \\
&= \tilde{\mathbf{H}}_v^H \left(\mathbf{I}_{N_r} \otimes \Psi\Psi^H\right) \tilde{\mathbf{H}}_v
\end{aligned} \tag{11}$$

where tr $(\cdot)$ is the trace function, $\mathbf{I}_n$ is an $n\times n$ identity matrix, $(\cdot)^H$ denotes the Hermitian, and $\tilde{\mathbf{H}}_v = \mathbf{R}_s^{\frac{1}{2}} \mathrm{vec}\left(\acute{\mathbf{H}}^H\right)$, where vec $(\mathbf{B})$ is the vectorisation operator, and $\mathbf{R}_s = \mathbf{R}_{\mathrm{Rx}} \otimes \mathbf{R}_{\mathrm{Tx}}$, with $\otimes$ being the Kronecker product.

From [10], and (11), the MGF in (10) equals,

$$\Phi(s) = \frac{\exp\left(s \times \mathbf{u}_{\tilde{\mathbf{H}}}^H \mathbf{\Lambda}\left(\mathbf{I}_{N_r N_t} - s\mathbf{L}_{\tilde{\mathbf{H}}}\mathbf{\Lambda}\right)^{-1} \mathbf{u}_{\tilde{\mathbf{H}}}\right)}{\left|\mathbf{I} - s\mathbf{L}_{\tilde{\mathbf{H}}}\mathbf{\Lambda}\right|}, \tag{12}$$

where $\mathbf{\Lambda} = \mathbf{I}_{N_r} \otimes \Psi\Psi^H$, $\mathbf{u}_{\tilde{\mathbf{H}}} = u_{\mathbf{H}}\mathbf{R}_{\mathbf{s}}^{\frac{1}{2}} \times \mathrm{vec}\left(\mathbf{1}_{N_r N_t}\right)$, $\mathbf{L}_{\tilde{\mathbf{H}}} = \sigma_{\mathbf{H}}^2 \mathbf{R}_{\mathbf{s}} + \sigma_e^2 \mathbf{I}_{N_r N_t}$, where $\mathbf{1}_{NrNt}$ is an $N_r \times N_t$ all ones matrix, and $\times$ is the product operation.

$$\begin{aligned}
\mathrm{E}_{\mathbf{H}}\left\{\Pr_{\text{error}}\right\} &= \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \frac{\exp\left(-\frac{1}{2\sigma_\Gamma^2\sin^2\theta}\mathbf{u}_{\tilde{\mathbf{H}}}^H \Lambda\left(\mathbf{I} + \frac{1}{2\sigma_\Gamma^2\sin^2\theta}\mathbf{L}_{\tilde{\mathbf{H}}}\Lambda\right)^{-1}\mathbf{u}_{\tilde{\mathbf{H}}}\right)}{\left|\mathbf{I} + \frac{1}{2\sigma_\Gamma^2\sin^2\theta}\mathbf{L}_{\tilde{\mathbf{H}}}\Lambda\right|} d\theta \\
&\leq \frac{1}{2}\frac{\exp\left(-\frac{1}{\sqrt{2}\sigma_\Gamma^2}\mathbf{u}_{\tilde{\mathbf{H}}}^H \Lambda\left(\mathbf{I} + \frac{1}{\sqrt{2}\sigma_\Gamma^2}\mathbf{L}_{\tilde{\mathbf{H}}}\Lambda\right)^{-1}\mathbf{u}_{\tilde{\mathbf{H}}}\right)}{\left|\mathbf{I} + \frac{1}{\sqrt{2}\sigma_\Gamma^2}\mathbf{L}_{\tilde{\mathbf{H}}}\Lambda\right|}
\end{aligned} \tag{13}$$

From (10) and (12), the general PEP for space modulation systems in the presence of CSEs is given in (13). The bound in (13) takes existing bounds [8, 11] one step further. The PEP bound in (13), as shown in Section 4, is a tight closed-form bound that does not require numerical evaluation of integrals. Moreover, (13) is applicable for any fading channel where the mean and the variance are the same for the real and the imaginary part. This clearly is the case for Rayleigh and Rician channels, but would also apply to other channels such as Nakagami–m fading channels, and Weibull fading channels.

The ABER for the considered space modulation systems in the presence of CSEs can be calculated using the PEP in (13) and the upper bound in (7), which is shown to be tight upper bound for both GSM and SM MIMO systems.

4. RESULTS

In the following, Monte Carlo simulation results for at least 10^6 transmitted bits for each considered SNR value over Rayleigh and Rician fading channel realisations are conducted. This is to validate the derived upper bound and to compare the performance of GSM and SM in the presence of channel estimation errors. 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 [12]. For Rician channels, a K-factor of $K = 5$ dB is considered in all results. Because of the limited space, only results for correlated channels are shown. It assumed that correlated channels better model real–world scenarios.

4.1. Analytical Performance of SM and GSM in the Presence of Channel Estimation Errors

Simulation results of the ABER versus SNR for GSM with $N_u = 2$ and SM are depicted in Figs. 1 and 2, respectively. Monte Carlo simulation results are compared with the analytical results obtained from the derived bound in (13), where $\eta = 8$, $N_t = 16$, and $N_r = 4$. GSM uses quadrature phase shift keying (QPSK) modulation to achieve the target spectral efficiency, whereas SM uses 16–QAM for the same spectral efficiency. Analytical and simulation results for both systems demonstrate a close match for a wide and pragmatic range of SNR values. Moreover, the same observation extends to the two considered fading channels.

Considering the Rayleigh fading channel results demonstrated in Fig. 1, the GSM system is shown to outperform its SM counterpart by about 1 dB. This is mainly because SM uses a higher modulation order when compared with GSM to achieve the same spectral efficiency. However, SM is shown to outperform GSM by about 1 dB in the presence of a strong LOS component as shown in Fig. 2. This can be attributed to the existence of high spatial correlation among transmit antennas as a result of the LOS component. Spatial correlation degrades the performance of both systems when compared with the performance in Rayleigh fading channels. However, this effect is more pronounced on GSM than on SM, since GSM requires the detection of several active transmit antennas, whereas SM only requires the estimation of a single transmit antenna for every channel use.

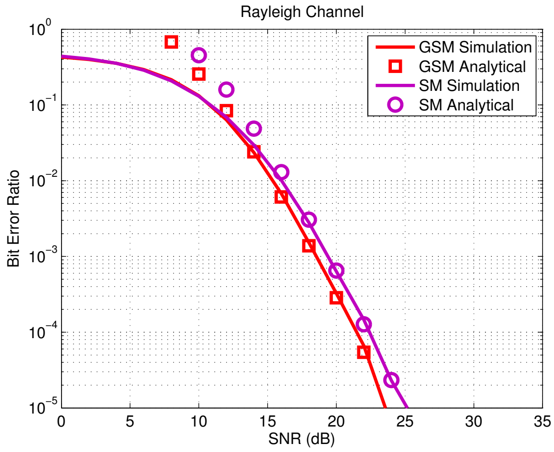

Fig. 1. ABER versus the SNR for GSM and SM over correlated Rayleigh fading channels for $\eta = 8$, and $N_t = 16$.

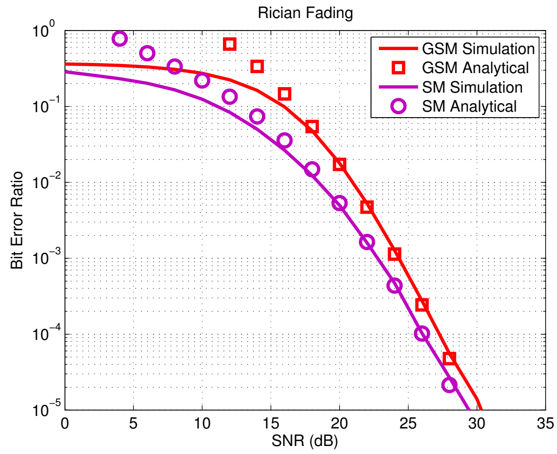

Fig. 2. ABER versus the SNR for GSM and SM over correlated Rician fading channels with $K = 5$ dB for $\eta = 8$, and $N_t = 16$.

4.2. GSM and SM ABER Performance Comparison in the Presence of Channel Estimation Errors

Large scale MIMO GSM and SM results are considered in the sequel. Monte Carlo simulation results for the ABER performance of GSM with $N_u = 2$ and SM in the presence of CSEs are shown in Fig. 3 for a Rayleigh fading channel and in Fig. 4 for a Rician fading channel. In this study, a spectral efficiency of $\eta = 12$ with $N_t = 128$ and $N_r = 4$ is assumed. As a consequence, the GSM system can use binary phase shift keying (BPSK) modulation for transmission whereas SM requires the use of 32–QAM modulation to achieve the target spectral efficiency. Note that, from (1) GSM can transmit 11 bits in the spatial constellation diagram with $N_t = 128$ and $N_u = 2$, while the SM system can transmit only 7 bits in the spatial constellation diagram.

A comparison of the GSM results with the SM results in Fig. 3 for Rayleigh fading channels shows that GSM outperform SM by about 5 dB with perfect channel knowledge and by 3 dB when considering estimation errors. Again, this is because SM uses a higher modulation order when compared with GSM to achieve the same spectral efficiency. However, SM is shown to be very robust to channel estimation errors and the performance only degrades by about 0.7 dB. This is because the probability of error for SM is not determined by the actual channel realization, rather by the differences between channels associated with the different transmit antennas. However, in GSM a group of transmit antennas need to be estimated at each particular time instant and the effect of CSEs is more pronounced. The ABER results for a Rician fading channel are shown in Fig. 4 for both GSM and SM. The presence of a LOS path degrades the performance of both systems as anticipated and as demonstrated in Sec. 4.1. GSM is shown to outperform SM with perfect channel knowledge by about 4 dB. In contrast, GSM shows similar performance to SM system when taking CSEs into account.

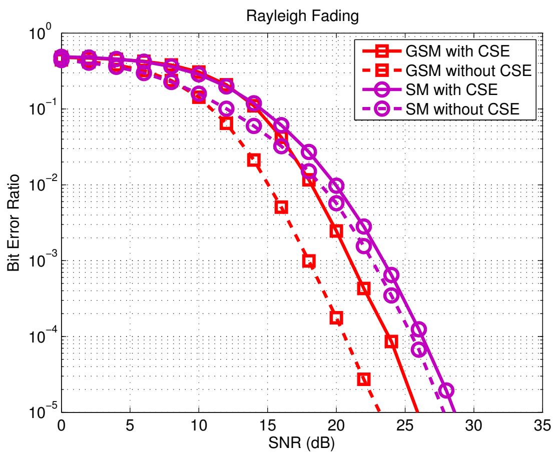

Fig. 3. ABER versus the SNR for GSM and SM over correlated Rayleigh fading channels for $\eta = 12$, and $N_t = 128$.

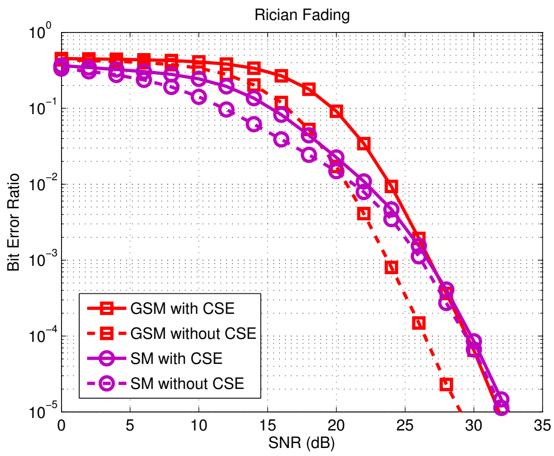

Fig. 4. ABER versus the SNR for GSM and SM over correlated Rician fading channels with $K = 5$ dB for $\eta = 12$, and $N_t = 128$.

5. CONCLUSION

In this paper, the ABER performance of large scale GSM and SM over correlated Rayleigh and Rician fading channels and in the presence of CSEs is studied and compared. A new, simple and accurate analytical closed-form upper bound is proposed. The results show that GSM is more suitable for large scale MIMO configurations than SM, and SNR performance gains of about 5 dB are reported. It is also confirmed that SM is very robust to CSEs and almost negligible performance loss occurs when taking CSEs into account. Furthermore, both techniques use only one RF chain, and guarantee high energy efficiency, and low implementation complexity. Thus, space modulation techniques have the potential to be very efficient and are highly suitable for large scale MIMO deployment.

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