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Spatially Modulated Non Orthogonal Space Time Block Code: Construction and design from cyclic codes over Galois Field

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ABSTRACT

A new class of non-binary Spatially Modulated Non-orthogonal Space Time Block Code designs (SM-NSTBC) has been proposed. These designs employ full rank, length $n, (n|q^m - 1, m \leq n)$ cyclic codes defined over $GF(q^m)$. The underlying cyclic code constructions have the property that the codewords when viewed as $m \times n$ matrices over $GF(q)$ have rank equal to m (Full rank). These codes are punctured to yield $m \times m$ full rank matrices over $GF(q)$. Rank preserving transformations are used to map the codewords of full rank codes over a finite field to full rank Space Time Block Codes. The proposed scheme can be generalized to handle any number of transmit antenna greater than two. Due to the characteristics of Full rank cyclic codes employed, a coding gain of approximately 1.5 dB to 5 dB is obtained over conventional STBC-SM and SM-OSTBC schemes. This is demonstrated for spectral efficiencies of 4, 5, 7 and 8 bpcu. Analytical as well as Monte-Carlo simulations show that proposed SM-NSTBC outperforms STBC-SM and its variants. The upper bound on average bit error rate has been derived and the computation complexity for ML detection has been estimated.

1. Introduction

It has been shown that the most effective and constructive way of increasing the performance of a wireless system is to use multiple antenna combinations at both transmitter and receiver terminals [1]. This type of arrangement has an inherent advantage of sustaining various signal fading effects. The two major limitations in wireless communication systems are power and the bandwidth. Multiple Input Multiple Output (MIMO) schemes have been deployed to improve the system throughput and reliability. MIMO systems have brought about significant improvements in the performance of wireless systems. These improvements are obtained at the cost of enhancement in the complexity at the transmitter and receiver. Thus, improvements in spectral efficiency, reliability of information transfer and throughput eventually increases the system complexity and power consumption. Recently, there has been an active interest in designing novel wireless architectures that can yield some or all of the desirable goals with reduced complexity or power consumption [2]. Developments in MIMO technology such as Spatial Multiplexing (SMX) and Space Time Block Codes (STBCs) have contributed to the improvement in spectral efficiencies and reliability of information transfer.

Space Time Block Codes were first introduced in [3], which demonstrated the concept of Space-Time Codes for high data rate Wireless communications by using transmit antenna diversity. The next generation 5G networks demand very high levels of reliability, high data rate and energy efficiency. Keeping these objectives in mind, researchers have developed intelligent and novel physical layer modulation techniques such as Spatial Modulation (SM) [2] and STBC-SM schemes [4]. In SM systems, the spatial domain is used as one of the information bearing dimensions to convey data bits that adds to the extra spectral efficiency. This scheme also has the potential to reduce energy consumption as well as complexity by activating only one RF chain through the process of transmission. The requirement for having multiple RF chains is eliminated. However, it has to be noted that single active antenna transmission will not improve the performance of the system in terms of transmit diversity. Transmit diversity plays a major role in optimizing the performance of a system when evaluated in a deep fading environment. The concatenation of orthogonal codes designed as STBCs are used in conjunction with SM systems to yield a scheme known as Space Time Block Coded Spatial Modulation (STBC-SM) system [4]. This STBC-SM method represents information in the form of a STBC matrix, which is transmitted from different combinations of transmit antennas of a MIMO system. STBC-SM was first implemented with the Alamouti code by Basar et al. in [4]. In this design, bits are divided into two parts, the information bearing bits and the antenna selection bits. The information bearing bits are used

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to constitute two complex information symbols which are used to form Alamouti STBC, and the antenna selection bits (antenna indices) are used to determine the particular antenna which will radiate the information symbols for transmission.

Another higher order transmit diversity scheme namely Space-Time Shift Keying (STSK) was proposed by Sugiura et al. [5] in 2010. A number of other high rate STBC-SM systems have been proposed in the recent past [6,7] and [8]. Lie Wang in [8] has proposed a STBC-SM with an embedded (n, k) cyclic error correcting code (ECC). This scheme provides a high spectral efficiency as a result of the large number of codewords employed. Another approach was designed to enhance the spectral efficiency of SM systems and is designated as Space Time Block Coded Spatial Modulation with Cyclic structure (STBC-CSM), given by Xaiofeng Li et al. in [9]. Helmy et al. [10] had proposed STBC-TSM which utilizes the temporal modulation to further improve the performance of STBC-SM. In STBC-TSM, cyclic spatially modulated rotated Alamouti STBCs is employed to obtain full transmit diversity over four consecutive time slots. The concept of utilizing the active antennas as layers was proposed in [11], which would ensure diversity and coding gain at each antenna selection due to the use of multi-layer coding scheme.

A comprehensive study of the above literature has inspired us to explore the use of full rank codes. Full rank codes obtained from cyclic codes are used in synthesizing designs for Spatially Modulated Non Orthogonal Space Time Block Codes (SM-NSTBC). We have derived $n-$ length cyclic codes over $GF(q^m)$ where $m \leq n$, has rank m when viewed as $m \times n$ matrices over $GF(q)$ [12] and [13]. Using Rank preserving transformations, these full rank $m \times n$ matrices over $GF(q)$ are transformed into equivalent full rank matrices over the complex number field. This collection of all full rank matrices over the complex number field constitutes the SM-NSTBC design. The full rank property of the codewords contributes to the improvement in terms of bit error rate (BER) performance over STBC-SM, SM schemes and its variants.

The major contributions of this paper are enumerated below:

- A new SM-MIMO transmission system, called SM-NSTBC, is presented. Following [13] we have designed several full rank NSTBCs from Non-binary cyclic codes over $GF(q^m)$.
- The proposed transmission scheme makes use of Gaussian and Eisenstein integers maps as given in [14–16] to transform full rank codewords constructed over a finite field to full rank codewords over the complex number field. Further, unlike the standard STBC-SM schemes the codewords are radiated using different transmit antenna combinations from the column vectors of the derived NSTBC matrix.
- A comprehensive framework for the SM-NSTBC is devised in the form of choosing and adapting the same procedure for different active antenna combinations. We have explored the use of $N_a = 2$ and $N_a = 4$ active antenna combinations.
- It is shown by computer simulations that the proposed SM-NSTBC scheme exhibits significant performance improvements over the STBC-SM, SM described in the literature. A closed form expression for the union bound on the bit error rate of the SM-NSTBC scheme is also derived to support our results. The derived upper bound is shown to become very tight with increasing signal-to-noise (SNR) ratio. Finally, the computational complexity for the proposed scheme has been estimated.

The rest of the paper is organized as follows. Section 2, describes the design and characterization of SM-NSTBC. In Section 3, we describe the SM-NSTBC system model. Analytical performance is described in Section 4, Section 5, gives the complexity analysis followed by simulation results in Section 6. Section 7 contains conclusions.

Notation. In this paper, vectors and matrices are represented by bold lowercase or uppercase letters, $\mathbf{X}^T$ represents the transpose of matrix $\mathbf{X}$, $\mathbf{X}^H$ represents the matrix Hermitian of $\mathbf{X}$, $\|\mathbf{X}\|_F^2$ denotes the Frobenius norm of the matrix. $GF(q)$ represents the Galois Field of size q and $GF(q^m)$ is m^{th} Galois Field extension of $GF(q)$.

2. Characterization of cyclic codes from transform domain

The preliminary information required to understand the design of codewords of the proposed SM-NSTBC scheme is provided in this section. The codewords used for NSTBC scheme design are derived from full rank cyclic codes by employing a Rank preserving map. Full rank cyclic codes over $GF(q)$ are obtained by making use of the Galois Field Fourier Transform (GFFT) description. The GFFT Transform Pair are defined as follows:

Let $\mathbf{a} = (a_0, a_1, \ldots, a_{n-1}) \in GF(q^m)$, where q is prime and $gcd(q, n) = 1$, Let τ be the smallest positive integer such that $n|q^{m\tau} - 1$ and $\alpha \in GF(q^{m\tau})$ be an element of order n. Then the Galois Field Fourier Transform (GFFT) of $\mathbf{a}$ is given by $\mathbf{A} = (A_0, A_1, \ldots A_{n-1})$, where

$$A_j = \sum_{i=0}^{n-1} \alpha^{ij} a_i \qquad j = 0, 1, \ldots, n-1 \tag{1}$$

The Inverse Galois Field Fourier Transform (IGFFT) is defined as,

$$a_i = \frac{1}{n\ modulo\ p} \sum_{j=0}^{n-1} \alpha^{-ij} A_i \qquad i = 0, 1, \ldots, n-1 \tag{2}$$

Here, $\mathbf{a}$ and $\mathbf{A}$ are considered to be time domain and transform domain Fourier Transform pair [17].

An $n-$length vector $\mathbf{a}$ over $GF(q^m)$ can be represented as $m \times n$ matrix over $GF(q)$. The rank of the vector is equivalent to the rank of matrix over $GF(q)$. In brief, consider a vector $\mathbf{a}$ of length n over $GF(q^m)$. It can be expressed as a $m \times n$ matrix over $GF(q)$ as given below:

$$\begin{bmatrix} a_{0,0} & a_{0,1} & \cdots & a_{0,n-1} \\ a_{1,0} & a_{1,1} & \cdots & a_{1,n-1} \\ \vdots & \vdots & \ldots & \vdots \\ a_{m-1,0} & a_{m-1,1} & \cdots & a_{m-1,n-1} \end{bmatrix} \tag{3}$$

In this matrix, $a_{i,j} \in GF(q)$. The rank of this matrix is equivalent to the rank of codeword $\mathbf{a}$, where $\mathbf{a} = \{\mathbf{a_0}, \mathbf{a_1} \ldots \mathbf{a_{n-1}}\}$ and the individual $\mathbf{a_i} \in GF(q^m)$. In order to obtain cyclic codes with full rank property, we will employ the following two theorems from [12,18].

Theorem 1 (*[12]*). *Consider $\mathcal{C}$ to be a cyclic code of length n, $(n|q^m - 1)$ over $GF(q^m)$ i.e. n divides $q^m - 1$, let $A_{jq^s} = A_{[j]}, (|[j]| = e_j, 0 \leq s \leq e_j - 1)$ be a single free transform domain component and all other transform components be constrained to zero, then the $Rank_q(\mathcal{C}) = e_j$. Here, e_j is order of the q-cyclotomic coset $[j]$.*

Theorem 2 (*[18]*). *Let $\mathcal{C}$ be a cyclic code of length n, $(n|q^m - 1)$ over $GF(q^m)$ with a single free transform component $A_{jq^s}, 0 \leq s \leq e_j - 1$ and all other transform components constrained to zero. For any non-zero codeword $\mathbf{a} = (a_0, a_1, \ldots a_{e_j-1}, a_{e_j}, \ldots, a_{n-1}) \in \mathcal{C}$ the entries of $(a_{e_j}, a_{e_j+1}, \ldots, a_{n-1})$ can be expressed as a linear combination of the entries $(a_0, a_1, \ldots, a_{e_j-1})$.*

Incorporating Theorems 1 and 2, it becomes clear that the rank of every codeword of a cyclic code $\mathcal{C}$ characterized by a single free transform domain component drawn from a q-cyclotomic coset of size e_j with the remaining transform domain components constrained to zero is equal to e_j. The first e_j columns of this codeword are linearly independent and the remaining $(n-e_j)$ columns can be expressed as linear combination of first e_j columns.

Thus, all the non-zero codewords of this code $\mathcal{C}$ (which are $m \times n$ matrices over $GF(q)$) have rank equal to e_j. To ensure that the rank of the code is as large as possible, we choose a single free transform component from a full size q-cyclotomic coset of size m. The first m columns are linearly independent and the remaining $(n - m)$ columns are linearly dependent on the first m columns. The code can be punctured without a reduction in the rank by dropping the last $(n-m)$ columns of every codeword. Thus, we will henceforth work with the $m \times m$ full rank codeword matrices of the punctured cyclic code $\mathcal{C}$. Let us illustrate the process of code construction with a suitable example. Let A_j be the single free transform domain component drawn from a full size q-cyclotomic coset of size m. Making use of the definition of IGFFT, the time domain vector can be represented as,

$$\mathbf{a} = [A_j, \alpha^{-j}A_j, \alpha^{-2j}A_j, \ldots, \alpha^{-(n-1)j}A_j] \quad (4)$$

Here, $A_j \in GF(q^m)$ and α is the nth root of unity.

Each of the n coefficients of this vector can be represented as a n tuple over the field $GF(q)$. From Theorem-1, it follows that by arranging these n-tuples as columns we obtain an $m \times n$ full rank matrix over $GF(q)$.

This technique of constructing full rank cyclic codes can be used to synthesize binary as well as non-binary cyclic codes [13]. We have used this technique to synthesize full rank codes over $GF(5)$, $GF(7)$, $GF(13)$ and $GF(17)$. These full rank codes have been used to derive suitable SM-NSTBC designs.

The full rank code with $m \times m$ matrices over $GF(q)$ can be transformed into an equivalent collection of full rank $m \times m$ matrices over the complex number field by the use of suitable rank-preserving maps. Two such maps, namely the Gaussian Integer map and the Eisenstein Integer map are described in literature. The Gaussian Integer map is applicable only for prime integers of the form $4K + 1$, where K is an integer. The Eisenstein map is applicable only for prime integers of the form $6K + 1$, where K is an integer. Hence, conventional signal constellations where the number of symbols is a power of 2 cannot be employed in these schemes. The following section gives brief insight about the mapping scheme employed in the design of proposed SM-NSTBC.

2.1. Mapping technique employed for SM-NSTBC

Gaussian and Eisenstein mapping techniques are bijective, isometric and rank preserving. Lusina et al. in [16] proposed the use of rank preserving maps for finite fields, which could retain the full rank property from $GF(q)$ to a complex signal set in designing of STBCs. The mapping was carried by the use of Gaussian Integers ($q = 1 \; mod \; 4$). A prime number q which is in the form of $q = 4K + 1$, can be expressed as $q = u^2 + v^2$ for suitable u and v. These are known as Gaussian integers which are represented by $\omega = u + iv$; $u, v \in Z$ [15]. This map is defined as follows:

$$\mathcal{Z}_i = i \; mod \; \Pi = i - \left[\frac{i\Pi'}{\Pi\Pi'}\right]\Pi; \qquad i = 0, 1 \ldots, q-1 \quad (5)$$

Here, $\Pi = (u+iv)$ and $\Pi' = (u-iv)$, $[\cdot]$ stands for rounding of operation to the nearest integer. The values of ω for $q = 5, 13, 17$ is tabulated in Table 1.

A similar interpretation for Eisenstein–Jacobi integers was given by Huber [14], where a rank preserving map for $GF(q)$, (q is a prime) which can be expressed as $q = 1 \; mod \; 6$ was given. The Eisenstein–Jacobi integer can be expressed as $\omega = \alpha + \rho\beta$, where ρ is a complex number given by $\rho = (-1 + i\sqrt{3})/2$, the Eisenstein–Jacobi prime is given by $q = \alpha^2 + 3\beta^2$[19]. The mapping for Eisenstein map is given as following:

$$\zeta(i) = i \mod \Pi \triangleq i - \left[\frac{i\Pi^*}{\Pi\Pi^*}\right]\Pi \; for \; i = 0, 1, 2, \ldots, q-1 \quad (6)$$

Where $\Pi = \alpha + \beta + \rho 2\beta$ and $\Pi^* = \alpha + \beta + \rho^2 2\beta$. The values $GF(q)$ represented using the Gaussian and Eisenstein integer maps for certain values of q are listed in Table 1.

Table 1
Gaussian and Eisenstein mapping values.

q	Π	u	v	q	Π	α	β
5	2+i	−1	1+i	7	$3+2\rho$	2	1
13	3+2i	−2	1+2i	13	$3+4\rho$	1	2
17	4+i	−2	2+i	19	$5+2\rho$	4	1

Table 2
Values of q, m, n.

q	m	$n \mid q^m - 1$
5	4	$13 \mid 5^4 - 1$
7	4	$5 \mid 7^4 - 1$
13	4	$5 \mid 13^4 - 1$
17	4	$5 \mid 17^4 - 1$

Employing these two dimensional complex map of Gaussian mapping (for $GF(5)$, $GF(13)$, $GF(17)$) or Eisenstein map (for $GF(7)$) [20,21], the full rank cyclic codewords are constructed for the proposed SM-NSTBC. We now describe the system model of the proposed SM-NSTBC scheme. Note: In SM systems, inactive state of the antenna is represented by symbol zero. In order to avoid ambiguity between the symbol zero of the codeword and zero which represents the inactive state of antenna, we have mapped symbol zero to a non-zero prime number from Gaussian or Eisenstein integers (symbol zero can take values $\mathcal{Z}_{2+i}$, $\mathcal{Z}_{3+2i}$, $\mathcal{Z}_{4+i}$ for Gaussian map over $GF(5)$, $GF(13)$ and $GF(17)$, $\zeta_{3+2\rho}$ for Eisenstein map over $GF(7)$).

3. SM-NSTBC system model

Fig. 1, shows the block diagram of the proposed SM-NSTBC scheme. In the transmitter section, the information bits are encoded by a full rank n-length cyclic code $\mathcal{C}$ over $GF(q^m)$ which yields full rank $m \times m$ codeword matrices after puncturing. Since, we assume that the incoming bit stream is binary, the number of bits that can be associated with a codeword is a power of 2. However, the number of codewords is a prime power q^m which is not necessarily a power of 2. Hence, we map $\lfloor \log_2(q^m) \rfloor$ bits to a single codeword of $\mathcal{C}$ over $GF(q)$.

In the following discussion, we have derived two SM-NSTBC schemes which differ in the number of active antennas in use over a given time slot. The number of active transmit antennas $(N_a) = 2, 4$ for case 1 and case 2 respectively. We have employed cyclic codes over $GF(5^4)$, $GF(7^4)$, $GF(13^4)$ and $GF(17^4)$. Thus the codewords of all the codes are $4 \times n$ matrices over different values of n, $n|q^m - 1$. The various values of q, m and n employed in our constructions is specified in Table 2.

Case 1: Two active antenna combination
The number of active antennas are 2 for a given time slot, so the minimum number of transmit antennas required is 3 or more. The elements of the codeword matrix are conveyed column by column. The 4×4 matrix is viewed as a set of four 4×1 vectors. Each column is then divided into two parts, upper two symbols

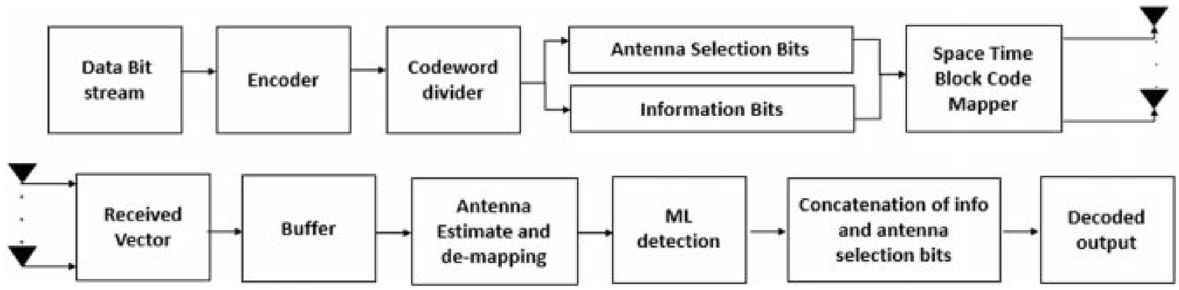

Fig. 1. Block diagram of the proposed SM-NSTBC scheme.

are used for antenna selection. These two symbols are responsible for selecting the two antennas which will radiate during this time slot. The third and the fourth symbols of the column are now radiated using the selected antennas. These lower two symbols are transmitted after modulation using Gaussian or Eisenstein mapping from the activated transmit antennas. This process is then repeated for the remaining columns. Thus, $\lfloor \log_2 (q^m) \rfloor$ bits ($m = 4$) of information are transmitted by the use of one codeword over 4 time slots.

Here, every column of the codeword matrix selects the pair of active transmit antenna individually. Thus each column of the codeword may be transmitted through a different set of active transmit antennas. One example of a possible transmitted codeword with $N_t = 4$ is illustrated in Eq. (7)

$$\begin{bmatrix} 0 & 0 & \Im\{a_{2,2}\} & \Im\{a_{2,3}\} \\ \Im\{a_{2,0}\} & \Im\{a_{2,1}\} & 0 & 0 \\ \Im\{a_{3,0}\} & 0 & \Im\{a_{3,2}\} & 0 \\ 0 & \Im\{a_{3,1}\} & 0 & \Im\{a_{3,3}\} \end{bmatrix} \tag{7}$$

The 0s in each column of the matrix denote inactive antennas. Let us examine the first column of Eq. (7). The first and the last entries are 0. This implies that antennas 1 and 4 are inactive during this time slot. The antenna selection bits have chosen antennas 2 and 3 as the active antennas and the information symbols $a_{2,0}$ and $a_{3,0}$ are conveyed through these antennas after suitable Gaussian or Eisenstein mapping. The total number of ways in which two active antennas can be selected from N_t transmit antennas is $\binom{N_t}{2}$, while two symbols from $GF(q)$ can address as many as q^2 different combinations. For small values of N_t, this can represent a many to one mapping. To avoid ambiguity in antenna selection, we have constructed a look up table which maps all possible q^2 values to appropriate antenna pairs. Since, the number of transmit antennas is fixed, and the values of q^2 are relatively large, many two tuples of symbols from $GF(q)$ will point to the same active transmit antenna pair. However, it must be kept in mind that at most $\left\lfloor q^2/\binom{N_t}{2} \right\rfloor + 1$ symbols pointing to a given antenna pair. After an entire SM-NSTBC codeword is received, we employ ML decoding to identify the transmitted symbols as well as the antenna selection symbols.

We have incorporated constellation rotation by θ in order to have two distinct constellations, the value of θ is selected to have maximum Euclidean distance between the two constellation points.

The algorithm given below describes the encoding procedure for SM-NSTBC with $N_a = 2$.

Algorithm 1 SM-NSTBC encoding algorithm for $N_a = 2$

Start
Consider $2^{\lfloor \log_2 (q^m) \rfloor}$ bits
Step 1:Segregation into $\mathcal{M}_{m\times m}$ full rank matrices
Step 2: Obtain
  $\mathcal{M}_{m\times m} \rightarrow [\mathcal{M}_1]_{m/2\times m}$ antenna selection symbols + $[\mathcal{M}_2]_{m/2\times m}$ information symbols
Step 3:for count=1:number of time slots for full codeword transmission
Step 4: Antenna selection
  $[\mathcal{M}_1]_{m/2\times m} \rightarrow \mathbf{C}_1[a(1)a(2)\cdots a(2m)]$
Step 5: Mapping of Information symbols
  $\mathbf{C}_2[i(1)i(2)\cdots i(2m)]$
  Define map $\varphi : [\mathcal{M}_2]_{m/2\times m} \rightarrow \Im(\mathbf{C}_2) * e^{j\theta}$
  Obtain suitable rank preserving map $X = \varphi(\mathbf{C}_2)$
end
Step 6: Obtain X_{SM}

Here, $\mathbf{C}_1$ represents the antenna selection, and $\mathbf{C}_2$ represent information symbol mapping. The total number of transmit antenna considered here are $N_t = 4$ and $N_t = 6$. Example 1, gives a brief description of SM-NSTBC case 1 implementation.

Example 1. Consider $N_t = 4, N_a = 2$ and $N_r = 4$. The possible antenna pair combinations are $\binom{N_t}{2}$ in number. Let the codeword matrix over $GF(q)$ be represented as,

$$\begin{bmatrix} x_{0,0} & x_{0,1} & x_{0,2} & x_{0,3} \\ x_{1,0} & x_{1,1} & x_{1,2} & x_{1,3} \\ x_{2,0} & x_{2,1} & x_{2,2} & x_{2,3} \\ x_{3,0} & x_{3,1} & x_{3,2} & x_{3,3} \end{bmatrix} \tag{8}$$

A brief description of the encoding procedure is presented. Consider the first column, $[x_{0,0}\ x_{1,0}\ x_{2,0}\ x_{3,0}]^T$. The first two symbols, $x_{0,0}$ and $x_{1,0}$ are used to select the transmit antenna pair and the remaining two symbols, $x_{2,0}$ and $x_{3,0}$ are mapped into suitable complex symbols by using the Gaussian or Eisenstein map. These symbols are then radiated by using the active antenna pair selected for that time slot. This process is repeated for the remaining three time slots as well.

Case 2: Four active antenna combination

In case 2, we have turned our attention to systems possessing a large number transmit antennas in which four antennas (out of N_t) are active over any given time slot. We have designed SM-NSTBC schemes for six and eight transmit antennas. We employ the same codeword structure as in Section 2. Here, a total of four transmit antennas are activated over every time slot. This essentially lead to transmission of four symbols in a single time slot. The codeword matrix is divided into set of two sub-matrices instead of four (as considered in case 1). Here a total of eight bits are transmitted in a single time slot, where four bits are conveyed by antenna selection (as transmit antenna index) while other four bits are transmitted in the form of information bits. This is given below:

$$\begin{bmatrix} x_{0,0} & x_{0,1} & x_{0,2} & x_{0,3} \\ x_{1,0} & x_{1,1} & x_{1,2} & x_{1,3} \\ x_{2,0} & x_{2,1} & x_{2,2} & x_{2,3} \\ x_{3,0} & x_{3,1} & x_{3,2} & x_{3,3} \end{bmatrix} \rightarrow \begin{bmatrix} x_{0,0} & x_{0,2} \\ x_{1,0} & x_{1,2} \\ x_{0,1} & x_{0,3} \\ x_{1,1} & x_{1,3} \\ x_{2,0} & x_{2,2} \\ x_{3,0} & x_{3,2} \\ x_{2,1} & x_{2,3} \\ x_{3,1} & x_{3,3} \end{bmatrix} \tag{9}$$

Two time slots are required to transmit the whole codeword matrix. The total number of distinct antenna combinations is $\binom{N_t}{4}$, while four tuples of symbols over $GF(q)$ can address as many as q^4 different antenna combinations. In the first time slot the symbols $[x_{0,0}\ x_{1,0}\ x_{0,1}\ x_{1,1}\ x_{2,0}\ x_{3,0}\ x_{2,1}\ x_{3,1}]^T$ are employed with first four symbols $[x_{0,0}\ x_{1,0}\ x_{0,1}\ x_{1,1}]$ being used for antenna selection and the remaining four symbols $[x_{2,0}\ x_{3,0}\ x_{2,1}\ x_{3,1}]$ being radiated through the selected antennas.

The transmission process employed in Case 2 is summarized by the algorithm 2.

Algorithm 2 SM-NSTBC encoding algorithm for $N_a = 4$

1: Start
2: Consider $2^{\lfloor \log_2 (q^m) \rfloor}$ bits
3: Step 1:Segregation into $\mathcal{M}_{m\times m}$
4: Step 2:Obtain modified matrix:
5: $\mathcal{M}_{m\times m} \rightarrow [\mathcal{M}]_{2m\times m/2}$
6: Step 3:Obtain Full rank matrix
7: $[\mathcal{M}_1]_{2m\times m/2} \rightarrow [\mathcal{M}_1]_{m\times m/2}$ antenna selection symbols + $[\mathcal{M}_2]_{m\times m/2}$ information symbols
8: Step 4:for count=1:number of time slots for full codeword transmission
9: Step 5:Antenna selection
10: $[\mathcal{M}_1]_{m\times m/2} \rightarrow \mathbf{C}_1[a(1)a(2)\cdots a(2m)]$
11: Step 6:Mapping Information symbols
12: $[\mathcal{M}_2]_{m\times m/2} \rightarrow \mathbf{C}_2[i(1)i(2)\cdots i(2m)]$
13: Obtain suitable rank preserving map $X = \varphi(\mathbf{C}_2)$
14: end
15: Step 7: Obtain X_{SM}

3.1. Spectral efficiency

The spectral efficiency for the proposed SM-NSTBC scheme is given by

$$\eta = \frac{\lfloor \log_2 (q^m) \rfloor}{n_s} \tag{10}$$

Here, the value of $n_s = 2m/N_a$, which corresponds to the time slots required to transmit the SM-NSTBC codeword, N_a represents the number of active antennas, q is a prime ($q = 5, 7, 13, 17$), m is the order of the extension field ($m = 4$).

3.2. ML Decoding for SM-NSTBC

In this section, the detailed discussion about the Maximum likelihood (ML) detection strategy employed is outlined. Let the total number of time slots required to transmit a full codeword matrix be specified as n_s. At the receiver the information bits are buffered till the full codeword is obtained. The received $N_r \times n_s$ array is given by,

$$\mathbf{Y} = \mathbf{H}\mathbf{X}_{SM} + n \tag{11}$$

Here, $\mathbf{Y}$ represents the received array, $\mathbf{X}_{SM}$ is the transmitted codeword matrix, $\mathbf{H}$ is the channel matrix and n is a realization of circularly symmetric complex independent and identically distributed Gaussian variable $C\mathcal{N}(0, 1)$.

Once the full codeword is obtained, the active antenna index is extracted and the ML decoding of the received array is performed. This procedure is summarized by Eq. (12).

$$\hat{\mathbf{X}}_{SM} = argmin \left\| \mathbf{Y} - \mathbf{H}\tilde{\mathbf{X}}_{SM} \right\|_F^2 \tag{12}$$

Here, $\hat{\mathbf{X}}_{SM}$ is the estimated codeword.

Algorithm 3 Decoding Algorithm

1: Start
2: Input: $\mathbf{Y}$
3: Output: $\hat{\mathcal{M}}$
4: Step1: Obtain the count of Slot
5: Step 2:For count=1:length of slot
6: Step 3: Obtain the receiver array
7: Step 4:Compute $\hat{\mathbf{X}}_{SM} = argmin \left\| \mathbf{Y} - \mathbf{H}\tilde{\mathbf{X}} \right\|_F^2 \forall \tilde{\mathbf{X}} \in \mathbf{C}$
8: Step 5: Obtain $\hat{\mathbf{X}}_{SM}$

4. Analytical performance of the proposed scheme

In this section, we give a detailed description about the error performance of proposed SM-NSTBC system. An analytical upper bound on the Average Bit Error Rate (ABER) has been derived to verify the correctness of Monte Carlo simulation performed. A theoretical upper bound on ABER is given by [22–24] and [25].

$$ABER_{SM-NSTBC} \leq \frac{1}{2^{\lfloor \log_2 q^m \rfloor}} \sum_{i=1}^{2^{\lfloor \log_2 q^m \rfloor}} \sum_{j=1}^{2^{\lfloor \log_2 q^m \rfloor}} \frac{N(\mathbf{X}_{SM} - \hat{\mathbf{X}}_{SM})}{2\eta} \times (P(\mathbf{X}_{SM} \rightarrow \hat{\mathbf{X}}_{SM})) \tag{13}$$

Here, η is the spectral efficiency of the SM-NSTBC scheme, $N(\mathbf{X}_{SM} - \hat{\mathbf{X}}_{SM})$ is the number of non-zero elements of the difference matrix of $\mathbf{X}_{SM} - \hat{\mathbf{X}}_{SM}$, and $P(\mathbf{X}_{SM} \rightarrow \hat{\mathbf{X}}_{SM})$ is given below

$$P(\mathbf{X}_{SM} \rightarrow \hat{\mathbf{X}}_{SM}) = Pr\left(\|\mathbf{Y} - \mathbf{H}\mathbf{X}_{SM}\|_F^2 > \left\| \mathbf{Y} - \mathbf{H}\hat{\mathbf{X}}_{SM} \right\|_F^2 |\mathbf{H} \right) \tag{14}$$

here, for a given channel $\mathbf{H}$, $(P(\mathbf{X}_{SM} \rightarrow \hat{\mathbf{X}}_{SM}))$ is the pairwise error probability (PEP) of detecting $\hat{\mathbf{X}}_{SM}$, when $\mathbf{X}_{SM}$ is transmitted. The pairwise error probability is evaluated using [23–25]:

$$P(\mathbf{X}_{SM} \rightarrow \hat{\mathbf{X}}_{SM}) = Q\left(\sqrt{\frac{\rho}{2}} \left\| \mathbf{H}(\mathbf{X}_{SM} - \hat{\mathbf{X}}_{SM}) \right\| \right) \tag{15}$$

Here $Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} \exp(-y^2/2)dy$. The unconditional PEP is obtained by the use of moment generating function [23].

$$P(\mathbf{X}_{SM} \rightarrow \hat{\mathbf{X}}_{SM}) \leq \frac{1}{\pi} \int_0^{\frac{\pi}{2}} \prod_{l=1}^{L} \left(\frac{1}{1 + \frac{\rho\lambda_{i,j_l}}{4 sin^2\theta}} \right)^{N_r} d\theta \tag{16}$$

here, $L = 4$ and λ_{i,j_l} are the Eigen values of the distance matrix $(\mathbf{X}_{SM} - \hat{\mathbf{X}}_{SM})^H (\mathbf{X}_{SM} - \hat{\mathbf{X}}_{SM})$. Further PEP can be calculated as given in Eq. (17)

$$P(\mathbf{X}_{SM} \rightarrow \hat{\mathbf{X}}_{SM}) =$$

$$\frac{1}{\pi} \int_0^{\frac{\pi}{2}} \left(\frac{1}{1 + \frac{\rho\lambda_{i,j_1}}{4\sin^2\theta}} \right)^{N_r} \left(\frac{1}{1 + \frac{\rho\lambda_{i,j_2}}{4\sin^2\theta}} \right)^{N_r}$$

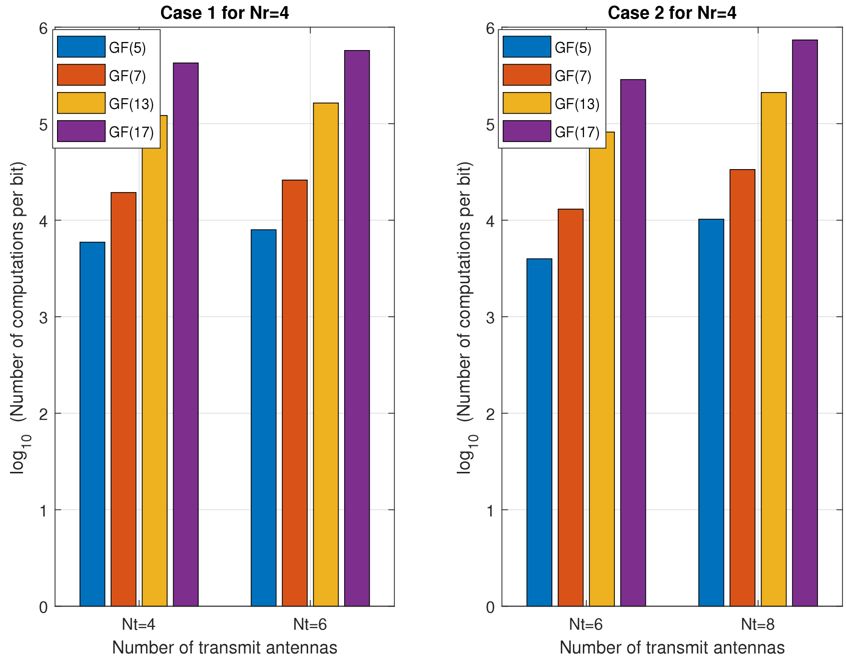

Fig. 2. Computation complexity of ML decoder for different transmit antenna configurations.

$$\times \left(\frac{1}{1+\frac{\rho\lambda_{i,j_3}}{4\sin^2\theta}}\right)^{N_r}\left(\frac{1}{1+\frac{\rho\lambda_{i,j_4}}{4\sin^2\theta}}\right)^{N_r} d\theta \tag{17}$$

Let $c_l = \frac{\rho\lambda_{i,j_l}}{4}$. The closed form solution of Eq. (17) can be obtained from [23], and is simplified as shown in Eq. (18).

$$P(\mathbf{X}_{SM} \to \hat{\mathbf{X}}_{SM}) \le \frac{1}{2}\prod_{l=1}^{L}\frac{1}{(1+c_l)^{N_r}} \tag{18}$$

Substituting Eq. (18) in Eq. (13), the analytical upper bound on the Average Bit Error Rate (ABER) is calculated as given in Eq. (19).

$$ABER_{SM-NSTBC} \le \frac{1}{2^{\lfloor \log_2 q^m \rfloor}} \sum_{i=1}^{2^{\lfloor \log_2 q^m \rfloor}} \sum_{j=1}^{2^{\lfloor \log_2 q^m \rfloor}} \frac{N(\mathbf{X}_{SM} - \hat{\mathbf{X}}_{SM})}{4\eta} \times \prod_{l=1}^{L}\frac{1}{(1+c_l)^{N_r}} \tag{19}$$

5. Complexity analysis for the proposed SM-NSTBC

In this section, we estimate the complexity of the proposed ML decoder. The computation complexity is given in terms of total complex multiplications. For an antenna selection decoding the total number of iterations required are $\binom{N_t}{N_a} n_s$. Now consider the ML decoding as $\left\|\mathbf{Y} - \mathbf{H}\tilde{\mathbf{X}}\right\|_F^2 \in \mathbf{C}$ the total number of complex multiplication required for estimating $\mathbf{H}\tilde{\mathbf{X}}$ is given by $(N_r N_t n_s)$. For estimating the Frobenius norm, the total number of complex multiplications required are $(N_r n_s)$. So the total number of complex multiplications required for ML decoder is evaluated as follows:

$$2^{\lfloor \log_2 q^m \rfloor}\left(\binom{N_t}{N_a} n_s + (N_r N_t n_s) + (N_r n_s)\right) \tag{20}$$

The total number of complex additions required are

$$2^{\lfloor \log_2 q^m \rfloor}\left((N_r n_s(N_t+1)) + (N_r n_s - 1)\right) \tag{21}$$

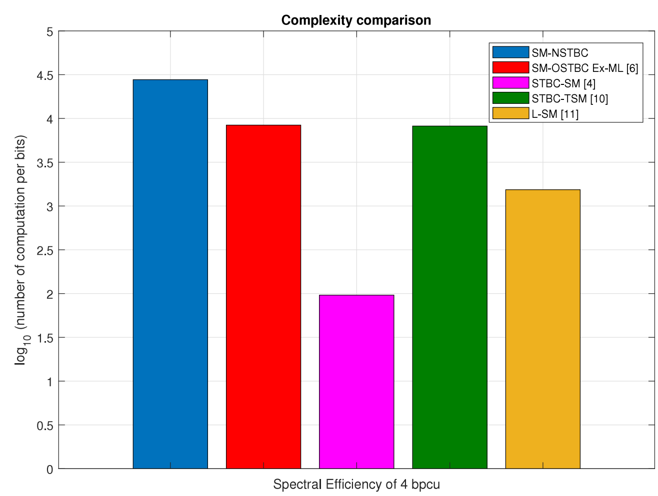

Fig. 3. Comparison of the computation complexity of SM-NSTBC, SM-OSTBC, STBC-SM, STBC-TSM and L-SM.

Fig. 2, shows the computational complexity of the proposed schemes which employ ML detection for two and four active antenna combinations. For this comparison we have used $N_r = 4$. The computational complexity increases if the number of possible antenna combination increases or as the size of the constellation increases. Fig. 3 gives the comparison of computational complexity of the proposed SM-NSTBC scheme, with other competing schemes namely, SM-OSTBC [6], STBC-SM [4], STBC-TSM [10] and L-SM [11] techniques. Here, the number of receive antenna for all the schemes is 1 and the spectral efficiency obtained is 4 bpcu.

6. Simulation results

This section demonstrates the simulation results of the proposed SM-NSTBC scheme for different values of N_t and N_a. The proposed SM-NSTBC is represented as $C(N_t, N_r, N_a)$ for all simulations. The proposed scheme has been compared with the variants of STBC-SM and SM systems. All performance comparisons are

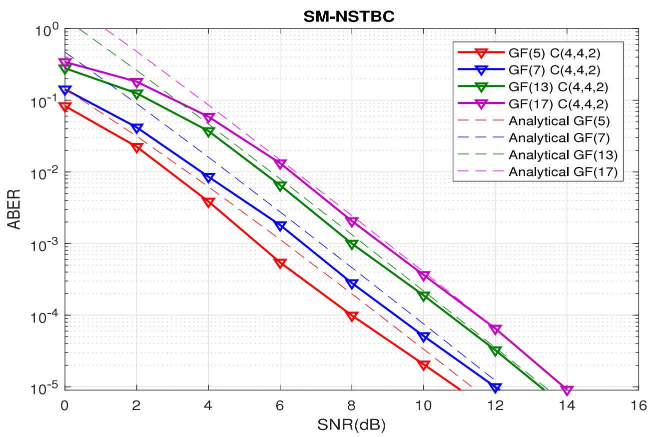

Fig. 4. ABER performance of SM-NSTBC for ($N_t = 4, N_r = 4, N_a = 2$) over $q = 5, 7, 13, 17$.

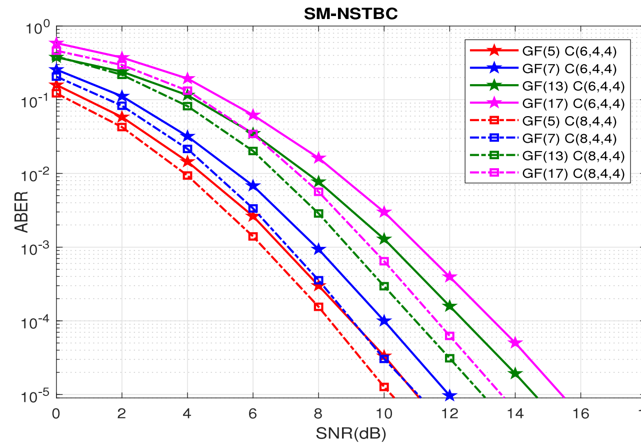

Fig. 6. ABER performance of SM-NSTBC for $q = 5, 7, 13, 17$ for ($N_t = 6, N_r = 4, N_a = 4$) and ($N_t = 8, N_r = 4, N_a = 4$).

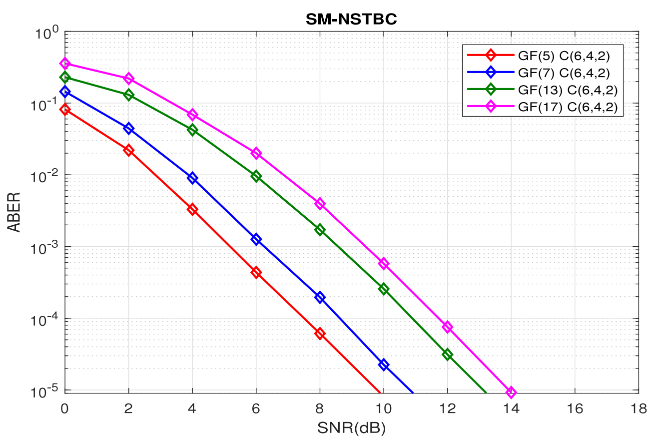

Fig. 5. ABER performance of SM-NSTBC for $q = 5, 7, 13, 17$ for ($N_t = 6, N_r = 4, N_a = 2$).

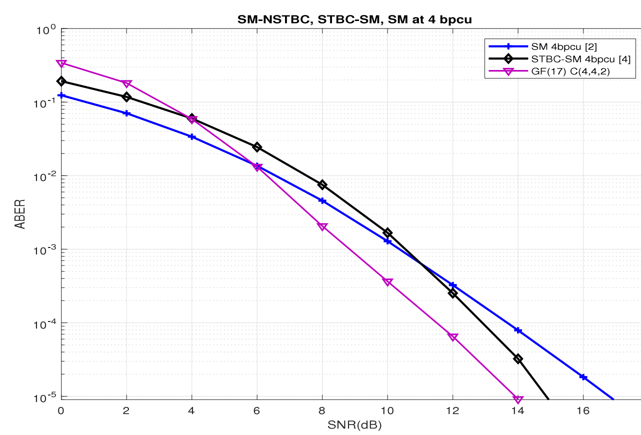

Fig. 7. ABER performance comparison of SM-NSTBC, STBC-SM [4] and SM [2] at 4 bpcu.

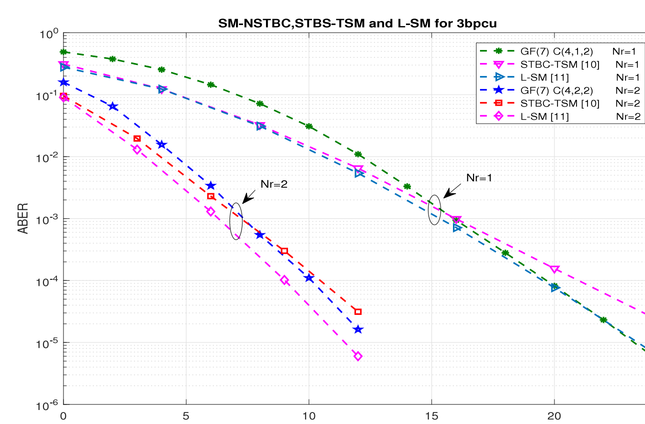

Fig. 8. ABER performance comparison of SM-NSTBC, STBC-TSM [10] and L-SM [11] at 3 bpcu.

carried out for a ABER value of 10^{-5}. It has been observed that for a considered spectral efficiency the proposed SM-NSTBC scheme outperforms other STBC-SM schemes and SM schemes by ∼1.5 dB to ∼5 dB over a Rayleigh fading channel.

In Fig. 4, the ABER curves of SM-NSTBC with $N_t = 4$, $N_r = 4$ and the active antennas are equal to 2 as specified in Case-1 are presented. A theoretical upper bound is shown in order to verify the correctness of the Monte-Carlo simulations. A close correspondence between the Analytical upper bound and simulation results is observed at higher SNR values. Simulation results are demonstrated for codewords obtained over $GF(5)$, $GF(7)$, $GF(13)$ and $GF(17)$. The spectral efficiency achieved for a system which utilizes $C(N_t = 4, N_r = 4, N_a = 2)$ is 2.322 bpcu for $q = 5$, 2.807 bpcu for $q = 7$, 3.701 bpcu for $q = 13$, and 4.087 bpcu is obtained for $q = 17$.

Fig. 5, shows a SM-NSTBC system which employs 6 transmit antennas, 4 receive antennas and 2 active antenna combinations represented as $C(6, 4, 2)$. The simulation results demonstrate the ABER performance for $q = 5, 7, 13, 17$ with similar spectral efficiencies as indicated for $C(4, 4, 2)$ system.

Fig. 6 demonstrates the simulation results for a system which employs $C(6, 4, 4)$ and $C(8, 4, 4)$ combinations as indicated in Case-2. These systems are evaluated over $GF(q)$. The corresponding values of $q = 5, 7, 13, 17$ yields a spectral efficiency of 4.643 bpcu, 5.614 bpcu, 7.4 bpcu and 8.175 bpcu respectively for a ABER of 10^{-5}.

Fig. 7, gives the comparison plot of SM-NSTBC with STBC-SM [4] and SM [2] systems. The proposed SM-NSTBC which uses $C(4, 4, 2)$ with codewords derived from $GF(17)$ achieves a BER performance improvement of ∼1.5 dB over STBC-SM and ∼3 dB over SM systems for a spectral efficiency of 4 bpcu. SM-NSTBC produces a fractional spectral efficiency of 4.087 bpcu.

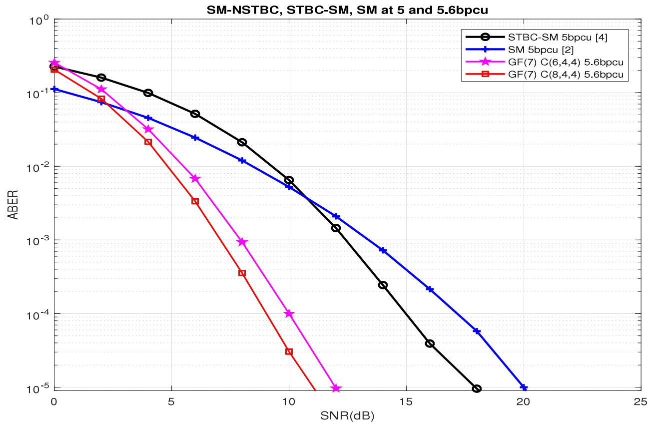

Fig. 9. ABER performance comparison of SM-NSTBC, STBC-SM [4] and SM [2] at 5 bpcu. SM-NSTBC has a higher spectral efficiency of 5.6 bpcu.

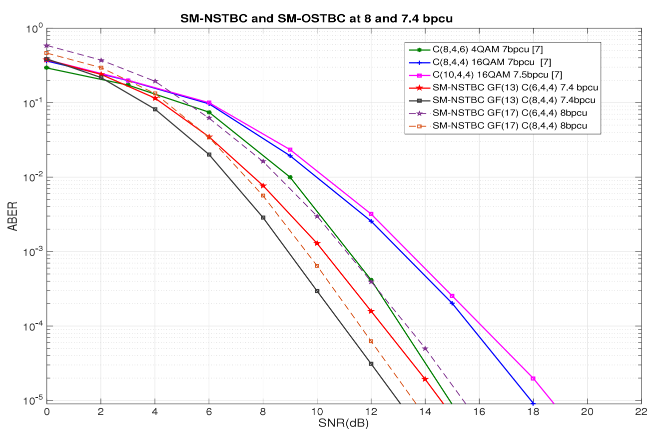

Fig. 10. ABER performance comparison of SM-NSTBC, SM-OSTBC [7] systems. (For interpretation of the references to color in this figure legend, the reader is referred to the web version of this article.)

In Fig. 8 performance of the proposed SM-NSTBC is compared with STBC-TSM [10] and L-SM [11] schemes. All schemes employ four transmit antennas, two active antennas, and $N_r = 1$ and 2. By the use of these antenna configurations STBC-TSM and L-SM achieves a spectral efficiency of 3 bpcu, while SM-NSTBC has spectral efficiency of 2.8 bpcu. It is observed that for $N_r = 2$, at very low SNR values STBC-TSM and L-SM give a better performance compared to the proposed scheme. At higher values of SNR, SM-NSTBC outperforms STBC-TSM by ~1 dB, while L-SM gives better performance as compared to SM-NSTBC. However, for $N_r = 1$, at high values of SNR, SM-NSTBC outperforms both STBC-TSM and L-SM schemes as shown in Fig. 8.

Fig. 9, shows the BER comparison of the SM-NSTBC, STBC-SM [4] and SM [2] schemes for a spectral efficiency of 5 bpcu. The proposed SM-NSTBC with codewords derived over $GF(7)$, $C(6,4,4)$ and $C(8,4,4)$ produces a spectral efficiency of 5.615. It is observed that the proposed scheme shows significant improvement of ~6 dB over STBC-SM and ~7.5 dB over SM systems.

Fig. 10, delineates the simulation results of SM-NSTBC and SM-OSTBC [7] systems. The STBC-SM [4] and SM [2] systems cannot be compared with this plot, since, STBC-SM and SM systems will not provide four active antenna combinations. It is observed that the proposed SM-NSTBC scheme with $C(8,4,4)$ over $GF(13)$ (produces 7.4 bpcu) shows an improvement in the BER performance of ~2 dB (Black Line), over SM-OSTBC system which uses $C(8,4,6)$ employing 4-QAM constellations to obtain a spectral efficiency of 7 bpcu (Green Line). Similarly, the proposed SM-NSTBC with $C(8,4,4)$ (Brown dotted line) over $GF(17)$ scheme (produces 8 bpcu) shows an improvement in the BER performance of ~6 dB, over SM-OSTBC system which uses $C(10,4,4)$ employing 16-QAM constellations (magenta) to obtain a spectral efficiency of 7.5 bpcu.

7. Conclusions

A new SM-NSTBC scheme employing a full rank cyclic code over $GF(q^m)$ has been proposed. These codes have codewords which can be viewed as full rank $m \times n, (m \leq n)$ matrices over $GF(q)$. By suitable puncturing, the codeword matrices are transformed into full rank $m \times m$ matrices over $GF(q)$. These $m \times m$ matrices are mapped into codewords of a SM-NSTBC using a suitable rank preserving map (Gaussian or Eisenstein map). The desired spectral efficiency can be achieved by selecting appropriate number of active antennas and order of Galois Field ($q = 5, 7, 13, 17$) over which codewords are constructed. An analytic upper bound on the ABER has been determined. Monte-Carlo simulations have been carried out to quantify the performance of these codes over Rayleigh fading channels. A close correspondence between the upper bound values and simulation results is observed. It is seen that proposed SM-NSTBC achieves an improvement of ~1.5 dB to ~5 dB over STBC-SM and SM-OSTBC schemes.

Declaration of competing interest

The authors declare that they have no known competing financial interests or personal relationships that could have appeared to influence the work reported in this paper.

References

[1] E. Telatar, Capacity of multi-antenna Gaussian channels, Eur. Trans. Telecommun. 10 (6) (1999) 585–595.
[2] R. Mesleh, H. Haas, S. Sinanovic, C.W. Ahn, S. Yun, Spatial modulation, IEEE Trans. Veh. Technol. 57 (4) (2008) 2228.
[3] V. Tarokh, N. Seshadri, A. Calderbank, Space-time codes for high data rate wireless communication: Performance criterion and code construction, IEEE Trans. Inf. Theory 44 (2) (1998) 744–765.
[4] E. Basar, U. Aygolu, E. Panayirci, H.V. Poor, Space-time block coded spatial modulation, IEEE Trans. Commun. 59 (3) (2011) 823–832.
[5] S. Sugiura, S. Chen, L. Hanzo, Coherent and differential space-time shift keying: A dispersion matrix approach, IEEE Trans. Commun. 58 (11) (2010) 3219–3230.
[6] M.-T. Le, V.D. Ngo, H.-A. Mai, X.N. Tran, M. D Renzo, Spatially modulated orthogonal space-time block codes with non-vanishing determinants, IEEE Trans. Commun. 62 (1) (2014) 85–99.
[7] L. Wang, Z. Chen, Correction to "Spatially modulated orthogonal space-time block codes with non-vanishing determinants", IEEE Trans. Commun. 62 (10) (2014) 3723–3724.
[8] L. Wang, Z. Chen, X. Wang, A space-time block coded spatial modulation from (n, k) error correcting code, IEEE Wirel. Commun. Lett. 3 (1) (2014) 54–57.
[9] X. Li, L. Wang, High rate space-time block coded spatial modulation with cyclic structure, IEEE Commun. Lett. 18 (4) (2014) 532–535.
[10] A.G. Helmy, M. Di Renzo, N. Al-Dhahir, Enhanced-reliability cyclic generalized spatial-and-temporal modulation, IEEE Commun. Lett. 20 (12) (2016) 2374–2377.
[11] S.A. Alkhawaldeh, Layered spatial modulation, EURASIP J. Wireless Commun. Networking 2018 (1) (2018) 233.
[12] U. Sripati, B. Rajan, On the rank distance of cyclic codes, in: IEEE International Symposium on Information Theory, 2003. Proceedings, IEEE, 2003, p. 72.
[13] U. Sripati, B.S. Rajan, V. Shashidhar, Full-diversity STBCs for block-fading channels from cyclic codes, in: Global Telecommunications Conference, 2004. GLOBECOM'04. IEEE, Vol. 1, IEEE, 2004, pp. 566–570.
[14] K. Huber, Codes over eisenstein-jacobi integers, Contemp. Math. 168 (1994) 165.
[15] K. Huber, Codes over Gaussian integers, IEEE Trans. Inform. Theory 40 (1) (1994) 207–216.

[16] P. Lusina, E. Gabidulin, M. Bossert, Maximum rank distance codes as space-time codes, IEEE Trans. Inform. Theory 49 (10) (2003) 2757–2760.
[17] T.K. Moon, Error Correction Coding: Mathematical Methods and Algorithms, Vol. 750, John Wiley & Sons, 2005, p. 750.
[18] U. Sripati, Space time block codes for MIMO fading channels from codes over finite field (Ph.D. thesis), IISc Bangalore, India, 2004.
[19] S. Puchinger, S. Stern, M. Bossert, R.F. Fischer, Space-time codes based on rank-metric codes and their decoding, in: Wireless Communication Systems (ISWCS), 2016 International Symposium on, IEEE, 2016, pp. 125–130.
[20] G.D. Goutham Simha, M. Raghavendra, K. Shriharsha, U.S. Acharya, Signal constellations employing multiplicative groups of Gaussian and eisenstein integers for enhanced spatial modulation, Phys. Commun. 25 (2017) 546–554.
[21] M. Raghavendra, G. Goutham Simha, U.S. Acharya, Non-orthogonal full rank space-time block codes over Eisenstein-Jacobi integers for MIMO systems, in: 2017 4th International Conference on Electronics and Communication Systems (ICECS), IEEE, 2017, pp. 83–87.
[22] J.G. Proakis, Digital Communications, third ed., McGraw-Hill, 1995.
[23] M.K. Simon, M.S. Alouini, Digital communication over fading channels, Vol. 95, John Wiley & Sons, 2005.
[24] E. Basar, U. Aygolu, E. Panayirci, H.V. Poor, Performance of spatial modulation in the presence of channel estimation errors, IEEE Commun. Lett. 16 (2) (2012) 176–179.
[25] G.D. Goutham Simha, Design and implementation of modulation and detection strategies for spatial modulation MIMO systems (Ph.D. thesis), NITK, Surathkal, India, 2018.

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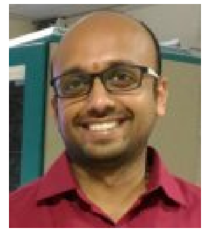

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