Upper Bounds for the Analysis of Trellis Coded Spatial Modulation over Correlated Fading Channels

Marco Di Renzo, Raed Mesleh, Harald Haas, Peter Grant

Upper Bounds for the Analysis of Trellis Coded Spatial Modulation over Correlated Fading Channels

Marco Di Renzo(1), Raed Y. Mesleh(2), Harald Haas(3), Peter M. Grant(3)
(1) French National Center for Scientific Research (CNRS), Laboratory of Signals and Systems (LSS)
École Supérieure d'Électricité (SUPÉLEC), 3 rue Joliot–Curie, 91192 Gif–sur–Yvette (Paris), France
(2) Jacobs University Bremen, School of Engineering and Science, 28759 Bremen, Germany
(3) The University of Edinburgh, Institute for Digital Communications (IDCOM),
Joint Research Institute for Signal and Image Processing, Edinburgh, EH9 3JL, Scotland (UK)
E–Mail: marco.direnzo@lss.supelec.fr, r.mesleh@jacobs–university.de, {h.haas, peter.grant}@ed.ac.uk

***Abstract*— Trellis Coded Spatial Modulation (TCSM) is a novel transmission technology for Multiple–Input–Multiple–Output (MIMO) systems, which has been recently proposed to improve the performance of Spatial Modulation (SM) over correlated fading channels. The fundamental principle of TCSM is to use convolutional encoding and Maximum–Likelihood Sequence Estimation (MLSE) decoding to increase the free distance between sequences of spatial constellation points, thus improving, especially over spatially correlated fading channels, the end–to–end system performance. In this paper, we propose tight analytical bounds for performance analysis of TCSM over correlated fading channels. In particular, the contributions of this paper are as follows: i) we propose two asymptotically tight (for high Signal–to–Noise–Ratios, SNRs) upper bounds for the analysis of uncoded SM schemes, which offer a better accuracy than already existing frameworks, ii) we propose a simple Chernoff bound for performance analysis of TCSM, which, although weak, can well capture the diversity order of the system, and iii) we propose an asymptotically tight (for high SNRs) true union bound for the accurate performance prediction of TCSM over correlated fading channels. Analytical frameworks and findings will also be substantiated via Monte Carlo simulations.**

I. Introduction

Space Shift Keying (SSK) and Spatial Modulation (SM) are two novel and recently proposed wireless transmission techniques for Multiple–Input–Multiple–Output (MIMO) wireless systems [1], [2]. Recent research efforts have pointed out that they can be promising candidates to the design of low–complexity modulation schemes and transceiver architectures for MIMO systems over fading channels [3]–[5]. In particular, SSK and SM can offer better performance, with a significant reduction in transmitter and receiver complexity and simplification in system design, compared to other MIMO communication systems, *e.g.*, V–BLAST (Vertical Bell Laboratories Layered Space–Time), Alamouti, as well as Amplitude Phase Modulation (APM) schemes [3]–[5]. Moreover, with respect to SM, SSK modulation can reduce further the receiver complexity owing to the absence of conventional modulation schemes for data transmission [5].

The underlying and fundamental principle of SSK and SM is twofold: i) at the transmitter, a one–to–one mapping of information bits to transmit–antennas, thus allowing the transmit–antenna index to convey information, and ii) at the receiver, the exploitation, due to the properties of wireless fading channels [6], of distinct multipath profiles received from different transmit–antennas. Numerical studies in, *e.g.*, [3], [5], have pointed out that the fundamental issue to be taken into account for the accurate analysis, design, and optimization of SSK and SM is channel correlation among the transmit–receive wireless links. As a matter of fact, at the receiver–side, the optimal detector [4] is designed to exploit the distinct multipath profiles along any transmit–receive wireless link. If correlation exist among the different paths, the detector may be unable to distinguish the different transmit–antennas. In order to cope with channel correlation, a novel scheme named Trellis Coded Spatial Modulation (TCSM) has been introduced in [7], which exploits convolutional encoding and Maximum–Likelihood Sequence Estimation (MLSE) decoding to increase the free distance between sequences of spatial constellation points. Simulation results have pointed out that TCSM can provide better performance than SM and V–BLAST schemes over correlated fading channels[1], while still guaranteeing the same spectral efficiency [7]. The performance of coded SSK modulation is also analyzed in [5], where a framework for the analysis of Bit Interleaved Coded Modulation (BICM) over uncorrelated fading channels is proposed.

To the best of the authors knowledge, all performance evaluations conducted for TCSM over correlated fading scenarios to date are based on Monte Carlo numerical simulations, which, however, besides being computational intensive, only yield limited insights into the system performance. Motivated by this consideration, the main aim of this paper is to develop a simple but accurate analytical framework for the performance analysis of TCSM over correlated fading channels, which can help for a quick and simple system performance and optimization study. More specifically, we will focus our attention on a MISO (Multiple–Input–Single–Output) system setup with MLSE detection and hard–decision Viterbi decoding at the receiver, and tight upper bounds for the Average Bit Error Probability (ABEP) will be provided. Since convolutional (*i.e.*, trellis) encoding is applied to transmit–antenna indexes only in [7], in this contribution we will assume that each transmit–antenna, when activated, transmits unmodulated data. This is equivalent to applying trellis–based encoding to SSK [5], thus yielding a transmission scheme that could be named Trellis Coded Space Shift Keying (TCSSK)[2].

In particular, the specific contributions of this paper are as follows: i) we propose two asymptotically tight (for high

[1]Note that a soft–decision Viterbi decoder [8] is required in [7] to fully exploit the distance properties of the Ungerboeck's mapping by set partitioning [9]. Moreover, due to the random constellation points inherent in SSK and SM [5], the mapping rule will be optimized only on average.

[2]Throughout this paper, and for the sake of generality, we will retain the term TCSM because SSK modulation is a special case of SM.

978-1-4244-2519-8/10/$26.00 ©2010 IEEE

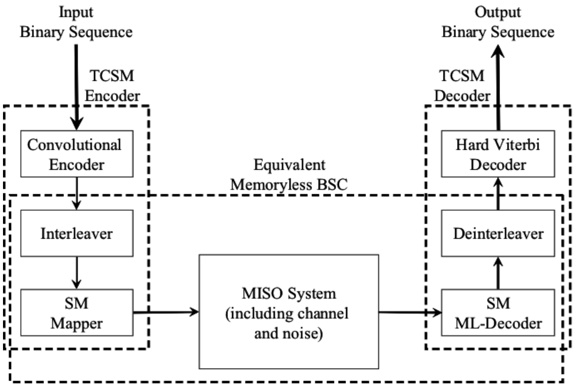

Fig. 1. TCSM system model.

Signal–to–Noise–Ratios, SNRs) upper bounds for the analysis of uncoded SSK modulation schemes, which will be shown to provide better performance estimates than the performance bounds already available in [5], ii) we compute a simple Chernoff bound for performance analysis of TCSM, which will be shown to well capture the diversity order of the system, and iii) we develop a simple and asymptotically tight (for high SNRs) true union bound for the accurate performance prediction of TCSM over correlated fading channels. With regard to i), it is worth mentioning that some improved semi–analytic bounds are proposed in [5] to take into account the effective constellation space of SSK modulation. The bounds proposed in this paper avoid the need to resort to semi–analytic modeling, and account for fading correlation as well. Moreover, TCSM instead of BICM is considered, which leads to a different analytical framework and requires tight performance bounds for the analysis of uncoded SSK and SM schemes. The extension of the framework to multiple receive antennas is left to a future contribution.

The remainder of the paper is organized as follows. In Section II, system and channel models are introduced. In Section III, the analytical framework for performance analysis over correlated fading channels is developed. In Section IV, numerical and simulation results are shown to substantiate the tightness of the proposed bounds. Finally, Section V concludes the paper.

II. System Model

A. Background

Let us consider the TCSM–based $M \times 1$ MISO system depicted in Fig. 1, where M denotes the number of transmit–antennas. TCSM works as follows[3]. Random binary i.i.d. (independent and identically distributed) information bits are input to a rate $R_c = m_c/n_c$ convolutional (*i.e.*, trellis) encoder whose output bits are then randomly block–interleaved[4] to break up fading bursts [10]. Groups of n_c interleaved bits are then mapped[5] into M spatial constellation points, which are chosen from the set of $M = 2^{n_c}$ transmit–antennas. The SM mapper in Fig. 1 is responsible for this mapping. Depending on the n_c–long codeword, a single transmit–antenna is activated for data transmission, according to the principle of SM and SSK modulation [3], [5]. In general, a simple unmodulated sinusoidal carrier [1] or a single pulse [5] can be transmitted by the activated antenna to convey the desired information, *i.e.*, the n_c–long codeword of information bits, to the final destination. At the receiver–side, the faded and noisy received signal is processed by a ML detector with full Channel State Information (CSI) [4] to retrieve the n_c–long codeword, *i.e.*, the coded transmit–antenna index. Then, this ML–optimum detector passes the hard–quantized coded bits through a random block–deinterleaver to recreate the codewords temporarily scrambled by the interleaver at the transmitter. The hard–quantized and deinterleaved bits are then passed to a hard–decision Viterbi [8] decoder, which retrieves the information bits originally emitted by the binary source at the transmitter.

[3]Note that, due to space constraints, a detailed discussion of TCSM, SM, and convolutional encoding is beyond the scope of this paper. The reader is referred to [3], [7], and [10] for a thorough treatment of them.

[4]For analytical tractability, we will assume an infinite interleaving depth, thereby resulting in an ideal memoryless channel. In practice, the depth of the interleaving will be finite and chosen in relation to the maximum duration of fade anticipated.

[5]Note that, owing to the random constellation points inherent in SM and SSK modulation, as well as to the adoption of a hard–decision Viterbi decoder, we do not consider any optimized mapping rule. The investigation of the optimal mapping rule to minimize the ABEP is left to a future contribution.

B. Notation

Let us briefly summarize the main notations used in what follows: i) we adopt a complex–envelope signal representation; ii) $\delta(\cdot)$ is the Dirac delta function; iii) $|\cdot|^2$ denotes square absolute value; iv) $\mathrm{E}\{\cdot\}$ is the expectation operator; v) T_m denotes the signaling interval for the transmission of each antenna index; vi) ρ_{AB} denotes the correlation coefficient of Random Variables (RVs) A and B; vii) $Q(x) = \left(1/\sqrt{2\pi}\right)\int_x^{+\infty} \exp\left(-t^2/2\right) dt$ is the Q–function; viii) E_m is the average energy transmitted by each antenna that emits a non–zero signal; ix) the noise at the receiver input is assumed to be Additive White Gaussian (AWG–) distributed, with both real and imaginary parts having a double–sided power spectral density equal to N_0; x) for ease of notation, we set $\bar{\gamma}=E_m/(4N_0)$; xi) $j = \sqrt{-1}$ is the imaginary unit; xii) $G \sim N\left(\mu_G, \sigma_G^2\right)$ is a Gaussian–distributed RV with mean μ_G and standard deviation σ_G; xiii) p_{BSC} is the average error probability of the equivalent Binary Symmetric Channel (BSC) shown in Fig. 1; xiv) $T(D,N)$ denotes the transfer function of the augmented state diagram of the convolutional encoder used at the transmitter; xv) d_{free} denotes the free distance of the convolutional code; vxi) $\binom{\cdot}{\cdot}$ is the binomial coefficient; xvii) $\mathrm{PEP}(\mathrm{TX}_k \to \mathrm{TX}_h)$ denotes the Pairwise Error Probability (PEP) between the transmit–antennas TX_k and TX_h with $k,h = 1,2,\ldots,M$, *i.e.*, the probability of detecting TX_h when, instead, TX_k is actually transmitting. Likewise, $\mathrm{PEP}\left(\mathrm{TX}_k \to \mathrm{TX}_h | A,B\right)$ is the same performance metric when conditioning onto RVs A and B; xviii) $N(k,h)$ is the number of information bit errors committed by choosing TX_h instead of TX_k as transmit–antenna; and xix) K is the constraint length of the convolutional encoder.

C. Channel Model

We consider a frequency–flat fading channel model, with fading envelopes distributed according to a Rayleigh distribution [11]. Moreover, we assume the fading gains not to be necessarily identically distributed, and spatial correlation among them will be accounted for in this paper. In particular:

- $\{h_i(t)\}_{i=1}^M = \beta_i \exp(j\varphi_i)\,\delta(t-\tau_i)$ is the channel impulse response from the i–th transmit–antenna to the single receive antenna of the MISO system depicted in Fig. 1, with $\{\beta_i\}_{i=1}^M$, $\{\varphi_i\}_{i=1}^M$, and $\{\tau_i\}_{i=1}^M$ denoting gain,

$$P_k = \begin{cases} \sum\limits_{e=(k+1)/2}^{k} \binom{k}{e} p_{\text{BSC}}^{e} \left(1 - p_{\text{BSC}}^{k-e}\right) & k \text{ odd} \\ \frac{1}{2}\binom{k}{k/2} p_{\text{BSC}}^{k/2} \left(1 - p_{\text{BSC}}^{k/2}\right) + \sum\limits_{e=(k/2)+1}^{k} \binom{k}{e} p_{\text{BSC}}^{e} \left(1 - p_{\text{BSC}}^{k-e}\right) & k \text{ even} \end{cases} \quad (3)$$

phase, and delay of the related wireless link. Moreover, $\{\alpha_i\}_{i=1}^M = \beta_i \exp(j\varphi_i)$ denotes the channel complex–gain of i–th transmit–receive path.

- According to a Rayleigh fading channel model, the channel complex–gains, $\{\alpha_i\}_{i=1}^M$, reduce to $\{\alpha_i\}_{i=1}^M = \{\alpha_i^R\}_{i=1}^M + j\{\alpha_i^I\}_{i=1}^M$, where $\{\alpha_i^R\}_{i=1}^M \sim N(0, \sigma_i^2)$ and $\{\alpha_i^I\}_{i=1}^M \sim N(0, \sigma_i^2)$, with $\{\alpha_i^R\}_{i=1}^M$ being independent from $\{\alpha_i^I\}_{i=1}^M$. Accordingly, $\{\beta_i\}_{i=1}^M$ and $\{\varphi_i\}_{i=1}^M$ will be Rayleigh and uniform distributed RVs, respectively.
- The following spatial correlation model between each pair (i, l), with $i = 1, 2, \ldots, M$ and $l = 1, 2, \ldots, M$, of transmit–receive wireless links is assumed: i) $\rho_{\alpha_i^R \alpha_i^I} = \rho_{\alpha_l^R \alpha_l^I} = \rho_{\alpha_i^R \alpha_l^I} = \rho_{\alpha_l^R \alpha_i^I} = 0$, and ii) $\rho_{\alpha_i^R \alpha_l^R} = \rho_{\alpha_i^I \alpha_l^I} = \rho_{i,l}$. Although this correlation model is not the most general one, it will allow us to get insightful and simple closed–form results for the system under analysis, while still guaranteeing good adherence to physical reality. An analytical framework for a more general fading and correlation model can be found in [12].
- $\{\tau_i\}_{i=1}^M$ and $\{\varphi_i\}_{i=1}^M$ are assumed to be independent and uniformly distributed in $[0, T_m)$ and $[0, 2\pi)$, respectively.

III. Upper Bounds for the ABEP of TCSM

A. Methodology

The methodology used for performance analysis of TCSM is as follows. Owing to the adoption of a random and infinite long interleaver and a hard–decision Viterbi decoder at the receiver, the block diagram in Fig. 1 can be readily shown to be equivalent to a memoryless BSC whose inputs are the outputs of a generic convolutional encoder, and whose outputs are the inputs of a MLSE decoder, which implements the hard–decision Viterbi algorithm. This equivalent memoryless BSC is highlighted in Fig. 1, and includes the interleaver/deinteleaver, the mapping onto spatial (*i.e.*, antenna) constellation points, the optimum ML detector for SM, as well as the MISO system. So, the TCSM system model shown in Fig. 1 reduces to the equivalent communication system in, *e.g.*, [13, Fig. 10].

In the light of the above equivalence with the BSC, the ABEP of TCSM can be computed by resorting to the general theory for performance analysis of convolutional codes over memoryless BSCs [13, Sec. VII–A], which, however, needs to be specialized to the specific signal structure of SM and SSK modulation schemes over correlated fading channels: this is the main aim of this section.

B. ABEP of Convolutional Codes over BSCs

A general approach for the analysis of the performance of convolutional codes over memoryless channels is described in [13] in a comprehensive fashion. In particular, for BSCs the ABEP of MLSE detection can be computed either using the simple Chernoff Bound (CB) or the True Union Bound (TUB), as summarized in what follows.

1) Chernoff Bound: The Chernoff Bound for the ABEP of TCSM is as follows [13, Eq. (21)]:

$$\text{ABEP} \le \text{ABEP}_{\text{CB}} = \left. \frac{dT(D,N)}{dN} \right|_{N=1,\, D=2\sqrt{p_{\text{BSC}}(1-p_{\text{BSC}})}} \quad (1)$$

2) True Union Bound: The True Union Bound for the ABEP of TCSM is as follows [13, Eq. (20)]:

$$\text{ABEP} \le \text{ABEP}_{\text{TUB}} = \sum_{k=d_{\text{free}}}^{+\infty} c_k P_k \quad (2)$$

with P_k being defined as shown in (3) on top of this page, and c_k being weighting coefficients that can be obtained from $T(D,N)$ as follows:

$$\left. \frac{dT(D,N)}{dN} \right|_{N=1} = \sum_{k=d_{\text{free}}}^{+\infty} c_k D^k \quad (4)$$

According to (1)–(4), both CB and TUB can be computed when both p_{BSC} and $T(D,N)$ are known in closed–form. In particular, p_{BSC} depends on the specific transmission technology used to convey the information from the transmitter to the receiver, *i.e.*, either SM or SSK modulation in this paper. On the other hand, $T(D,N)$ depends on the particular trellis encoder used at the transmitter. In Section III-C, tight bounds for computing p_{BSC} over correlated Rayleigh fading channels and an arbitrary number of transmit–antennas will be presented. In Section III-D, two case studies for $T(D,N)$ will analyzed when four (*i.e.*, $M = 4$) and eight (*i.e.*, $M = 8$) transmit–antennas are considered.

C. Bounds for Computing p_{BSC}

By carefully looking at Fig. 1, we can readily observe that p_{BSC} is the ABEP of an uncoded $M \times 1$ MISO system with SSK modulation at the transmitter and CSI–assisted ML (antenna index–by–antenna index) detection [4] at the receiver. Performance bounds for computing the ABEP of uncoded SSK modulation can be found in [5] for uncorrelated Rayleigh fading and an arbitrary number of transmit– and receive–antennas. However, these bounds have two main limitations: i) the bound in [5, Eq. (8)] is relatively weak, and ii) the bound in [5, Eq. (9)] is semi–analytic and requires the estimation of the effective constellation points of SSK modulation. Moreover, they have been derived and analyzed for uncorrelated fading. In this section, we propose two new upper bounds for performance prediction.

1) Symbol–based Union Bound for p_{BSC}: The first bound, which is called Symbol–based Union Bound (SUB), can be obtained by using typical methods for performance analysis of multi–level modulation schemes with optimum detection [14, Sec. 5.7]. In particular, the average error probability for the antennas index, *i.e.*, the Average Constellation Error Probability (ACEP), can be readily upper bounded via union bound techniques, as shown in what follows [14, Eq. (5.86)]:

$$\mathrm{ACEP} \le \frac{1}{M}\sum_{k=1}^{M}\sum_{h\neq k=1}^{M} \mathrm{PEP}\left(\mathrm{TX}_k \to \mathrm{TX}_h\right) \quad (5)$$

From the ACEP in (5), an upper bound, $p_{\rm BSC}^{\rm SUB}$, for $p_{\rm BSC}$ can be obtained by using [14, Eq. (5.101)], which assumes that the errors for all antenna indexes are equally likely, as follows:

$$p_{\rm BSC} \le p_{\rm BSC}^{\rm SUB} = \frac{M/2}{M-1}\mathrm{ACEP} \quad (6)$$

2) Codeword–based Union Bound for $p_{\rm BSC}$: The second bound, which is called Codeword–based Union Bound (CUB), can be obtained by using typical methods for performance analysis of MLSE detectors [11, Sec. 13.1.3]. In particular, an upper bound, $p_{\rm BSC}^{\rm CUB}$, for $p_{\rm BSC}$ can be obtained from [11, Sec. 13.44], as shown in what follows ($p_{\rm BSC} \le p_{\rm BSC}^{\rm CUB}$):

$$p_{\rm BSC}^{\rm CUB} = \frac{1}{\log_2(M)}\frac{1}{M}\sum_{k=1}^{M}\sum_{h\neq k=1}^{M} N(k,h)\,\mathrm{PEP}\left(\mathrm{TX}_k \to \mathrm{TX}_h\right) \quad (7)$$

where we have taken into account the following facts: i) in an uncoded system the number of information bits per transmission is $\log_2(M)$, and ii) due to the interleaver in Fig. 1 the antenna indexes are transmitted with equal probability, which is equal to $1/M$.

3) Computation of $\{\mathrm{PEP}(\mathrm{TX}_k \to \mathrm{TX}_h)\}_{k\neq h=1}^{M}$: Both bounds in (6), (7) need a closed–form expression of $\{\mathrm{PEP}(\mathrm{TX}_k \to \mathrm{TX}_h)\}_{k\neq h=1}^{M}$. For uncorrelated Rayleigh fading channels the PEP can be found in [5], while for correlated Nakagami–*m* fading channels a general framework has been recently proposed in [12]. In particular, the PEP conditioned onto fading channel statistics is as follows [5], [12]:

$$\mathrm{PEP}\left(\mathrm{TX}_k \to \mathrm{TX}_h|\,\alpha_k,\alpha_h\right) = Q\left(\sqrt{\bar{\gamma}\,|\alpha_k-\alpha_h|^2}\right) \quad (8)$$

By following the same methodology described in [12], and specializing the result for the fading and correlation model described in Section II-C, the PEP can be written, after a few algebraic manipulations, which are here omitted due to space constraints, as shown in what follows:

$$\begin{aligned}\mathrm{PEP}\left(\mathrm{TX}_k \to \mathrm{TX}_h\right) &= \mathrm{E}\left\{\mathrm{PEP}\left(\mathrm{TX}_k \to \mathrm{TX}_h|\,\alpha_k,\alpha_h\right)\right\} \\ &= \frac{1}{2} - \frac{1}{2}\sqrt{\frac{\bar{\sigma}_{h,k}^2\bar{\gamma}}{1+\bar{\sigma}_{h,k}^2\bar{\gamma}}}\end{aligned} \quad (9)$$

where, by using the notation in Section II-C, we have defined $\bar{\sigma}_{h,k}^2 = \sigma_h^2 + \sigma_k^2 - 2\rho_{h,k}\sigma_h\sigma_k$.

D. Examples of $T(D,N)$ for $M=4$ and $M=8$

Let us now consider two examples of convolutional encoders in order to clearly illustrate how the coefficients of the series in (3) and (4) can be computed in practice. These encoders will be used in Section IV.

1) Convolution Encoder for a 4×1 MISO System: An example of convolutional encoder for a 4×1 MISO system can be found in [13, Fig. 1]. The coder has constraint length $K=3$, coding rate $R_c = 1/2$, free distance $d_{\rm free} = 5$, and the augmented transfer function is $T(D,N)$:

$$T(D,N) = \frac{ND^5}{1-2ND} = \sum_{k=5}^{+\infty} 2^{k-5}N^{k-4}D^k \quad (10)$$

Accordingly, from (10), the weighting coefficients in (4) are $\{c_k\}_{k=5}^{+\infty} = (k-4)\,2^{k-5}$, from which the $\mathrm{ABEP}_{\rm TUB}$ in (2) can be easily computed.

2) Convolution Encoder for a 8×1 MISO System: An example of convolutional encoder for a 8×1 MISO system can be found in [15, Fig. 8.2.2]. The coder has constraint length $K=3$, coding rate $R_c = 1/3$, free distance $d_{\rm free} = 6$, and the augmented transfer function is $T(D,N)$:

$$T(D,N) = \frac{ND^6}{1-2ND^2} = \sum_{\substack{k=6 \\ k\ \text{even}}}^{+\infty} 2^{(k-6)/2}N^{(k-4)/2}D^k \quad (11)$$

Accordingly, from (11), the weighting coefficients in (4) are $\{c_k\}_{k=6,\ k\ \text{even}}^{+\infty} = 0.5\,(k-4)\,2^{2^{(k-6)/2}}$, from which the $\mathrm{ABEP}_{\rm TUB}$ in (2) can be easily computed.

IV. Numerical and Simulation Results

In this section, we show some numerical results in order to substantiate the accuracy of the proposed bounds for TCSM over correlated fading channels. The following system setup is considered. As far as the fading scenario is concerned, we consider two case studies: balanced and unbalanced setups. In the former case, we set $\{\sigma_i^2\}_{i=1}^{M} = 1$, while in latter case $\{\sigma_i^2\}_{i=1}^{M}$ are assumed to be independent and uniformly distributed (one–shot realization) in $[1,3]$. Moreover, both uncorrelated and correlated fading is considered. In the latter case we set $\{\rho_{i,l}\}_{i,l=1}^{M} = \exp(-|i-l|/2)$. As far as Monte Carlo simulations are concerned: i) we require 1000 bit error events for estimating the ABEP; ii) the interleaver depth is set to 1000; and iii) the traceback depth is set to 15. Regarding the computation of (2), we truncate the infinite series after the first 10 non–zero terms. Both convolutional encoders in Section III-D have been considered. Finally, although no optimized SM mapping rules are used, to enable the reproducibility of the results we adopt a natural mapping[6] for the spatial constellation points.

In Figs. 2, 3, the error propagability $p_{\rm BSC}$ for various system settings is shown and compared with the bounds proposed in Section III-C, along with the bound proposed in [5, Eq. (4)]. We can observe a very good accuracy of the proposed frameworks for various system and channel conditions. We note that the bound introduced in [5] overestimates the simulated $p_{\rm BSC}$ of approximately 3dB and 5dB for the 4×1 and 8×1 MISO systems, respectively. On the contrary, both bounds proposed in this paper are asymptotically tight for high (but pragmatic) SNRs and yield almost the same performance.

In Figs. 4, 5, the ABEP of TCSM is shown and the proposed bounds are compared to each other and with Monte Carlo simulations. The error probability of the equivalent BSC is computed by using $p_{\rm BSC}^{\rm SUB}$. We can observe that both the CB in (1) and the TUB in (2) can capture the diversity order of the system, *i.e.*, the slope of the ABEP for high SNRs. However, the CB in (1) is relatively weak and errors between 2.5dB and 5dB can be observed in the analyzed system setups. On the other hand, the TUB in (2) is fairly accurate and asymptotically tight for high SNRs. These results substantiate the analytical frameworks introduced in this paper, and confirm that they can be used for the accurate analysis

[6]Each antenna index is identified by its base–2 (binary) equivalent representation.

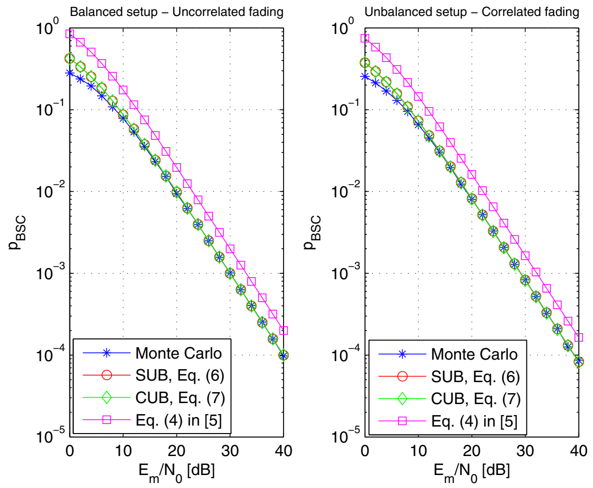

Fig. 2. p_{BSC} for a 4×1 MISO system: simulation vs. bounds.

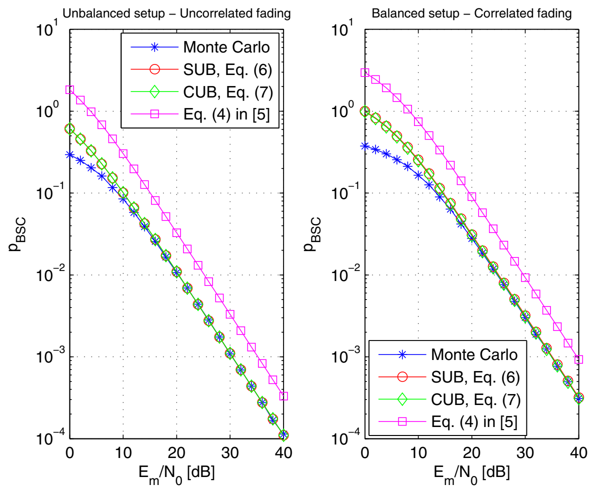

Fig. 3. p_{BSC} for a 8×1 MISO system: simulation vs. bounds.

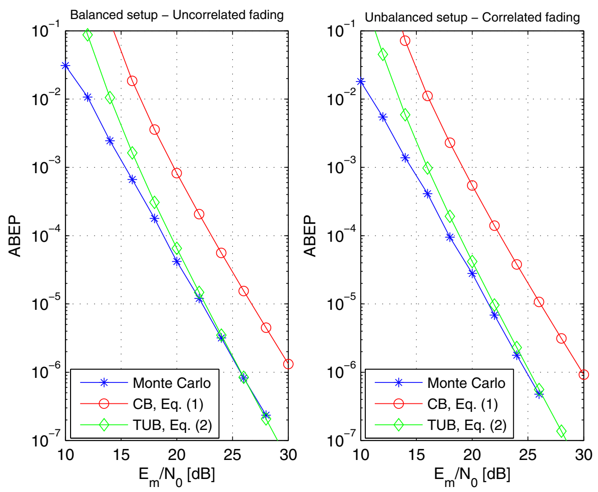

Fig. 4. ABEP for a 4×1 MISO system: simulation vs. bounds.

and optimization of TCSM over correlated fading channels. Finally, by comparing Figs. 2–5 we observe that, due to the coding gain of convolutional encoding at the transmitter and MLSE decoding at the receiver, the performance of coded SM (*i.e.*, ABEP) is much better than uncoded SM (*i.e.*, p_{BSC}).

V. Conclusion

In this paper, we have proposed some analytical bounds for performance analysis of TCSM over correlated Rayleigh fad-

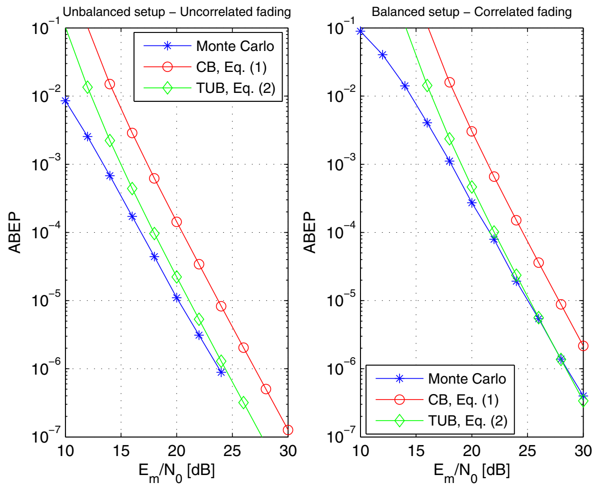

Fig. 5. ABEP for a 8×1 MISO system: simulation vs. bounds.

ing channels. The comparison with Monte Carlo simulations has shown a good accuracy of the frameworks for various system setups, channel fading conditions, and convolutional encoders. The bounds introduced for uncoded SM offer tighter estimates than other frameworks available in the literature.

Acknowledgment

We gratefully acknowledge support from the EPSRC (EP/G011788/1) for this work. Harald Haas acknowledges the Scottish Funding Council support of his position within the Edinburgh Research Partnership in Engineering and Mathematics between the University of Edinburgh and Heriot Watt University.

References

[1] Y. Chau and S.–H. Yu, "Space modulation on wireless fading channels", *IEEE Veh. Technol. Conf. – Fall*, vol. 3, pp. 1668–1671, Oct. 2001.
[2] H. Haas, E. Costa, and E. Schultz, "Increasing spectral efficiency by data multiplexing using antennas arrays", *IEEE Int. Symp. Personal, Indoor, Mobile Radio Commun.*, vol. 2, pp. 610–613, Sept. 2002.
[3] R. Y. Mesleh, H. Haas, S. Sinanovic, C. W. Ahn, and S. Yun, "Spatial modulation", *IEEE Trans. Veh. Technol.*, vol. 57, vol. 4, pp. 2228–2241, July 2008.
[4] J. Jeganathan, A. Ghrayeb, and L. Szczecinski, "Spatial modulation: Optimal detection and performance analysis", *IEEE Commun. Lett.*, vol. 12, no. 8, pp. 545–547, Aug. 2008.
[5] J. Jeganathan, A. Ghrayeb, L. Szczecinski, and A. Ceron, "Space shift keying modulation for MIMO channels", *IEEE Trans. Wireless Commun.*, vol. 8, no. 7, pp. 3692–3703, July 2009.
[6] L. Xiao, L. Greenstein, N. Mandayam, and W. Trappe, "Using the physical layer for wireless authentication in time–variant channels", *IEEE Trans. Wireless Commun.*, vol. 7, no. 7, pp. 2571–2579, July 2008.
[7] R. Y. Mesleh, I. Stefan, H. Haas, and P. M. Grant, "On the performance of trellis coded spatial modulation", *Int. ITG Workshop on Smart Antennas*, pp. 235–241, Feb. 2009.
[8] A. J. Viterbi, "Error bounds for convolutional codes and an asymptotically optimum decoding algorithm", *IEEE Trans. Inform. Theory*, vol. 13, no. 2, pp. 260–269, Apr. 1967.
[9] G. Ungerboeck, "Channel coding with multilevel/phase signals", *IEEE Trans. Inform. Theory*, vol. 28, no. 1, pp. 55–67, Jan. 1982.
[10] E. Biglieri, *Coding for Wireless Channels*, Springer, 1st ed., May 2005.
[11] M. K. Simon and M.–S. Alouini, *Digital Communication over Fading Channels*, John Wiley & Sons, 1st ed., 2000.
[12] M. Di Renzo and H. Haas, "On the performance of SSK modulation over correlated Nakagami–*m* fading channels", *IEEE Int. Conf. Commun.*, Cape Town, South Africa, May 2010.
[13] A. J. Viterbi, "Convolutional codes and their performance in communication systems", *IEEE Trans. Commun. Technol.*, vol. 19, no. 5, pp. 751–772, Oct. 1971.
[14] S. Haykin, *Communication Systems*, Wiley, 4th ed., 2000.
[15] J. Proakis, *Digital Communications*, McGraw–Hill, 4th ed., 2000.