Generalized Space Time Block Coded Spatial Modulation for Open-Loop Massive MIMO Downlink Communication Systems

Lixia Xiao, *Member, IEEE*, Da Chen, *Member, IEEE*, Ibrahim Hemadeh, *Member, IEEE*, Pei Xiao, *Senior Member, IEEE*, and Tao Jiang, *Fellow, IEEE*

Abstract—In this paper, we propose a generalized space-time block coded spatial modulation (GSTBC-SM) scheme for open-loop massive multiple-input and multiple-output (MIMO) downlink communication systems. Specifically, we firstly partition the information bits into multiple groups with each group modulated by the spatial modulation (SM), where the SM symbols are invoked for orthogonal STBC (OSTBC) and quasi-orthogonal STBC (Q-OSTBC) structures. Then, message passing (MP) and block minimum mean square equalization (B-MMSE) detectors are designed for our GSTBC-SM systems, to achieve near-optimal performance with significantly reduced complexity in massive MIMO configurations. Finally, we derive the theoretical average bit error probability (ABEP) of the proposed scheme. The main contribution is that the propose scheme achieves high transmission rate and diversity gain even with small number of radio frequency (RF) chains at the transmitter. Simulation results verify the theoretical derivations and show that the proposed GSTBC-SM scheme provides near 20 dB gain over the conventional GSTBC scheme under massive MIMO configurations.

Index Terms—Spatial Modulation (SM), Space Time Block Coding (STBC), High throughput, Diversity gain.

I. INTRODUCTION

SPACE time block coding (STBC), which transmits identical data information via different space and time domains to attain a high diversity gain, has been considered as the key multiple input multiple output (MIMO) technique in wireless communications [1]. There are two classic categories of conventional STBC schemes, i.e., orthogonal STBC (OSTBC) [2]-[7] and quasi-orthogonal STBC (Q-OSTBC) schemes [8]-[14]. Specifically, Alamouti first proposed the OSTBC scheme with $N_t = 2$ transmit antennas (TAs) in [2], which has been widely used in multicarrier systems [3][4]. Then, the authors of [5]-[7] further extended the OSTBC concept into other numbers of TAs. Meanwhile, the Q-OSTBC schemes with a small number of TAs were developed in [8]-[14]. With the

requirements of high throughput and reliability for wireless communications, massive MIMO has been considered as the key technique in the next generation communication [15]-[17]. However, conventional STBC schemes are unsuitable for massive MIMO systems since they significantly complicate the code design and increase the cost of radio frequency (RF) chains under massive MIMO configurations. As a result, it is urgent to design new STBC schemes for massive MIMO systems to improve the throughput and diversity gain with low implementation cost.

Recently, several precoding based conventional STBC systems have been investigated for massive MIMO scenarios in [18]-[24]. Specifically, noncoherent massive space-time block codes is designed for uplink communications in [18], while the massive STBC systems of [19]-[24] are designed for downlink communications. In [19], the authors proposed a novel massive OSTBC scheme by embedding smaller OSTBCs into massive configurations at the expense of degraded performance. Then, the authors of [20]-[22] proposed omnidirectional OSTBC and Q-OSTBC for massive MIMO systems, which attain the diversity gain by employing the precoding concept. Based on the same concept, the authors of [23] proposed full-diversity STBC for massive MIMO. Finally, the authors of [24] presented the design criteria for the precoding based omnidirectional massive STBC systems.

However, the aforementioned schemes in [19]-[24] are designed for closed-loop massive MIMO downlink communication systems, which require the channel state information (CSI) at the transmitter and significantly increase the complexity. These massive STBC schemes also require numerous RF chains, resulting in substantial implementation cost and signal processing complexity. Furthermore, their common shortcoming is the low transmission rate. Therefore, it is meaningful to design a STBC scheme with high throughput for open-loop massive MIMO downlink communication systems.

With the development of massive MIMO, spatial modulation (SM), which implicitly transmits information relying on the activated antenna indices, has been considered as a promising low-cost single RF chain based open-loop massive MIMO candidate [25]-[30]. To further improve the diversity gain of SM, the STBC based SM (STBC-SM) scheme was proposed in [31]. Then, several variants of STBC-SM with high transmission rate and modified codeword were proposed in [32]-[35]. However, the common feature of these STBC-SM schemes is that the information bits are transmitted as

This work was supported in part by the National Key R&D Program of China with Grant number 2019YFB180003400, National Science Foundation of China with Grant number 61831013, 61631015, and Innovation Fund of WNLO.

L. Xiao is with the Wuhan National Laboratory for Optoelectronics, Huazhong University of Science and Technology, Wuhan 430074, China. (e-mail: lixiaoxiao@hust.edu.cn)

D. Chen, and T. Jiang are with Wuhan National Laboratory for Optoelectronics, School of Electronic Information and Communications, Huazhong University of Science and Technology, Wuhan 430074, China (e-mail: chen-da@hust.edu.cn; Tao.Jiang@ieee.org).

I. Hemadeh and P. Xiao are with 5G Innovation Center of University of Surrey, UK (e-mail: ibrahimhemadeh@gmail.com; p.xiao@surrey.ac.uk).

amplitude phase modulation (APM) symbols by one activated antenna combination (AC). In massive MIMO configurations, the selection process of AC becomes more complicated, which makes the bit-to-symbol mapping impractical. More importantly, the diversity gains of these schemes also become slight in the context of massive MIMO setup. In order to address the above issues, we proposed a novel STBC-SM in [36], where the SM symbols were taken as a whole vector to invoke for the Alamouti coding, which significantly simplified the AC design and the bit-to-symbol mapping. We refer to this scheme as Alamouti-SM in this paper. Moreover, it is demonstrated in [36] that Alamouti-SM outperforms STBC-SM in [31] in terms of bit error rate (BER). Non-coherent STBC-SM schemes were also developed in [37]-[41]. However, the above mentioned STBC-SM schemes were mainly designed for two activated TAs. To make STBC-SM applicable for large number of activated TAs, the authors of [42] proposed the spatially modulated orthogonal space-time block codes, where four RF chains were considered at the transmitter and double space-time transmit diversity was achieved. However, it significantly complicated the space codeword design since the total number of space codewords was extremely large, i.e., 4^{N_t-2} at massive MIMO setups, where N_t is the number of TAs. Then, we proposed a generalized STBC-SM (GSTBC-SM) system [43] to further exploit the benefits of the generalized STBC structure and SM scheme. However, it only considered low transmission rate based OSTBC-SM schemes and was only suitable for small-scale MIMO.

Against the above background, the contributions of this paper are summarized as follows.

- 1) We propose a novel GSTBC-SM structure for open-loop massive MIMO downlink communication systems. In the proposed GSTBC-SM scheme, the information bits are partitioned into K groups with each group modulated by SM, which will be taken as a node for conventional STBC coding (i.e OSTBC, Q-OSTBC). Hence the transmitter is capable of obtaining a flexible transmission rate and diversity order depending on the specific conventional STBC structure and the number of TAs. More specifically, both OSTBC-SM and Q-OSTBC-SM schemes are studied in the context of massive MIMO configurations.
- 2) We propose a low-complexity message passing (MP) detector and a low-complexity block minimum mean square equalization (B-MMSE) detector for different GSTBC-SM systems, which are capable of approaching the maximum likelihood (ML) detector's performance with a significant complexity reduction. Furthermore, the proposed low-complexity detectors based GSTBC-SM are capable of providing substantial performance gains over the conventional GSTBC, SM family [47] and zero-forcing (ZF) precoded MIMO [48] systems in the context of massive MIMO configurations.
- 3) Both the accurate and approximate average bit error probability (ABEP) are derived for our proposed GSTBC-SM scheme, which are verified by simulation results. Specifically, the performance of the proposed

approximate ABEP approaches that of accurate counterpart with significantly reduced complexity.

The remainder of this paper is organized as follows. Section II gives a brief introduction of the conventional STBC schemes. Section III introduces the system model of the proposed GSTBC-SM. In Section IV, we propose low-complexity detectors for our GSTBC-SM schemes. In Section V, we derive the ABEP upper bound of the proposed GSTBC-SM scheme. Simulation results are presented in Section VI. Finally, Section VII concludes this paper.

Notation: $\|\cdot\|^2$ denotes the two norm of a matrix. $|\cdot|$ represents the magnitude of a complex quantity. $(\cdot)^*$, $(\cdot)^T$ and $(\cdot)^H$ stand for the conjugate, transpose and the Hermitian transpose of a vector/matrix, respectively. $\mathbf{I}_N$ denotes $N \times N$ identity matrix. $\text{Tr}(\mathbf{A})$ represents the trace of the matrix $\mathbf{A}$.

II. AN OVER REVIEW OF EXISTING STBC SCHEMES

For a conventional STBC-MIMO having N_t^c TAs and N_r receive antennas (RAs), K APM symbols are transmitted over T time slots by STBC coding, hence the coding ratio R is defined as $R = K/T$. As mentioned in introduction, there are lots of STBC schemes having varied R . For simplicity, we mainly illustrate the GSTBC-SM system based on the OSTBC scheme with $K = 3$ $R = 0.75$ and the Q-OSTBC with $K = 4$ $R = 1$ in this paper [1]. To ease understanding, we first introduce the two schemes as follows.

A. OSTBC scheme with $K = 3$ $R = 3/4$

For an OSTBC system having $N_t^c = 4$ and $R = 3/4$, the information bits are partitioned into $K = 3$ groups, which are modulated into three M -APM symbols as x_1 , x_2 and x_3 . According to [1], the transmitted signal of OSTBC are expressed as

$$\mathbf{G}_{434} = \begin{bmatrix} x_1 - x_2^* & x_3^* & 0 \\ x_2 & x_1^* & 0 & x_3^* \\ x_3 & 0 & -x_1^* - x_2^* \\ 0 & x_3 & x_2 & -x_1 \end{bmatrix}_{4 \times 4}, \quad (1)$$

where $\mathbf{G}_{434} \mathbf{G}_{434}^H = 3\mathbf{I}_4$.

B. Q-OSTBC scheme with $K = 4$ $R = 1$

For a Q-OSTBC scheme with $R = 1$, the information bits are partitioned into four groups, which are modulated into four M -APM symbols as x_1 , x_2 , x_3 and x_4 . According to [1], the transmitted signal of Q-OSTBC are expressed as

$$\mathbf{G}_{Q1} = \begin{bmatrix} x_1 & -x_2^* & -x_3^* e^{-j\phi} & x_4 e^{j\phi} \\ x_2 & x_1^* & -x_4^* e^{-j\phi} & -x_3 e^{j\phi} \\ x_3 e^{j\phi} & -x_4^* e^{-j\phi} & x_1^* & -x_2 \\ x_4 e^{j\phi} & x_3^* e^{-j\phi} & x_2^* & x_1 \end{bmatrix}. \quad (2)$$

Other generalized STBC schemes can be found in [5]-[14].

Fig. 1. System models of the conventional GSTBC and the proposed GSTBC-SM schemes.

III. PROPOSED GENERALIZED SPACE TIME BLOCK CODING SPATIAL MODULATION

We consider a GSTBC-SM structure with N_t TAs and N_r receive antennas over flat Rayleigh fading channels via T time slots. Fig. 1 presents the system model of GSTBC-SM, where the information bits with length of B are split into K blocks to map K SM symbols and then conventional GSTBC structure is invoked for these symbols. The coding ratio is $R = K/T$, which is the same as the conventional counterpart. Explicitly, for the k -th block information bits with length of $B_k = \lceil \log_2 n_t \rceil + M_k$ are further partitioned into two parts to map an SM symbol $\mathbf{x}_k$ using n_t TAs and M_k -APM modulation, where $\lceil \log_2 x \rceil$ denotes the ceil operator. If n_t is a power of 2, the SM mapping method of [25] is employed. Otherwise, $\mathbf{x}_k$ is obtained by the mapping method in [36]. Then, the K SM symbols are invoked into a specific STBC structure [1]. Hence, the transmission rate of the proposed GSTBC-SM is

$$R_p = \frac{\sum_{k=1}^K \lceil \log_2 n_t \rceil + M_k}{T}. \quad (3)$$

According to [1], there are numerous STBC codes with different R . In the proposed GSTBC-SM, the value of n_t is selected based on the specific STBC structure and the number of TAs N_t . Hence the throughput is determined by the specific STBC structure and the value of N_t .

Next, our proposed GSTBC-SM structure with specific examples are introduced and its generalization design will be developed.

A. Proposed OSTBC-SM with $K = 3$ and $R = 3/4$

1) Transmit structure:

In this subsection, GSTBC-SM with OSTBC structure is introduced. For an OSTBC-SM scheme with $R = 3/4$, the

information bits are partitioned into three groups, which are modulated into three SM symbols using $n_t = N_t/4$ TAs as

$$\mathbf{x}_k = (\underbrace{0, \dots, 0}_{q_k-1}, s_k, \underbrace{0, \dots, 0}_{N_t/4-q_k}), k = 1, 2, 3, \quad (4)$$

where q_k is the activated TA index in $\mathbf{x}_k$ and s_k belongs to the constellation symbol set $\mathbb{M}$. According to (1), the transmitted signal $\mathbf{S}_{3/4} \in \mathbb{C}^{N_t \times 4}$ of the OSTBC-SM scheme can be obtained via (1) using $\mathbf{x}_1$, $\mathbf{x}_2$ and $\mathbf{x}_3$ as

$$\mathbf{S}_{3/4} = \begin{bmatrix} \mathbf{x}_1 & -\mathbf{x}_2^* & \mathbf{x}_3^* & \mathbf{O}_{N_t/4} \\ \mathbf{x}_2 & \mathbf{x}_1^* & \mathbf{O}_{N_t/4} & \mathbf{x}_3^* \\ \mathbf{x}_3 & \mathbf{O}_{N_t/4} & -\mathbf{x}_1^* & -\mathbf{x}_2^* \\ \mathbf{O}_{N_t/4} & \mathbf{x}_3 & \mathbf{x}_2 & -\mathbf{x}_1 \end{bmatrix}_{N_t \times 4}. \quad (5)$$

2) Diversity order and coding gain analysis:

In this subsection, the diversity and coding gain of the proposed OSTBC-SM with $R = 3/4$ are analyzed. Assume that $\mathbf{S}_{3/4}^i$ and $\mathbf{S}_{3/4}^j$ are two different OSTBC-SM symbols as

$$\mathbf{S}_{3/4}^i = \begin{bmatrix} \mathbf{x}_1^i - \mathbf{x}_2^{i*} & \mathbf{x}_3^{i*} & 0 \\ \mathbf{x}_2^i & \mathbf{x}_1^{i*} & 0 & \mathbf{x}_3^{i*} \\ \mathbf{x}_3^i & 0 & -\mathbf{x}_1^{i*} & -\mathbf{x}_2^{i*} \\ 0 & \mathbf{x}_3^i & \mathbf{x}_2^i & -\mathbf{x}_1^i \end{bmatrix}, \mathbf{S}_{3/4}^j = \begin{bmatrix} \mathbf{x}_1^j - \mathbf{x}_2^{j*} & \mathbf{x}_3^{j*} & 0 \\ \mathbf{x}_2^j & \mathbf{x}_1^{j*} & 0 & \mathbf{x}_3^{j*} \\ \mathbf{x}_3^j & 0 & -\mathbf{x}_1^{j*} & -\mathbf{x}_2^{j*} \\ 0 & \mathbf{x}_3^j & \mathbf{x}_2^j & -\mathbf{x}_1^j \end{bmatrix}, \quad (6)$$

the coding gain of the OSTBC-SM scheme is defined as

$$G_{3/4} = \min_{\mathbf{S}_{3/4}^i \neq \mathbf{S}_{3/4}^j} |\det(\Delta_{3/4}^H \Delta_{3/4})|, \quad (7)$$

where $\Delta_{3/4} = \mathbf{S}_{3/4}^i - \mathbf{S}_{3/4}^j$. The value of $\Delta_{3/4}^H \Delta_{3/4}$ can be further obtained as

$$\Delta_{3/4}^H \Delta_{3/4} = \begin{bmatrix} A_{3/4} & 0 & 0 & 0 \\ 0 & A_{3/4} & 0 & 0 \\ 0 & 0 & A_{3/4} & 0 \\ 0 & 0 & 0 & A_{3/4} \end{bmatrix}, \quad (8)$$

where $A_{3/4} = \|\mathbf{x}_1^i - \mathbf{x}_1^j\|^2 + \|\mathbf{x}_2^i - \mathbf{x}_2^j\|^2 + \|\mathbf{x}_3^i - \mathbf{x}_3^j\|^2$. Moreover, the value of $A_{3/4}$ can be further obtained as

$$A_{3/4} = \begin{cases} 0, & \text{if } \mathbf{x}_k^i = \mathbf{x}_k^j, k = 1, 2, 3; \\ \|\mathbf{x}_k^i - \mathbf{x}_k^j\|^2, & \text{if } \mathbf{x}_k^i = \mathbf{x}_k^j, k = 2, 3; \\ \|\mathbf{x}_1^i - \mathbf{x}_1^j\|^2 + \|\mathbf{x}_2^i - \mathbf{x}_2^j\|^2, & \text{if } \mathbf{x}_k^i = \mathbf{x}_k^j, k = 3; \\ \|\mathbf{x}_1^i - \mathbf{x}_1^j\|^2 + \|\mathbf{x}_2^i - \mathbf{x}_2^j\|^2 + \|\mathbf{x}_3^i - \mathbf{x}_3^j\|^2, & \text{if } \mathbf{x}_k^i \neq \mathbf{x}_k^j, \forall k. \end{cases} \quad (9)$$

It can be seen from (8) that the rank of $\Delta_{3/4}^H \Delta_{3/4}$ is four, so that the diversity order of the OSTBC-SM is four. According to (7)-(9), its coding gain can be finally obtained as

$$G_{3/4} = \min_{\mathbf{S}_{3/4}^i \neq \mathbf{S}_{3/4}^j} A_{3/4}^4 = \min_{s_i \neq s_j \in \mathbb{M}} [|s_i - s_j|^8, (|s_i|^2 + |s_j|^2)^4]. \quad (10)$$

B. Proposed Q-OSTBC-SM scheme with $K = 4$ and $R = 1$

1) Transmit structure:

In this subsection, GSTBC-SM with Q-OSTBC structure is introduced. For an Q-OSTBC-SM scheme associated with $R = 1$, the information bits are split into four groups, which

are modulated into four SM symbols using $N_t/4$ TAs by (4). The transmitted signal $\mathbf{S}_{Q1} \in \mathbb{C}^{N_t \times 4}$ of the Q-OSTBC-SM scheme can be obtained via (2) using four SM symbols $\mathbf{x}_1, \mathbf{x}_2, \mathbf{x}_3$ and $\mathbf{x}_4$ as

$$\mathbf{S}_{Q1} = \begin{bmatrix} \mathbf{x}_1 & -\mathbf{x}_2^* & -\mathbf{x}_3^* e^{-j\phi} & \mathbf{x}_4 e^{j\phi} \\ \mathbf{x}_2 & \mathbf{x}_1^* & -\mathbf{x}_4^* e^{-j\phi} & -\mathbf{x}_3 e^{j\phi} \\ \mathbf{x}_3 e^{j\phi} & -\mathbf{x}_4^* e^{-j\phi} & \mathbf{x}_1^* & -\mathbf{x}_2 \\ \mathbf{x}_4 e^{j\phi} & \mathbf{x}_3^* e^{-j\phi} & \mathbf{x}_2^* & \mathbf{x}_1 \end{bmatrix}. \quad (11)$$

2) Diversity order and coding gain analysis:

Assuming that $\mathbf{S}_{Q1}^i$ and $\mathbf{S}_{Q1}^j$ are mapping by (11) using $(\mathbf{x}_1^i, \mathbf{x}_2^i, \mathbf{x}_3^i, \mathbf{x}_4^i)$ and $(\mathbf{x}_1^j, \mathbf{x}_2^j, \mathbf{x}_3^j, \mathbf{x}_4^j)$, respectively, the coding gain can be obtained as

$$G_Q = \min_{\mathbf{S}_{Q1}^i \neq \mathbf{S}_{Q1}^j} |\det(\Delta_{Q1}^H \Delta_{Q1})|, \quad (12)$$

where $\Delta_{Q1} = \mathbf{S}_{Q1}^i - \mathbf{S}_{Q1}^j$ is expressed as

$$\Delta_{Q1} = \begin{bmatrix} \mathbf{x}_1^i - \mathbf{x}_1^j & (\mathbf{x}_2^j)^* - (\mathbf{x}_2^i)^* & [(\mathbf{x}_3^j)^* - (\mathbf{x}_3^i)^*] e^{-j\phi} & (\mathbf{x}_4^i - \mathbf{x}_4^j) e^{j\phi} \\ \mathbf{x}_2^i - \mathbf{x}_2^j & (\mathbf{x}_1^j)^* - (\mathbf{x}_1^i)^* & [(\mathbf{x}_4^j)^* - (\mathbf{x}_4^i)^*] e^{-j\phi} & (\mathbf{x}_3^j - \mathbf{x}_3^i) e^{j\phi} \\ (\mathbf{x}_3^i - \mathbf{x}_3^j) e^{j\phi} & [(\mathbf{x}_4^j)^* - (\mathbf{x}_4^i)^*] e^{-j\phi} & (\mathbf{x}_1^i)^* - (\mathbf{x}_1^j)^* & \mathbf{x}_2^j - \mathbf{x}_2^i \\ [(\mathbf{x}_4^i - \mathbf{x}_4^j) e^{j\phi}] & [(\mathbf{x}_3^j)^* - (\mathbf{x}_3^i)^*] e^{-j\phi} & (\mathbf{x}_2^j)^* - (\mathbf{x}_2^i)^* & \mathbf{x}_1^j - \mathbf{x}_1^i \end{bmatrix}. \quad (13)$$

According to (13), the value of $\Delta_{Q1}^H \Delta_{Q1}$ can be obtained by

$$\Delta_{Q1}^H \Delta_{Q1} = \begin{bmatrix} A_{Q1} & 0 & 0 & B_{Q1} \\ 0 & A_{Q1} B_{Q1} & 0 & \\ 0 & B_{Q1} A_{Q1} & 0 & \\ B_{Q1} & 0 & 0 & A_{Q1} \end{bmatrix}, \quad (14)$$

where the value of A_{Q1} and B_{Q1} can be calculated as

$$\begin{aligned} A_{Q1} &= \|\mathbf{\Lambda}_1\|^2 + \|\mathbf{\Lambda}_2\|^2 + \|\mathbf{\Lambda}_3\|^2 + \|\mathbf{\Lambda}_4\|^2, \\ B_{Q1} &= 2\Re[(\mathbf{\Lambda}_1)^H \mathbf{\Lambda}_4 e^{j\phi} - (\mathbf{\Lambda}_2)^H \mathbf{\Lambda}_3 e^{j\phi}], \end{aligned} \quad (15)$$

where $\mathbf{\Lambda}_1 = \mathbf{x}_1^i - \mathbf{x}_1^j \in \varphi$, $\mathbf{\Lambda}_2 = \mathbf{x}_2^i - \mathbf{x}_2^j \in \varphi$, $\mathbf{\Lambda}_3 = \mathbf{x}_3^i - \mathbf{x}_3^j \in \varphi$, $\mathbf{\Lambda}_4 = \mathbf{x}_4^i - \mathbf{x}_4^j \in \varphi$. Explicitly, the value of φ is defined by

$$\varphi = \left\{ [\cdots, s_i - s_j \cdots]^T, [\cdots, s_i, \cdots, -s_j \cdots]^T \right\}, s_i, s_j \in \mathbb{M}. \quad (16)$$

According to [1], we have

$$\det \begin{pmatrix} AB \\ CD \end{pmatrix} = \det(A) \det(D - CA^{-1}B). \quad (17)$$

Hence, the value of $\det(\Delta_{Q1}^H \Delta_{Q1})$ can be obtained via (14) and (17) as

$$\begin{aligned} \det(\Delta_{Q1}^H \Delta_{Q1}) &= (A_{Q1} + B_{Q1})^2 (A_{Q1} - B_{Q1})^2 \\ &= \begin{cases} \|\mathbf{\Lambda}_1\|^8, \text{ if } \mathbf{\Lambda}_k = \mathbf{0}_{N_t/4}, k = 2, 3, 4, \\ \|\mathbf{\Lambda}_1\|^8 |1 - e^{j2\phi}|^4, \text{ if } \mathbf{\Lambda}_k = \mathbf{0}_{N_t/4}, k = 2, 3, \mathbf{\Lambda}_1 = \mathbf{\Lambda}_4, \\ \left[\left(\|\mathbf{\Lambda}_1 - \mathbf{\Lambda}_4 e^{j\phi}\|^2 + \|\mathbf{\Lambda}_2 + \mathbf{\Lambda}_3 e^{j\phi}\|^2 \right)^2 \right. \\ \left. \times (\mathbf{\Lambda}_1 + \mathbf{\Lambda}_4 e^{j\phi^2} + \mathbf{\Lambda}_2 - \mathbf{\Lambda}_3 e^{j\phi^2})^2 \right], \text{ else.} \end{cases} \end{aligned} \quad (18)$$

Then, the coding gain can be finally obtained as

$$\begin{aligned} G_{Q1} &= \min_{\mathbf{S}_{Q1}^i \neq \mathbf{S}_{Q1}^j} |\det(\Delta_{Q1}^H \Delta_{Q1})| \\ &= \min_{s_i \neq s_j \in \mathbb{M}} [|(s_i - s_j)|^8 |1 - e^{j2\phi}|^4, |(s_i - s_j)|^8]. \end{aligned} \quad (19)$$

TABLE I
MAXIMUM CODING GAIN COMPARISONS BETWEEN THE
PROPOSED GSTBC-SM SCHEMES AND CONVENTIONAL STBC
SCHEMES

Rate=3 bpcu				
Scheme (N_t, K, M)	OSTBC (4,3,16)	OSTBC-SM (8,3,8)	Q-OSTBC (4,4,8)	Q-OSTBC-SM (8,4,4)
Coding gain	5.4e-4	0.1177	0.1177	16
Rate=6 bpcu				
Scheme (N_t, K, M)	OSTBC (4,3,256)	OSTBC-SM (64,3,16)	Q-OSTBC (4,4,64)	Q-OSTBC-SM (64,4,4)
Coding gain	1.3e-13	5.4e-4	8.6e-9	16

As a result, the diversity order of Q-OSTBC-SM is four and the optimal ϕ can be obtained as

$$\begin{aligned} \phi &= \max_{\phi \in (0, \pi)} (G_{Q1}) = \max_{\phi \in (0, \pi)} (|1 - e^{j2\phi}|^4, 1) \\ &= \max_{\phi \in (0, \pi)} (|1 - e^{j2\phi}|^4, 1) \\ &= \max_{\phi \in (0, \pi)} (|2 - 2\cos(2\phi)|^2, 1) \\ &= \max_{\phi \in (0, \pi)} [4(2\sin^2(\phi))^2, 1]. \end{aligned} \quad (20)$$

According to (20), the range of ϕ is calculated as

$$4(2\sin^2(\phi))^2 \geq 1 \Rightarrow \pi/6 \leq \phi \leq 5\pi/6. \quad (21)$$

The optimal coding gain of Q-OSTBC-SM is expressed as

$$G_{Q1}^{opt.} = |(s_i - s_j)|^8. \quad (22)$$

In order to provide further insights, Table I compares the maximum coding gains of the proposed OSTBC-SM and Q-OSTBC-SM systems to those of conventional counterpart with the identical number of RF chains and throughput, where bpcu denotes the bits per channel use. It is observed from Table I that the proposed system provides significant coding gain over the corresponding STBC counterpart. As N_t increases, this superiority becomes more apparent.

C. Generalized design

According to [1]-[15], there are numerous eminent STBC schemes, which can be straightforward to extend into our proposed GSTBC-SM structure. The generalize design of GSTBC-SM can be detailed as follows.

Step 1: Obtain a conventional STBC structure by [1]-[15];

Step 2: Obtain the setups (N_t^c, K, T) of the conventional GSTBC scheme;

Step 3: Obtain the setups (n_t, K, T) of the proposed GSTBC-SM scheme based on the values of (N_t^c, K, T) and N_t ;

Step 4: Obtain the APM modulation order M_k for each SM symbol based on the given throughput of (3).

Step 5: Obtain K SM symbols using n_t and M_k ;

Step 6: Obtain the proposed GSTBC-SM symbol by substituting K SM symbols into an APM symbol of conventional STBC structure obtained by Step 1;

Step 7: Analyze the diversity order and coding gain of the proposed GSTBC-SM system. For an OSTBC-SM system, the analysis can be carried out by the approach of (7)-(8), while that can be implemented by the method of (14)-(19) for a Q-OSTBC-SM system.

Fig. 2. Factor graph of the proposed MP detector.

Step 8: Perform signal detection for the proposed GSTBC-SM system.

D. Optimal detector

In this subsection, the ML detector for our proposed GSTBC-SM is designed. Specifically, according to (5) and (11), the received signal $\mathbf{Y} \in \mathbb{C}^{N_r \times T}$ can be formulated as

$$\mathbf{Y} = \mathbf{H}\mathbf{S} + \mathbf{N}, \quad (23)$$

where $\mathbf{H} \in \mathbb{C}^{N_r \times N_t}$ and $\mathbf{N} \in \mathbb{C}^{N_r \times T}$ represent the channel and noise matrices, whose elements obey the complex Gaussian distributions associated with $\mathcal{CN}(0, 1)$ and $\mathcal{CN}(0, \sigma^2)$, respectively, where σ^2 is the noise variance, and $\mathbf{S} \in \mathbb{C}^{N_t \times T}$ denotes the GSTBC-SM symbol. Then, the optimal ML detector can be formulated as

$$\hat{\mathbf{S}} = \arg \min_{\mathbf{S} \in \mathbb{S}} \|\mathbf{Y} - \mathbf{H}\mathbf{S}\|^2, \quad (24)$$

where $\mathbb{S}$ is the set of the proposed GSTBC-SM symbols. According to (3), it is obvious that the complexity order is $O(2^{R_p T})$. Hence the complexity of the ML detector will become extremely high in the context of massive MIMO configurations. Low-complexity detectors become important for our proposed GSTBC-SM in massive MIMO communication scenarios, which will be introduced in the following section.

IV. LOW-COMPLEXITY DETECTORS DESIGN FOR THE PROPOSED GSTBC-SM SYSTEMS

In this section, low-complexity MP and B-MMSE detectors are designed for our GSTBC-SM systems based on the specific STBC structures. Both of them are capable of approaching the optimal ML detector with significantly reduced complexity.

A. Proposed MP detector

Fig. 2 presents the factor graph of the proposed MP detector. As observed from Fig. 2, the messages are passing between

the observation node $\mathbf{y}_t$ to the variable node $\mathbf{x}_k$, which can be expressed by [45]

$$V_{\mathbf{y}_t \rightarrow \mathbf{x}_k}^i = \begin{cases} \mathbf{u}_{t,k}^i = \sum_{v \neq k} \mathbf{H}_{(v)} \sum_{q=1}^Q \mathbf{s}_q p_{tv}^{i-1}(\mathbf{s}_q) \\ (\sigma_{t,k}^2)^i = \sum_{v \neq k} \left(\sum_{q=1}^Q \|\mathbf{H}_{(v)} \mathbf{s}_q\|^2 p_{tv}^{i-1}(\mathbf{s}_q) \right. \\ \left. - \left\| \mathbf{H}_{(v)} \sum_{q=1}^Q \mathbf{s}_q p_{tv}^{i-1}(\mathbf{s}_q) \right\|^2 \right) \end{cases} \quad (25)$$

$$V_{\mathbf{y}_t \leftarrow \mathbf{x}_k}^i = p_{tk}^i(\mathbf{s}_q) = \prod_{j=1, j \neq t}^T p(\mathbf{y}_j | \mathbf{x}_k = \mathbf{s}_q),$$

where $\mathbf{u}_{t,k}^i$ and $(\sigma_{t,k}^2)^i$ denote the mean value and variance associated with the received signal $\mathbf{y}_t$ and transmit signal $\mathbf{x}_k$, respectively, $\mathbf{H}_{(k)}$ is the channel matrix associated with $\mathbf{x}_k$, $\mathbf{s}_q$ is the possible SM symbol, $p_{tv}^{i-1}(\mathbf{s}_q)$ is the probability of $\mathbf{s}_q$ associated with $\mathbf{y}_t$ and $\mathbf{x}_k$ in the $(i-1)$ -th iteration calculation. $V_{\mathbf{y}_t \rightarrow \mathbf{x}_k}^i$ denotes the message from the observation node $\mathbf{y}_t$ to the variable node $\mathbf{x}_k$, while $V_{\mathbf{y}_t \leftarrow \mathbf{x}_k}^i$ denotes the message from the variable node $\mathbf{x}_k$ to the observation node $\mathbf{y}_t$. According to (25), in the proposed MP detector, the calculation of $\mathbf{u}_{t,k}^i$, $(\sigma_{t,k}^2)^i$ and $p_{tv}^{i-1}(\mathbf{s}_q)$ are the key factors. Next, we will introduce these calculations by specific examples.

1) MP detector for OSTBC-SM with $R = 3/4$:

To obtain the values of $\mathbf{u}_{t,k}^i$, $(\sigma_{t,k}^2)^i$ and $p_{tv}^{i-1}(\mathbf{s}_q)$, we first obtain the factor graph by the received signal of (23). Specifically, assuming that $\mathbf{Y} = [\mathbf{y}_1, \mathbf{y}_2, \mathbf{y}_3, \mathbf{y}_4]$, $\mathbf{H} = [\mathbf{H}_1 \mathbf{H}_2 \mathbf{H}_3 \mathbf{H}_4]$, and $\mathbf{N} = [\mathbf{n}_1, \mathbf{n}_2, \mathbf{n}_3, \mathbf{n}_4]$, the received signal can be rewritten as

$$\begin{aligned} \mathbf{y}_1 &= \mathbf{H}_1 \mathbf{x}_1 + \underbrace{\mathbf{H}_2 \mathbf{x}_2 + \mathbf{H}_3 \mathbf{x}_3}_{\mathbf{w}_{1,1}} + \mathbf{n}_1 \\ \mathbf{y}_2 &= \mathbf{H}_2 \mathbf{x}_1^* - \underbrace{\mathbf{H}_1 \mathbf{x}_2^* + \mathbf{H}_4 \mathbf{x}_3}_{\mathbf{w}_{2,1}} + \mathbf{n}_2 \\ \mathbf{y}_3 &= -\mathbf{H}_3 \mathbf{x}_1^* + \underbrace{\mathbf{H}_1 \mathbf{x}_3^* + \mathbf{H}_4 \mathbf{x}_2}_{\mathbf{w}_{3,1}} + \mathbf{n}_3 \\ \mathbf{y}_4 &= -\mathbf{H}_4 \mathbf{x}_1 + \underbrace{\mathbf{H}_2 \mathbf{x}_3^* - \mathbf{H}_3 \mathbf{x}_2^*}_{\mathbf{w}_{4,1}} + \mathbf{n}_4 \end{aligned} \quad (26)$$

$$\begin{aligned}
y_1 &= \mathbf{H}_2 \mathbf{x}_2 + \underbrace{\mathbf{H}_1 \mathbf{x}_1 + \mathbf{H}_3 \mathbf{x}_3 + \mathbf{n}_1}_{\mathbf{w}_{1,2}} \\
y_2 &= -\mathbf{H}_1 \mathbf{x}_2^* + \underbrace{\mathbf{H}_2 \mathbf{x}_1^* + \mathbf{H}_4 \mathbf{x}_3 + \mathbf{n}_2}_{\mathbf{w}_{2,2}} \\
y_3 &= \mathbf{H}_4 \mathbf{x}_2 - \underbrace{\mathbf{H}_3 \mathbf{x}_1^* + \mathbf{H}_1 \mathbf{x}_3^* + \mathbf{n}_3}_{\mathbf{w}_{3,2}} \\
y_4 &= -\mathbf{H}_3 \mathbf{x}_2^* - \underbrace{\mathbf{H}_4 \mathbf{x}_1 + \mathbf{H}_2 \mathbf{x}_3^* + \mathbf{n}_4}_{\mathbf{w}_{4,2}}
\end{aligned} \tag{27}$$

and

$$\begin{aligned}
y_1 &= \mathbf{H}_3 \mathbf{x}_3 + \underbrace{\mathbf{H}_1 \mathbf{x}_1 + \mathbf{H}_2 \mathbf{x}_2 + \mathbf{n}_1}_{\mathbf{w}_{1,3}} \\
y_2 &= \mathbf{H}_4 \mathbf{x}_3 + \underbrace{\mathbf{H}_2 \mathbf{x}_1^* - \mathbf{H}_1 \mathbf{x}_2^* + \mathbf{n}_2}_{\mathbf{w}_{2,3}} \\
y_3 &= \mathbf{H}_1 \mathbf{x}_3^* - \underbrace{\mathbf{H}_3 \mathbf{x}_1^* + \mathbf{H}_4 \mathbf{x}_2}_{\mathbf{w}_{3,3}} + \mathbf{n}_3 \\
y_4 &= \mathbf{H}_2 \mathbf{x}_3^* - \underbrace{\mathbf{H}_4 \mathbf{x}_1 - \mathbf{H}_3 \mathbf{x}_2^* + \mathbf{n}_4}_{\mathbf{w}_{4,3}}
\end{aligned} \tag{28}$$

where $\mathbf{w}_{t,k}$ with $t = (1, 2, 3, 4)$ and $k = (1, 2, 3)$ denotes the interference from symbols $\mathbf{x}_v$ with $v \neq k$. Moreover, $\mathbf{u}_{t,k}^i$ and $(\sigma_{t,k}^2)^i$ denote the mean value and variance of $\mathbf{w}_{t,k}$ at the i -th iteration, which are calculated in the Appendix A, and $p(\mathbf{y}_j|\mathbf{x}_k = s_q)$ denotes the conditional probability, which can be obtained by

$$\begin{aligned}
p(\mathbf{y}_1|\mathbf{x}_1) &\propto \exp\left(\frac{-\|\mathbf{y}_1 - \mathbf{H}_1 \mathbf{x}_1 - \mathbf{u}_{1,1}\|^2}{2\sigma_{1,1}^2}\right), \\
p(\mathbf{y}_2|\mathbf{x}_1) &\propto \exp\left(\frac{-\|\mathbf{y}_2 - \mathbf{H}_2(\mathbf{x}_1)^* - \mathbf{u}_{2,1}\|^2}{2\sigma_{2,1}^2}\right), \\
p(\mathbf{y}_3|\mathbf{x}_1) &\propto \exp\left(\frac{-\|\mathbf{y}_3 + \mathbf{H}_3(\mathbf{x}_1)^* - \mathbf{u}_{3,1}\|^2}{2\sigma_{3,1}^2}\right), \\
p(\mathbf{y}_4|\mathbf{x}_1) &\propto \exp\left(\frac{-\|\mathbf{y}_4 + \mathbf{H}_4 \mathbf{x}_1 - \mathbf{u}_{4,1}\|^2}{2\sigma_{4,1}^2}\right),
\end{aligned} \tag{29}$$

$$\begin{aligned}
p(\mathbf{y}_1|\mathbf{x}_2) &\propto \exp\left(\frac{-\|\mathbf{y}_1 - \mathbf{H}_2 \mathbf{x}_2 - \mathbf{u}_{1,2}\|^2}{2\sigma_{1,2}^2}\right), \\
p(\mathbf{y}_2|\mathbf{x}_2) &\propto \exp\left(\frac{-\|\mathbf{y}_2 + \mathbf{H}_1(\mathbf{x}_2)^* - \mathbf{u}_{2,2}\|^2}{2\sigma_{2,2}^2}\right), \\
p(\mathbf{y}_3|\mathbf{x}_2) &\propto \exp\left(\frac{-\|\mathbf{y}_3 - \mathbf{H}_4 \mathbf{x}_2 - \mathbf{u}_{3,2}\|^2}{2\sigma_{3,2}^2}\right), \\
p(\mathbf{y}_4|\mathbf{x}_2) &\propto \exp\left(\frac{-\|\mathbf{y}_4 + \mathbf{H}_3(\mathbf{x}_2)^* - \mathbf{u}_{4,2}\|^2}{2\sigma_{4,2}^2}\right),
\end{aligned} \tag{30}$$

$$\begin{aligned}
p(\mathbf{y}_1|\mathbf{x}_3) &\propto \exp\left(\frac{-\|\mathbf{y}_1 - \mathbf{H}_3 \mathbf{x}_3 - \mathbf{u}_{1,3}\|^2}{2\sigma_{1,3}^2}\right), \\
p(\mathbf{y}_2|\mathbf{x}_3) &\propto \exp\left(\frac{-\|\mathbf{y}_2 - \mathbf{H}_4 \mathbf{x}_3 - \mathbf{u}_{2,3}\|^2}{2\sigma_{2,3}^2}\right), \\
p(\mathbf{y}_3|\mathbf{x}_3) &\propto \exp\left(\frac{-\|\mathbf{y}_3 - \mathbf{H}_1(\mathbf{x}_3)^* - \mathbf{u}_{3,3}\|^2}{2\sigma_{3,3}^2}\right), \\
p(\mathbf{y}_4|\mathbf{x}_3) &\propto \exp\left(\frac{-\|\mathbf{y}_4 - \mathbf{H}_2(\mathbf{x}_3)^* - \mathbf{u}_{4,3}\|^2}{2\sigma_{4,3}^2}\right).
\end{aligned} \tag{31}$$

According to (25)-(31), the process of the MP detector for OSTBC-SM with $R = 3/4$ is detailed as follows.

Step 1: Initialize the probability $p_{tk}^0(s_q)$ by $p_{tk}^0(s_q) = \frac{1}{Q} (\forall t, k, q)$, where $Q = 2^{R_p T}$.

Step 2: Compute the messages $\mathbf{u}_{t,k}^i, (\sigma_{t,k}^2)^i$ for $t = (1, 2, 3, 4), k = (1, 2, 3)$ via Appendix A using $p_{tk}^{i-1}(s_q)$.

Step 3: Compute the messages $p_{tk}^i(s_q)$ for $t = (1, 2, 3, 4), k = (1, 2, 3)$ using the values of $\mathbf{u}_{t,k}^i, (\sigma_{t,k}^2)^i$ by

$$p_{tk}^i(s_q) = \prod_{j=1, j \neq t}^M p(\mathbf{y}_j|\mathbf{x}_k = s_q). \tag{32}$$

Step 4: Repeat Steps 2 and 3 until $\|\mathbf{P}^i - \mathbf{P}^{i-1}\|^2 < \varsigma$, where $\varsigma \in (0, 1)$ is a small constant and $\mathbf{P}^i = [p_{tk}^i(s_q)]$ for $t = (1, 2, 3, 4)$ and $k = (1, 2, 3)$.

Step 5: $\hat{q} = \arg \max_{q \in (1, Q)} \prod_{t=1}^4 p(\mathbf{y}_t|\mathbf{x}_k = s_q); \hat{\mathbf{x}}_k = s_{\hat{q}}.$

2) *Proposed MP detector for GSTBC-SM:*

According to Fig. 2, (25) and the descriptions of MP detector for OSTBC-SM with $R = 0.75$, the process of MP for GSTBC-SM is detailed in Algorithm 1, where N_{iter} is the number of iterations and $\alpha \in (0, 1)$ is a damping factor. According to Algorithm 1, for a specific GSTBC-SM structure, we can calculate the values of $\mathbf{u}_{t,k}^i$ and $(\sigma_{t,k}^2)^i$ by (25), which can further help to attain the conditional probabilities of line 10. Hence, the proposed MP detector is suitable for any GSTBC-SM systems. However, in [45], the authors indicate that MP detector is more suitable for the case of $K < T$. For the case of $K = T$, it may exhibit performance loss, which will be demonstrated in simulation results. In order to make the signal detection of the proposed GSTBC-SM system more comprehensive, a novel B-MMSE detector will be developed in the next subsection.

Algorithm 1 The designed MP detector for the GSTBC-SM systems

Input: $y, \mathbf{H}, \sigma^2, p_{tk}^0(s_q) = 1/Q, (q = 1, \dots, Q), (t = 1, \dots, T), (k = 1, \dots, K).$

Output: $\hat{\mathbf{x}}_k (k = 1, \dots, K);$

```

1: for  $i \in (1, N_{\text{iter}})$  do
2:   for  $t \in (1, T)$  do
3:     for  $k \in (1, K)$  do
4:       Calculate  $\mathbf{u}_{t,k}^i$  and  $(\sigma_{t,k}^2)^i$  by the given GSTBC-SM structure;
5:     end for
6:   end for
7:   for  $t \in (1, T)$  do
8:     for  $k \in (1, K)$  do
9:       for  $q \in (1, Q)$  do
10:         $p_{tk}(s_q) \propto \prod_{j=1, j \neq t}^T p(\mathbf{y}_j|\mathbf{x}_k).$ 
11:         $p_{tk}^i(s_q) = \frac{1-\alpha}{C_{tk}} p_{tk}(s_q) + \alpha p_{tk}^{i-1}(s_q), C_{tk}$  is a normalizing constant.
12:      end for
13:    end for
14:  end for
15: end for
16: for  $k \in (1, K)$  do
17:    $\hat{q} = \arg \max_{q \in (1, Q)} \prod_{t=1}^T p(\mathbf{y}_t|\mathbf{x}_k = s_q);$ 
18:    $\hat{\mathbf{x}}_k = s_{\hat{q}}.$ 
19: end for

```

B. Proposed B-MMSE detector

In this section, low-complexity B-MMSE detector is designed for GSTBC-SM system. Firstly, the design for the Q-OSTBC-SM with $R = 1$ is introduced, then its extension for the GSTBC-SM is developed.

1) *B-MMSE for Q-OSTBC-SM with $R = 1$:*

In the Q-OSTBC-SM system, assuming that $\mathbf{Y} = [\mathbf{y}_1, \mathbf{y}_2, \mathbf{y}_3, \mathbf{y}_4]$, $\mathbf{H} = [\mathbf{H}_1 \mathbf{H}_2 \mathbf{H}_3 \mathbf{H}_4]$, and $\mathbf{N} = [\mathbf{n}_1, \mathbf{n}_2, \mathbf{n}_3, \mathbf{n}_4]$, the Q-OSTBC-SM received signal can be formulated via (11) and (23) as

$$\begin{aligned} \mathbf{y}_1 &= \mathbf{H}_1 \mathbf{x}_1 + \mathbf{H}_2 \mathbf{x}_2 + \mathbf{H}_3 \mathbf{x}_3 e^{j\phi} + \mathbf{H}_4 \mathbf{x}_4 e^{j\phi} + \mathbf{n}_1, \\ \mathbf{y}_2 &= \mathbf{H}_2 \mathbf{x}_1^* - \mathbf{H}_1 \mathbf{x}_2^* - \mathbf{H}_3 \mathbf{x}_4^* e^{-j\phi} + \mathbf{H}_4 \mathbf{x}_3^* e^{-j\phi} + \mathbf{n}_2, \\ \mathbf{y}_3 &= \mathbf{H}_3 \mathbf{x}_1^* + \mathbf{H}_4 \mathbf{x}_2^* - \mathbf{H}_1 \mathbf{x}_3^* e^{-j\phi} - \mathbf{H}_2 \mathbf{x}_4^* e^{-j\phi} + \mathbf{n}_3, \\ \mathbf{y}_4 &= \mathbf{H}_4 \mathbf{x}_1 - \mathbf{H}_3 \mathbf{x}_2 - \mathbf{H}_2 \mathbf{x}_3 e^{j\phi} + \mathbf{H}_1 \mathbf{x}_4 e^{j\phi} + \mathbf{n}_4. \end{aligned} \quad (33)$$

To easy detection, (33) can be further represented as

$$\underbrace{\begin{bmatrix} \mathbf{y}_1 \\ \mathbf{y}_2^* \\ \mathbf{y}_3^* \\ \mathbf{y}_4 \end{bmatrix}}_{\hat{\mathbf{y}}} = \underbrace{\begin{bmatrix} \mathbf{H}_1 & \mathbf{H}_2 & \mathbf{H}_3 & \mathbf{H}_4 \\ \mathbf{H}_2^* - \mathbf{H}_1^* & \mathbf{H}_4^* - \mathbf{H}_3^* & \mathbf{H}_3^* & \mathbf{H}_4^* \\ \mathbf{H}_3^* & \mathbf{H}_4^* - \mathbf{H}_1^* - \mathbf{H}_2^* & \mathbf{H}_1^* & \mathbf{H}_2^* \\ \mathbf{H}_4 - \mathbf{H}_3 - \mathbf{H}_2 & \mathbf{H}_1 & \mathbf{H}_2 & \mathbf{H}_3 \end{bmatrix}}_{\hat{\mathbf{H}}} \underbrace{\begin{bmatrix} \mathbf{x}_1 \\ \mathbf{x}_2 \\ \mathbf{x}_3 e^{j\phi} \\ \mathbf{x}_4 e^{j\phi} \end{bmatrix}}_{\hat{\mathbf{x}}} + \underbrace{\begin{bmatrix} \mathbf{n}_1 \\ \mathbf{n}_2^* \\ \mathbf{n}_3^* \\ \mathbf{n}_4 \end{bmatrix}}_{\hat{\mathbf{n}}}. \quad (34)$$

It is obvious that there are lots of zeros in the vector $\hat{\mathbf{x}}$, so that (34) can be simplified as

$$\hat{\mathbf{y}} = \hat{\mathbf{H}}_{\Omega} \mathbf{s} + \hat{\mathbf{n}}, \quad (35)$$

where Ω is the antenna combination and $\mathbf{s}$ is the symbol vector, which are expressed as

$$\begin{aligned} \Omega &= (q_1, n_l + q_2, 2n_l + q_3, 3n_l + q_4), \\ \mathbf{s} &= [s_1 s_2 s_3 e^{j\phi} s_4 e^{j\phi}]^T, \end{aligned} \quad (36)$$

where q_l denotes the activated index of $\mathbf{x}_l$ and $n_l = N_t/4$. Then, the optimal antenna combination can be obtained as

$$\hat{\Omega} = \arg \min_{\Omega \in \mathbb{I}} \|\hat{\mathbf{y}} - \hat{\mathbf{H}}_{\Omega} \mathbf{s}_{\Omega}\|^2, \quad (37)$$

where $\mathbb{I}$ is the set of antenna combination and $\mathbf{s}_{\Omega}$ is

$$\mathbf{s}_{\Omega} = ([\hat{\mathbf{H}}_{\Omega}^H \hat{\mathbf{H}}_{\Omega} + \sigma^2 \mathbf{I}_4]^{-1} \hat{\mathbf{H}}_{\Omega}^H \hat{\mathbf{y}}). \quad (38)$$

After obtaining the optimal antenna combination $\hat{\Omega}$, the symbol vector can be obtained as

$$\hat{\mathbf{s}} = ([\hat{\mathbf{H}}_{\hat{\Omega}}^H \hat{\mathbf{H}}_{\hat{\Omega}} + \sigma^2 \mathbf{I}_4]^{-1} \hat{\mathbf{H}}_{\hat{\Omega}}^H \hat{\mathbf{y}}). \quad (39)$$

To further reduce the complexity, if the condition satisfies

$$\|\hat{\mathbf{y}} - \hat{\mathbf{H}}_{\hat{\Omega}} \mathbf{s}_{\hat{\Omega}}\|^2 \leq V_{\text{th}}, \quad (40)$$

where $V_{\text{th}} = \beta N_r \sigma^2 T$, the detection process will terminate [44]. The parameter β is a constant to balance the tradeoff between the computation complexity and performance.

2) *B-MMSE for GSTBC-SM:*

In the GSTBC-SM system, assuming that we have $\mathbf{Y} = [\mathbf{y}_1, \mathbf{y}_2, \dots, \mathbf{y}_T]$, $\mathbf{H} = [\mathbf{H}_1 \mathbf{H}_2 \dots \mathbf{H}_T]$, and $\mathbf{N} = [\mathbf{n}_1, \mathbf{n}_2, \dots, \mathbf{n}_T]$. If the received signal can be reformulated as

$$\tilde{\mathbf{y}} = \tilde{\mathbf{H}} \tilde{\mathbf{x}} + \tilde{\mathbf{n}}, \quad (41)$$

where $\tilde{\mathbf{y}} \in \mathbb{C}^{N_r T \times 1}$ is associated with $\mathbf{y}_1, \mathbf{y}_2, \dots, \mathbf{y}_T$, $\tilde{\mathbf{H}} \in \mathbb{C}^{N_r T \times n_t K}$ is associated with $\mathbf{H}_1, \mathbf{H}_2, \dots, \mathbf{H}_T$ and $\tilde{\mathbf{n}} \in \mathbb{C}^{N_r T \times 1}$ is associated with $\mathbf{n}_1, \mathbf{n}_2, \dots, \mathbf{n}_T$, then B-MMSE detector can be employed; Otherwise, MP detector is more attractive.

TABLE II
COMPLEXITY OF PROPOSED MP DETECTOR

Operations (25)		Flops	Average times
$\mathbf{u}_{t,k}^i$	$\tilde{\mathbf{x}}_k = \sum_{q=1}^Q s_q p_{tk}(s_q)$	$2Q$	$(K-1)TKN_{\text{iter}}$
	$\mathbf{H}_{(k)} \tilde{\mathbf{x}}_k$	$6N_r$	
$(\sigma_{t,k}^2)^i$	$\tilde{\mathbf{x}}_k^2 = \sum_{q=1}^Q s_q ^2 p_{tk}(s_q)$	$3Q$	$(K-1)TKN_{\text{iter}}$
	$\ \mathbf{H}_{(k)}\ ^2 \tilde{\mathbf{x}}_k^2$	$4N_r$	
$p_{t,k}^i(s_q)$	$p(\mathbf{y}_j \mathbf{x}_k = s_q)$	$14N_r$	$Q(T-1)KTN_{\text{iter}}$
Total	$[(5Q + 10N_r)(K-1) + 14N_r Q(T-1)]TKN_{\text{iter}}$		

C. Complexity analysis

In this subsection, the complexities of the proposed MP and B-MMSE detectors are evaluated in terms of real-valued Flops. For specific matrices $\mathbf{A} \in \mathbb{C}^{m \times n}$, $\mathbf{B} \in \mathbb{C}^{n \times p}$, $\mathbf{c} \in \mathbb{C}^{n \times 1}$ and $\mathbf{d} \in \mathbb{C}^{n \times 1}$, the operations of $\mathbf{AB}$, $\|\mathbf{c}\|_F^2$ and $\mathbf{c} \pm \mathbf{d}$ require $8mnp - 2mp$, $4n - 1$, and $2n$ flops, respectively. The complexity of the ML detector can be expressed as

$$C_{\text{ML}} = (8N_r KT + 4N_r T)2^{R_p T}. \quad (42)$$

According to (25) and Algorithm 1, the complexity of the proposed MP detector is expressed as

$$C_{\text{MP}} = [(5Q + 10N_r)(K-1) + 14N_r Q(T-1)]TKN_{\text{iter}}, \quad (43)$$

where N_{iter} is the number of iterations and the detailed calculation process is presented in Table II.

On the other hand, the complexity of the B-MMSE detector is mainly imposed by the calculation of (39), whose complexity is $4K^3 + 12K^2 N_r + 7K^2 + 6N_r K$ [44]. Assuming the average number of MMSE detection of (39) is N_{avg} , the complexity of B-MMSE detector can be obtained as

$$C_{\text{B-MMSE}} = N_{\text{avg}}(4K^3 + 12K^2 N_r + 7K^2 + 6N_r K). \quad (44)$$

where N_{avg} is the average number of the operation in (38).

V. ABEP ANALYSIS FOR THE PROPOSED GSTBC-SM SYSTEMS

In this section, ABEP upper bound is derived for our coherent GSTBC-SM system. Assuming that the transmitted and received signals of GSTBC-SM are $\mathbf{S}_i$ and $\mathbf{S}_j$, respectively, the ABEP upper bound is given by

$$P_b = \frac{1}{TR_p 2^{TR_p}} \sum_{i=1}^{2^{TR_p}} \sum_{j=1, j \neq i}^{2^{TR_p}} d(\mathbf{S}_i, \mathbf{S}_j) P(\mathbf{S}_i \rightarrow \mathbf{S}_j), \quad (45)$$

where $d(\mathbf{S}_i, \mathbf{S}_j)$ is the number of error bits and $P(\mathbf{S}_i \rightarrow \mathbf{S}_j)$ denotes the pairwise error probability (PEP). According to [36], the PEP can be obtained by

$$\begin{aligned} P(\mathbf{S}_i \rightarrow \mathbf{S}_j) &= \mathbf{E}(P(\mathbf{S}_i \rightarrow \mathbf{S}_j) | \mathbf{H}) \\ &= \mathbf{E}\left(Q\left(\sqrt{\frac{\|\mathbf{H}\Delta\|^2}{2\sigma^2}}\right)\right) \\ &= \int Q\left(\sqrt{\frac{\gamma}{2\sigma^2}}\right) d\gamma, \end{aligned} \quad (46)$$

where $\mathbf{E}(\cdot)$ denotes expectation operator, $Q(x) = \int_x^\infty \frac{1}{\sqrt{2\pi}} e^{-x^2/2} dx$, $\gamma = \|\mathbf{H}\Delta\|^2$ with $\Delta = (\mathbf{S}_i - \mathbf{S}_j)$. According to [36], the value of $P(\mathbf{S}_i \rightarrow \mathbf{S}_j)$ is dominated by the MGF of γ and the value of $Q(x)$, which will be introduced as follows.

A. MGF of γ

The MGF of γ is expressed as $M_\gamma(s) = \int e^{\gamma s} f_\gamma(\gamma) d\gamma$ and calculated as follows. First, the value of γ is expressed as

$$\gamma = \text{Tr} [\mathbf{H}\Delta\Delta^H\mathbf{H}^H] = \text{Tr} [\tilde{\mathbf{H}}\mathbf{V}\tilde{\mathbf{H}}^H] = \sum_{r=1}^{N_r} \sum_{n=1}^{N_t} |\lambda_n \tilde{\mathbf{H}}_{rn}|^2, \quad (47)$$

where $\Delta\Delta^H = \mathbf{U}\mathbf{V}\mathbf{U}^H$ with $\mathbf{U}\mathbf{U}^H = \mathbf{I}_{N_t}$, $\mathbf{V} = \text{diag}(\lambda_1, \dots, \lambda_{N_t})$, $\tilde{\mathbf{H}}_{rn}$ denotes the r -th row and n -column of $\tilde{\mathbf{H}} = \mathbf{H}\mathbf{U}$, and λ_n denotes the n -th eigenvalues of $\Delta\Delta^H$. Assuming that $\gamma_r = \sum_{n=1}^{N_t} |\lambda_n \tilde{\mathbf{H}}_{rn}|^2$, we have $\gamma = \sum_{r=1}^{N_r} \gamma_r$, and the MGF of γ_r is expressed as

$$M_\gamma(s) = \int e^{\gamma s} f_\gamma(\gamma) d\gamma = \prod_{n=1}^{N_r} M_{\gamma_r}(s), \quad (48)$$

where $M_{\gamma_r}(s) = \int e^{\gamma_r s} f_{\gamma_r}(\gamma_r) d\gamma_r$. Furthermore, the value of γ_r can be reformulated as

$$\gamma_r = \sum_{n=1}^{N_t} |\lambda_n \tilde{\mathbf{H}}_{rn}|^2 = \sum_{n=1}^{\kappa} \lambda_n |\tilde{\mathbf{H}}_{rn}|^2 = \sum_{n=1}^{\kappa} \Phi_n. \quad (49)$$

where κ is the number of nonzero eigenvalues and $\Phi_n = \lambda_n |\tilde{\mathbf{H}}_{rn}|^2$. Hence, the MGF of γ_r can be obtained as

$$M_{\gamma_r}(s) = \prod_{n=1}^{\kappa} M_{\Phi_n}(s). \quad (50)$$

Note that, we have

$$2\Phi_n/\lambda_n = 2|\tilde{\mathbf{H}}_{rn}|^2 \in \chi^2(2) \rightarrow M_\chi(s) = (1 - 2s)^{-1}. \quad (51)$$

Then, the MGF of Φ_n is expressed as $M_{\Phi_n}(s) = (1 - \lambda_n s)^{-1}$. As a result, the MGF of γ can be obtained as

$$M_\gamma(s) = \prod_{r=1}^{N_r} \prod_{n=1}^{\kappa} M_{\Phi_n}(s) = \prod_{n=1}^{\kappa} (1 - \lambda_n s)^{-N_r}, \quad (52)$$

where $\lambda_1, \dots, \lambda_\kappa$ are the nonzero eigenvalues.

B. Accurate upper bound of ABEP

According to [36], $Q(x)$ can be accurately represented as

$$Q(x) = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} \exp\left(-\frac{x^2}{2 \sin^2 \theta}\right) d\theta. \quad (53)$$

Then, the PEP value can be obtained as

$$\begin{aligned} P(\mathbf{S}_i \rightarrow \mathbf{S}_j) &= \frac{1}{\pi} \int_0^{\pi/2} M_\gamma\left(\frac{-1}{4\sigma^2 \sin^2 \theta}\right) d\theta \\ &= \frac{1}{\pi} \int_0^{\pi/2} \prod_{n=1}^{\kappa} \left(1 + \frac{\lambda_n}{4\sigma^2 \sin^2 \theta}\right)^{-N_r} d\theta. \end{aligned} \quad (54)$$

By substituting (54) into (45), the accurate ABEP upper bound can be obtained.

C. Approximate upper bound of ABEP

Since it is difficult to obtain the closed form of (54), we need to calculate the integration in the calculation of the upper bound, which imposes extra complexity. In order to further simplify the calculation of the ABEP upper bound, the value of $Q(x)$ can be approximated as [46]

$$Q(x) \approx \frac{1}{12} e^{-\frac{x^2}{2}} + \frac{1}{4} e^{-\frac{2x^2}{3}}. \quad (55)$$

Then, the value of $P(\mathbf{x}_i^t \rightarrow \mathbf{x}_i^j)$ can be derived as

$$\begin{aligned} P(\mathbf{x}_i^t \rightarrow \mathbf{x}_i^j) &\approx \int \left(\frac{1}{12} e^{-\frac{\gamma}{4\sigma^2}} + \frac{1}{4} e^{-\frac{\gamma}{3\sigma^2}}\right) f_\gamma(\gamma) d\gamma \\ &= \frac{1}{12} M_\gamma\left(-\frac{1}{4\sigma^2}\right) + \frac{1}{4} M_\gamma\left(-\frac{1}{3\sigma^2}\right), \\ &= \frac{1}{12} \prod_{n=1}^{\kappa} \left(1 + \frac{\lambda_n}{4\sigma^2}\right)^{-N_r} + \frac{1}{4} \prod_{n=1}^{\kappa} \left(1 + \frac{\lambda_n}{3\sigma^2}\right)^{-N_r}. \end{aligned} \quad (56)$$

By substituting (56) into (45), the approximate ABEP upper bound can be obtained.

VI. SIMULATION RESULTS

In this section, we conduct simulations to verify the performances of our proposed GSTBC-SM systems in the context of different antenna configurations, where the perfect channel state information is assumed at the receiver.

A. Performance comparison of small-scale MIMO configurations

In this subsection, we compare the performances of the proposed GSTBC-SM systems to those of the conventional GSTBC and space shift keying (SSK) systems [47] in the context of small-scale MIMO configurations. Moreover, both the accurate and approximate ABEP upper bound are added for benchmarks. It can be observed from Figs. 3 and 6 that, the derived approximate upper bound approaches the accurate upper bound and both are closely tight with simulation results upon increasing the SNR values.

Specifically, Fig. 3 and 4 compare the performances of the proposed GSTBC-SM systems to those of conventional GSTBC and SSK systems at the throughput of 3 bpcu in conjunction with $N_r = 1$ and $N_r = 2$, respectively. To achieve 3 bpcu, $(N_t, K, M, T) = (8, 4, 4, 4)$, $(8, 3, 8, 4)$ and $(8, 2, 2, 2)$ are employed for proposed Q-OSTBC-SM, OSTBC-SM, Alamouti-SM [36] systems, while $(N_t, K, M, T) = (4, 4, 8, 4)$, $(4, 3, 16, 4)$ and $(2, 2, 8, 2)$ are used for conventional Q-OSTBC, OSTBC and Alamouti systems. It can be observed from Figs. 3-4 that, Q-OSTBC-SM with $\pi/4$ performs the best and outperforms corresponding Q-OSTBC by 8 dB and 5 dB at $\text{BER}=10^{-5}$ for the case of $N_r = 1$ and $N_r = 2$, respectively. OSTBC-SM outperforms corresponding OSTBC by around 6 dB at $\text{BER}=10^{-5}$ for both the cases of $N_r = 1$ and $N_r = 2$. Since the diversity order of proposed GSTBC-SM is four, which is larger than that of Alamouti-SM, Alamouti and SSK schemes, all the proposed GSTBC-SM systems are capable of providing significant performance gains over the Alamouti-SM, Alamouti and SSK systems.

Fig. 3. Performance comparisons for proposed GSTBC-SM systems and conventional GSTBC systems having $N_t = 8$ $N_r = 1$ at 3 bpcu.

Figs. 5-6 compare the performances of the proposed GSTBC-SM systems to those of conventional GSTBC and SSK systems at the throughput of 4 bpcu. To achieve 4 bpcu, $(N_t, K, M, T) = (16, 4, 4, 4)$, and $(16, 2, 2, 2)$ are employed for Q-OSTBC-SM, Alamouti-SM [36], while $(N_t, K, M, T) = (4, 4, 16, 4)$, and $(2, 2, 16, 2)$ are employed for Q-OSTBC and Alamouti. Since it is difficult for OSTBC-SM and OSTBC systems to attain 4 bpcu, the performance of them are not included for comparison. It can be observed from Fig. 4 that, Q-OSTBC-SM with $\phi = \pi/4$ still performs the best and outperforms corresponding Q-OSTBC by around 8 dB at $\text{BER}=10^{-5}$ for both the cases of $N_r = 1$ and $N_r = 2$. As N_r increases, the performance of Q-OSTBC-SM with $\phi = 0$ approaches that of Q-OSTBC-SM with $\phi = \pi/4$. It is also obvious that the proposed Q-OSTBC-SM system with $\phi = \pi/4$ outperforms Alamouti-SM, Alamouti and SSK systems by 5 dB, 12 dB and 13 dB at $\text{BER}=10^{-5}$ for the case of $N_r = 2$, respectively. The performance advantages become more dominant for the case of $N_r = 1$.

Fig. 7 compares the performances of the proposed Q-OSTBC-SM systems to those of SSK-vertical bell labs space-time (SSK-VBLAST) [47] systems having $N_t = 8$ at the throughput of 4 bpcu. To achieve 4 bpcu, $(N_t, K, M, T) = (8, 4, 8, 4)$ and $(8, 4, 1, 1)$ are employed for Q-OSTBC-SM and SSK-VBLAST systems, respectively. It is obvious that the proposed Q-OSTBC-SM system with $\phi = \pi/4$ outperforms SSK-VBLAST systems by 10 dB and 20 dB at $\text{BER}=10^{-4}$ for the case of $N_r = 2$ and $N_r = 1$, respectively.

B. Performance and complexity comparisons between different detectors of proposed GSTBC-SM systems

In order to achieve near optimal performance with significantly reduced complexity in the context of massive MIMO communication, low-complexity MP and B-MMSE detectors are designed for our OSTBC-SM and Q-OSTBC-SM sys-

Fig. 4. Performance comparisons for proposed GSTBC-SM systems and conventional GSTBC systems having $N_t = 8$ $N_r = 2$ at 3 bpcu.

Fig. 5. Performance comparisons for proposed GSTBC-SM systems and conventional GSTBC and SSK systems having $N_t = 16$ $N_r = 1$ at 4 bpcu.

tems, respectively. In this subsection, the performance and complexity of the proposed low-complexity detectors are compared with the ML detector in Figs. 8-9 and Table III, respectively, where we have $C_r = (C_{\text{ML}} - C_{\text{MP}})/C_{\text{ML}}$ and $C_r = (C_{\text{ML}} - C_{\text{B-MMSE}})/C_{\text{ML}}$ for the proposed OSTBC-SM and Q-OSTBC-SM, respectively.

Specifically, Fig. 8 compares the performances of the proposed MP and ML detectors for the QPSK-aided OSTBC-SM systems with $N_t = 8$ and $N_t = 16$. It can be observed from Fig. 8 that, the performance of the proposed MP detector approaches that of the ML detector for the cases of $N_t = 8$, $N_r = 4$ and $N_t = 16$, $N_r = 6$. For the case of small number of receiver antennas, the proposed MP detector suffer from near 2 dB performance loss over the ML detector. It can

Fig. 6. Performance comparisons for proposed GSTBC-SM systems and conventional GSTBC and SSK systems having $N_t = 16$ $N_r = 2$ at 4 bpcu.

Fig. 7. Performance comparisons for proposed GSTBC-SM systems and SSK-VBLAST systems having $N_t = 8$ at 4 bpcu.

be observed from Table III that, the proposed MP detector reduces the complexity to about 31.7% $\sim$ 7.7% of that of the ML detector.

Fig. 9 presents the performances of the proposed MP, the proposed B-MMSE and ML detectors for the QPSK-aided Q-OSTBC-SM systems with $N_t = 8$ at 3 bpcu. It can be observed from Fig. 9 that, the performance of the proposed B-MMSE detector depends on the value of β . To balance a tradeoff between the performance and the complexity, it is better to set $\beta = 1.5$, $\beta = 1.2$ and $\beta = 0.9$ for the cases of $N_r = 2$, $N_r = 4$ and $N_r = 8$, respectively. It can be observed from Fig. 9 that, the performance of the B-MMSE detector with the appropriate value of β always approaches that of the ML detector and outperforms that of the proposed MP detector. This is because

TABLE III
COMPLEXITY COMPARISON OF THE PROPOSED DETECTORS WITH ML DETECTOR IN TERMS OF FLOPS.

Scheme	Antenna setup	ML	MP	B-MMSE	C_r
OSTBC-SM	$N_t = 8$ $N_r = 4$	2.3×10^5	7.2×10^4	—	68.53%
	$N_t = 16$ $N_r = 4$	1.8×10^6	1.4×10^5	—	92.34%
Q-OSTBC-SM	$N_t = 8$ $N_r = 2$	1.2×10^6	—	7200 $\beta = 1.5$	99.39%
	$N_t = 8$ $N_r = 4$	2.4×10^6	—	13552 $\beta = 1.2$	99.43%

Fig. 8. Performance comparisons for OSTBC-SM systems associated with different detectors.

that the performance of the MP detector is associated with the value of N_r and K . By further increasing the value of N_r or decreasing the value of K , its performance is able to approach ML detector's counterpart. It can be observed from Table III that, the proposed B-MMSE detector reduces the complexity to about 0.5% $\sim$ 0.6% of that of the ML detector.

In a short, the B-MMSE detector is only suitable for the GSTBC-SM systems whose received signal can be expressed by (41), while MP is suitable for any GSTBC-SM system. For a specific GSTBC-SM system, if B-MMSE detector can be employed, B-MMSE is the more attractive than MP detector; Otherwise, the proposed MP detector is more suitable.

C. Performance comparison of massive MIMO configurations

In this subsection, the performance comparisons in massive MIMO antenna configurations are presented, where the ML detector becomes impractical. Therefore, only the performances of the low-complexity detectors are included for comparison. For fair comparison, both OSTBC-SM and OSTBC systems employ our proposed MP detectors, while both Q-OSTBC-SM and Q-OSTBC systems employ our proposed B-MMSE detectors. Fig. 10 compares the performances of the proposed OSTBC-SM systems to those of conventional OSTBC systems at the throughput of 3.75 bpcu and 4.5 bpcu, respectively. $(N_t, K, M, N_r, T) = (32, 3, 4, 6, 4)$, and $(64, 3, 4, 6, 4)$ are employed for OSTBC-SM, while

Fig. 9. Performance comparisons for Q-OSTBC-SM systems associated with different detectors at 3 bpcu.

Fig. 10. Performance comparisons for proposed OSTBC-SM systems and conventional OSTBC systems at 3.75 bpcu and 4.5 bpcu.

$(N_t, K, M, N_r, T) = (4, 3, 32, 6, 4)$, and $(4, 3, 64, 6, 4)$ are used for conventional OSTBC. For the case of 3 RF chains, it can be observed from Fig. 10 that the proposed OSTBC-SM system outperforms the conventional OSTBC system by around 15 dB and 20 dB at 3.75 bpcu and 4.5 bpcu, respectively, at $\text{BER}=10^{-5}$.

Fig. 11 compares the performance of the proposed Q-OSTBC-SM system to that of conventional Q-OSTBC, SSK and ZF precoded MIMO systems at the throughput of 6 bpcu. $(N_t, K, M, N_r, T) = (64, 4, 4, 2, 4)$ is employed for Q-OSTBC-SM, $(N_t, K, M, N_r, T) = (4, 4, 64, 2, 4)$ is used for conventional Q-OSTBC systems, and $(N_t, M, N_r, T) = (64, 8, 2, 1)$ is employed for ZF precoded MIMO. It can be observed from Fig. 11 that the proposed Q-OSTBC-SM system outperforms the conventional ZF precoded MIMO [48], SSK

Fig. 11. Performance comparisons for proposed Q-OSTBC-SM system and conventional Q-OSTBC, SSK and ZF precoded MIMO systems at 6 bpcu.

and Q-OSTBC systems by around 3 dB, 13 dB and 20 dB at $\text{BER}=10^{-5}$.

VII. CONCLUSIONS

In this paper, we proposed a novel GSTBC-SM scheme for open-loop massive MIMO downlink communication systems, which is capable of attaining high throughput and diversity gain even with small number of RF chains. Then, the performances of the specific OSTBC-SM and Q-OSTBC-SM systems were investigated, while low-complexity detectors were designed in the context of massive MIMO antenna configurations. Both simulation and theoretical results show that the proposed GSTBC-SM scheme provides considerable performance gains over the corresponding STBC schemes. Moreover, the proposed GSTBC-SM scheme employing a large number of TAs provides 20 dB gain over the GSTBC system with identical number of RF chains and throughput.

APPENDIX A

THE CALCULATION OF $u_{t,k}^i$ AND $\sigma_{t,k}^i$ IN (25)

In this Appendix, the values of $u_{t,k}^i$ and $\sigma_{t,k}^i$ in MP detector for OSTBC-SM system are calculated by (26)-(28) as follows

$$\begin{aligned} u_{1,1}^i &= \sum_{q=1}^Q \mathbf{H}_2 \mathbf{s}_q p_{12}^{i-1}(\mathbf{s}_q) + \sum_{q=1}^Q \mathbf{H}_3 \mathbf{s}_q p_{13}^{i-1}(\mathbf{s}_q), \\ \sigma_{1,1}^i &= \sum_{q=1}^Q \|\mathbf{H}_2 \mathbf{s}_q\|^2 p_{12}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_2 \mathbf{s}_q p_{12}^{i-1}(\mathbf{s}_q) \right\|^2 \\ &\quad + \sum_{q=1}^Q \|\mathbf{H}_3 \mathbf{s}_q\|^2 p_{13}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_3 \mathbf{s}_q p_{13}^{i-1}(\mathbf{s}_q) \right\|^2 + \sigma^2, \end{aligned} \quad (57)$$

$$\begin{aligned}
u_{2,1}^i &= - \sum_{q=1}^Q \mathbf{H}_1 \mathbf{s}_q^* p_{22}^{i-1}(\mathbf{s}_q) + \sum_{q=1}^Q \mathbf{H}_4 \mathbf{s}_q p_{23}^{i-1}(\mathbf{s}_q), \\
\sigma_{2,1}^i &= \sum_{q=1}^Q \left\| \mathbf{H}_1 \mathbf{s}_q^* \right\|^2 p_{22}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_1 \mathbf{s}_q^* p_{22}^{i-1}(\mathbf{s}_q) \right\|^2 \\
&+ \sum_{q=1}^Q \left\| \mathbf{H}_4 \mathbf{s}_q \right\|^2 p_{23}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_4 \mathbf{s}_q p_{23}^{i-1}(\mathbf{s}_q) \right\|^2 + \sigma^2,
\end{aligned} \quad (58)$$

$$\begin{aligned}
u_{3,1}^i &= \sum_{q=1}^Q \mathbf{H}_4 \mathbf{s}_q p_{32}^{i-1}(\mathbf{s}_q) + \sum_{q=1}^Q \mathbf{H}_1 \mathbf{s}_q^* p_{33}^{i-1}(\mathbf{s}_q), \\
\sigma_{3,1}^i &= \sum_{q=1}^Q \left\| \mathbf{H}_4 \mathbf{s}_q \right\|^2 p_{32}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_4 \mathbf{s}_q p_{32}^{i-1}(\mathbf{s}_q) \right\|^2 \\
&+ \sum_{q=1}^Q \left\| \mathbf{H}_1 \mathbf{s}_q^* \right\|^2 p_{33}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_1 \mathbf{s}_q^* p_{33}^{i-1}(\mathbf{s}_q) \right\|^2 \\
&+ \sigma^2,
\end{aligned} \quad (59)$$

$$\begin{aligned}
u_{4,1}^i &= \sum_{q=1}^Q \mathbf{H}_2 \mathbf{s}_q^* p_{43}^{i-1}(\mathbf{s}_q) - \sum_{q=1}^Q \mathbf{H}_3 \mathbf{s}_q^* p_{42}^{i-1}(\mathbf{s}_q), \\
\sigma_{4,1}^i &= \sum_{q=1}^Q \left\| \mathbf{H}_2 \mathbf{s}_q^* \right\|^2 p_{43}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_2 \mathbf{s}_q^* p_{43}^{i-1}(\mathbf{s}_q) \right\|^2 \\
&+ \sum_{q=1}^Q \left\| \mathbf{H}_3 \mathbf{s}_q^* \right\|^2 p_{42}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_3 \mathbf{s}_q^* p_{42}^{i-1}(\mathbf{s}_q) \right\|^2 + \sigma^2,
\end{aligned} \quad (60)$$

$$\begin{aligned}
u_{1,2}^i &= \sum_{q=1}^Q \mathbf{H}_1 \mathbf{s}_q p_{11}^{i-1}(\mathbf{s}_q) + \sum_{q=1}^Q \mathbf{H}_3 \mathbf{s}_q p_{13}^{i-1}(\mathbf{s}_q), \\
\sigma_{1,2}^i &= \sum_{q=1}^Q \left\| \mathbf{H}_1 \mathbf{s}_q \right\|^2 p_{11}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_1 \mathbf{s}_q p_{11}^{i-1}(\mathbf{s}_q) \right\|^2 \\
&+ \sum_{q=1}^Q \left\| \mathbf{H}_3 \mathbf{s}_q \right\|^2 p_{13}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_3 \mathbf{s}_q p_{13}^{i-1}(\mathbf{s}_q) \right\|^2 + \sigma^2,
\end{aligned} \quad (61)$$

$$\begin{aligned}
u_{2,2}^i &= \sum_{q=1}^Q \mathbf{H}_2 \mathbf{s}_q^* p_{21}^{i-1}(\mathbf{s}_q) + \sum_{q=1}^Q \mathbf{H}_4 \mathbf{s}_q p_{23}^{i-1}(\mathbf{s}_q), \\
\sigma_{2,2}^i &= \sum_{q=1}^Q \left\| \mathbf{H}_2 \mathbf{s}_q^* \right\|^2 p_{21}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_2 \mathbf{s}_q^* p_{21}^{i-1}(\mathbf{s}_q) \right\|^2 \\
&+ \sum_{q=1}^Q \left\| \mathbf{H}_4 \mathbf{s}_q \right\|^2 p_{23}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_4 \mathbf{s}_q p_{23}^{i-1}(\mathbf{s}_q) \right\|^2 + \sigma^2,
\end{aligned} \quad (62)$$

$$\begin{aligned}
u_{3,2}^i &= - \sum_{q=1}^Q \mathbf{H}_3 \mathbf{s}_q^* p_{31}^{i-1}(\mathbf{s}_q) + \sum_{q=1}^Q \mathbf{H}_1 \mathbf{s}_q^* p_{33}^{i-1}(\mathbf{s}_q), \\
\sigma_{3,2}^i &= \sum_{q=1}^Q \left\| \mathbf{H}_3 \mathbf{s}_q^* \right\|^2 p_{31}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_3 \mathbf{s}_q^* p_{31}^{i-1}(\mathbf{s}_q) \right\|^2 \\
&+ \sum_{q=1}^Q \left\| \mathbf{H}_1 \mathbf{s}_q^* \right\|^2 p_{33}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_1 \mathbf{s}_q^* p_{33}^{i-1}(\mathbf{s}_q) \right\|^2 + \sigma^2,
\end{aligned} \quad (63)$$

$$\begin{aligned}
u_{4,2}^i &= \sum_{q=1}^Q \mathbf{H}_4 \mathbf{s}_q p_{41}^{i-1}(\mathbf{s}_q) + \sum_{q=1}^Q \mathbf{H}_2 \mathbf{s}_q^* p_{43}^{i-1}(\mathbf{s}_q), \\
\sigma_{4,2}^i &= \sum_{q=1}^Q \left\| \mathbf{H}_4 \mathbf{s}_q \right\|^2 p_{41}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_4 \mathbf{s}_q p_{41}^{i-1}(\mathbf{s}_q) \right\|^2 \\
&+ \sum_{q=1}^Q \left\| \mathbf{H}_2 \mathbf{s}_q^* \right\|^2 p_{43}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_2 \mathbf{s}_q^* p_{43}^{i-1}(\mathbf{s}_q) \right\|^2 + \sigma^2,
\end{aligned} \quad (64)$$

$$\begin{aligned}
u_{1,3}^i &= \sum_{q=1}^Q \mathbf{H}_1 \mathbf{s}_q p_{11}^{i-1}(\mathbf{s}_q) + \sum_{q=1}^Q \mathbf{H}_2 \mathbf{s}_q p_{12}^{i-1}(\mathbf{s}_q), \\
\sigma_{1,3}^i &= \sum_{q=1}^Q \left\| \mathbf{H}_1 \mathbf{s}_q \right\|^2 p_{11}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_1 \mathbf{s}_q p_{11}^{i-1}(\mathbf{s}_q) \right\|^2 \\
&+ \sum_{q=1}^Q \left\| \mathbf{H}_2 \mathbf{s}_q \right\|^2 p_{12}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_2 \mathbf{s}_q p_{12}^{i-1}(\mathbf{s}_q) \right\|^2 + \sigma^2,
\end{aligned} \quad (65)$$

$$\begin{aligned}
u_{2,3}^i &= \sum_{q=1}^Q \mathbf{H}_2 \mathbf{s}_q^* p_{21}^{i-1}(\mathbf{s}_q) - \sum_{q=1}^Q \mathbf{H}_1 \mathbf{s}_q^* p_{22}^{i-1}(\mathbf{s}_q), \\
\sigma_{2,3}^i &= \sum_{q=1}^Q \left\| \mathbf{H}_2 \mathbf{s}_q^* \right\|^2 p_{21}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_2 \mathbf{s}_q^* p_{21}^{i-1}(\mathbf{s}_q) \right\|^2 \\
&+ \sum_{q=1}^Q \left\| \mathbf{H}_1 \mathbf{s}_q^* \right\|^2 p_{22}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_1 \mathbf{s}_q^* p_{22}^{i-1}(\mathbf{s}_q) \right\|^2 + \sigma^2,
\end{aligned} \quad (66)$$

$$\begin{aligned}
u_{3,3}^i &= - \sum_{q=1}^Q \mathbf{H}_3 \mathbf{s}_q^* p_{31}^{i-1}(\mathbf{s}_q) + \sum_{q=1}^Q \mathbf{H}_4 \mathbf{s}_q p_{32}^{i-1}(\mathbf{s}_q), \\
\sigma_{3,3}^i &= \sum_{q=1}^Q \left\| \mathbf{H}_3 \mathbf{s}_q^* \right\|^2 p_{31}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_3 \mathbf{s}_q^* p_{31}^{i-1}(\mathbf{s}_q) \right\|^2 \\
&+ \sum_{q=1}^Q \left\| \mathbf{H}_4 \mathbf{s}_q \right\|^2 p_{32}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_4 \mathbf{s}_q p_{32}^{i-1}(\mathbf{s}_q) \right\|^2 + \sigma^2,
\end{aligned} \quad (67)$$

$$\begin{aligned}
u_{4,3}^i &= \sum_{q=1}^Q \mathbf{H}_4 \mathbf{s}_q p_{41}^{i-1}(\mathbf{s}_q) - \sum_{q=1}^Q \mathbf{H}_3 \mathbf{s}_q^* p_{42}^{i-1}(\mathbf{s}_q), \\
\sigma_{4,3}^i &= \sum_{q=1}^Q \left\| \mathbf{H}_4 \mathbf{s}_q \right\|^2 p_{41}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_4 \mathbf{s}_q p_{41}^{i-1}(\mathbf{s}_q) \right\|^2 \\
&+ \sum_{q=1}^Q \left\| \mathbf{H}_3 \mathbf{s}_q^* \right\|^2 p_{42}^{i-1}(\mathbf{s}_q) - \left\| \sum_{q=1}^Q \mathbf{H}_3 \mathbf{s}_q^* p_{42}^{i-1}(\mathbf{s}_q) \right\|^2 + \sigma^2.
\end{aligned} \quad (68)$$

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