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RESEARCH ARTICLE

Space-Time Line Code Aided Offset Spatial Modulation

YASHAN PANG¹, YUE XIAO², (Member, IEEE), XIA LEI², (Member, IEEE),
YONG LIANG GUAN³, (Senior Member, IEEE),
AND JINGON JOUNG⁴, (Senior Member, IEEE)

¹National Key Laboratory of Advanced Communication Networks, Academy of Network and Communications of CETC, Shijiazhuang, Hebei 050081, China

²National Key Laboratory of Wireless Communications, University of Electronic Science and Technology of China, Chengdu 611731, China

³School of Electrical and Electronic Engineering, Nanyang Technological University, Singapore 639798

⁴Department of Electrical and Electronics Engineering, Chung-Ang University, Seoul 06974, South Korea

Corresponding author: Yue Xiao (xiaoyue@uestc.edu.cn)

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ABSTRACT In this paper, a space-time line code (STLC) is combined with offset spatial modulation (OSM) for reaping both their advantages toward efficient multiple-input multiple-output (MIMO) transmission. Specifically, the spatial diversity offered by STLC increases the resilience against fading channels, while the simplified radio frequency (RF) switching processing merits of OSM inherently are preserved. As a further advancement in the pursuit of enhancing the performance of index detection, angle rotation is conceived and optimized for various transmit conditions. Subsequently, based on the approximation of the Q -function, an upper bound for the average bit error rate (BER) of STLC-OSM is derived in the context of an equispaced (ES) angle arrangement. Eventually, we demonstrate that STLC-OSM achieves considerable BER performance while achieving a unique single RF structure with a low implementation cost. Simulation results confirm the tightness of the derived BER bound and the advantages of the proposed scheme.

INDEX TERMS Space-time line code (STLC), offset spatial modulation (OSM).

I. INTRODUCTION

Space-time line code (STLC) [1] constitutes a full spatial diversity design alternative to space-time block code (STBC) [2], where both single radio frequency (RF) chain at the transmitter and low-complexity detection on the receiver side can be realized in the presence of the channel state information (CSI) at the transmitter. In recent years, as a new promising space-time code, STLC has attracted substantial research interests [3], [4], [5], [6], [7], [8], [9], [10], [11], [12], [13]. Specifically, in contrast to [1], the authors of [3] extended STLC to massive multiple-input multiple-output (MIMO) and multi-user scenarios. Furthermore, the authors of [4] evaluated the bit error rate (BER) performance of STLC over Rayleigh channels. In [5], the authors presented

an STLC encoding/decoding scheme with four receive antennas for full rate. In [6], two multiplexed STLC streams were employed to achieve additional spatial multiplexing gain. Subsequently, on the basis of [6], the authors of [7] proposed a channel shuffling technique to boost performance. In [8], the authors proposed a differential STLC scheme to reduce the transmitter's need for CSI. For the sake of eliminating the need for partial channel information in STLC decoding, in [9], the authors proposed a novel blind decoding method upon machine learning. In [10], orthogonal frequency division multiplexing (OFDM) and the concept of STLC were jointly explored in the time and frequency domains. Moreover, in order to strengthen the security of information transmission, the authors of [11] and [12] inserted artificial noise (AN) into STLC signals. In the context of a two-way relay (TWR), the authors of [13] proposed a new TWR method with STLC to improve

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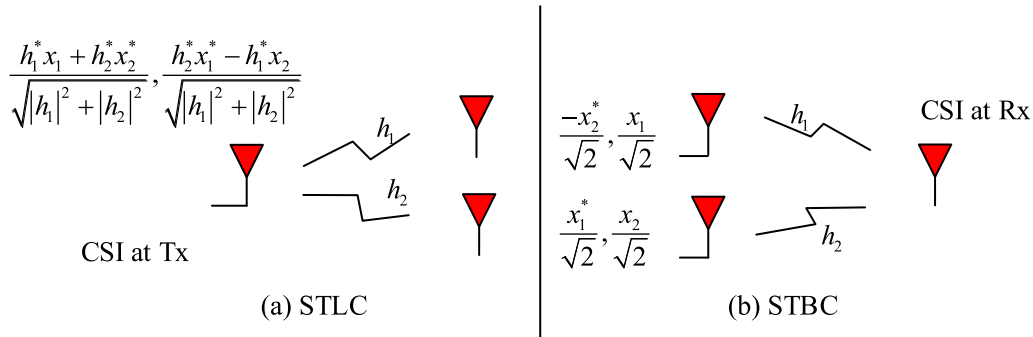


FIGURE 1. Space-time coding schemes: (a) STLC. (b) STBC.

the energy efficiency. In addition, as a benefit of single-RF transmission, it may be integrated with other single-RF MIMO techniques toward low-cost implementation in uplink transmission.

Meanwhile, as a particularly attractive alternative for single-RF MIMO transmission, spatial modulation (SM) [14], [15], [16] extends the antenna index for information modulation, while exhibiting high spectrum efficiency and low hardware costs. As a further advancement, to enhance the diversity gain of SM, current studies have explored the incorporation of space-time code (STC) [17], [18], [19]. In [17], space-time block coded spatial modulation (STBC-SM) was proposed with improvement in system performance. Correspondingly, the authors of [18] provided the performance analysis of STBC-SM over fading channels. By contrast, for the sake of maintaining the single RF chain, the authors of [19] proposed a novel space-time line coded spatial modulation (STLC-SM). However, SM has limitations in terms of frequent switching between the RF chain and the transmitting antenna, which directly affects the transmission rate and is therefore not conducive to the applications of future high-rate wireless communications [20]. Recently, in [21], the authors proposed a novel offset spatial modulation (OSM) to tackle the above-mentioned drawbacks of SM, while providing a performance improvement compared to SM. Specifically, the dynamic and static OSM systems were proposed in [21] to respond to different switching frequency requirements. For the former, a fixed antenna is selected to convey the symbols during transmission, so that the RF chain is fixed to this antenna, meaning that no RF switching is required, while in the latter, the activated antenna is with the maximum channel gain selected from a subset of antennas, and therefore the RF chain remains stationary during the coherence time. Furthermore, in [22], the OSM scheme under the channel estimation error scenario was considered, and the corresponding performance analysis was provided. However, the potential of transmit diversity in OSM has not yet been explored. Moreover, owing to the restriction of the activated antennas at the transmitter, STBC is not applicable to improve the diversity gain of OSM. Therefore, improving the diversity gain of OSM systems is still an open problem to be solved.

Against this backdrop, in this paper, a class of novel space-time line code aided offset spatial modulation (STLC-OSM) techniques are introduced for single-RF MIMO systems, in which the information is encoded jointly by STLC, the spatial index, the activated transmit antenna index and CSI. Furthermore, angle rotation [23], [24] is also tailored to the STLC-OSM signal to obtain more robust performance. Specifically, the optimal rotation angle is derived by maximizing the minimum Euclidean distance (ED). Furthermore, to circumvent the excessive complexity of the exhaustive search, a semi-definite relaxation (SDR) method is devised to obtain a sub-optimal rotation angle for STLC-OSM. In order to further reduce the computational complexity, angle rotation with an equispaced (ES) angle arrangement is applied to enhance the minimum ED statistically rather than dynamically. Furthermore, a theoretical derivation of the BER performance is presented to obtain insights. Afterward, with the channel estimation error, a comparison between STLC-OSM and the original OSM is performed to further verify the advantages of STLC-OSM. Finally, by computer simulations, we demonstrate the superiority of the proposed STLC-OSM with angle rotation over conventional STLC, OSM, and STBC-SM.

The contents of this paper are arranged as follows. In Section II, both the STLC and the proposed STLC-OSM systems are introduced in detail. The angle rotation and performance analysis for STLC-OSM are set out in Section III. Section IV describes the STLC-OSM scheme with imperfect channel estimation. In Section V, we present our numerical results and comparison to validate the superiority of the proposed system, before concluding in Section VI.

Notation: We use boldface variables to denote matrices and vectors. $(\cdot)^*$, $(\cdot)^T$, $(\cdot)^H$, and $\lfloor \cdot \rfloor$ denote conjugate, transpose, Hermitian transpose, and the floor function, respectively. $\mathcal{N}(m, \sigma^2)$ and $\mathcal{CN}(m, \sigma^2)$ denote the real and complex Gaussian distribution with mean $E\{\cdot\} = m$ and variance $D\{\cdot\} = \sigma^2$, respectively. $\Re(\cdot)$ and $\Im(\cdot)$ are the real and imaginary elements of a complex number, respectively. $\text{diag}(\cdot)$, $\|\cdot\|$ and $|\cdot|$ represent the transformation of a vector into a diagonal matrix, the norm and the absolute value, respectively. Furthermore, $\mathbf{I}_N$ is the $N \times N$ identity matrix, whereas $\Omega_j = \mathbf{I}_N[j]$ is the column j th from $\mathbf{I}_N$.

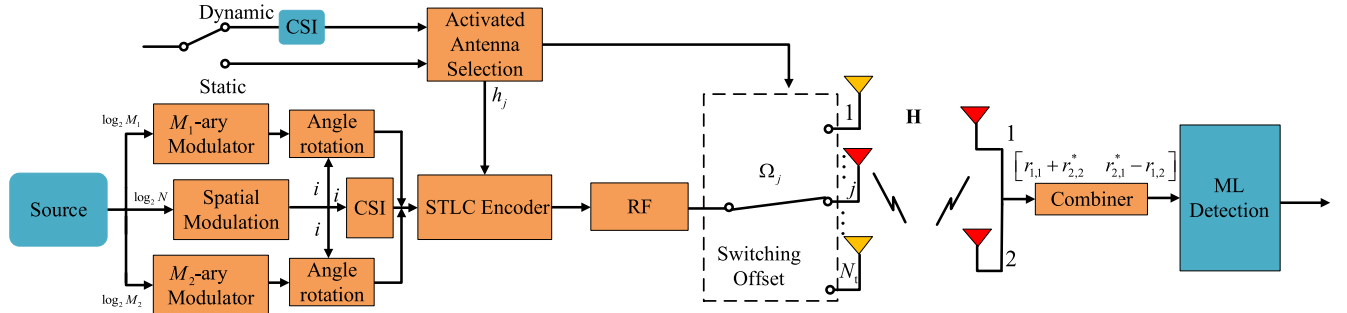


FIGURE 2. Block diagram of the proposed STLC-OSM system model where the active antennas are marked in red.

TABLE 1. Symbols and nomenclatures.

Symbol	nomenclature
s_t	The transmitted symbols, $t \in \{1, 2\}$
h	Channel fading coefficient
η	A normalization factor
χ	A predefined antenna index subset
p	A predefined antenna index subset χ containing p elements
q	The activated transmit antenna index is chosen by maximizing the effective channel gain
M	Modulation order
i	Spatial index
j	A randomly selected fixed antenna for static STLC-OSM
d	The distances between constellation points

Finally, $\mathbb{C}$ and $\mathbb{B}$ are the complex number field and a binary element matrix, respectively.

II. SYSTEM MODEL

A. SPACE-TIME LINE CODE

Consider a 1×2 STLC system [1], whose comparison with the original STBC system model is shown in Fig. 1. Under the assumption of a slow Rayleigh fading channel, the symbols transmitted over two consecutive symbol cycles are performed as

$$\begin{aligned} s_1 &= \eta(h_1^* x_1 + h_2^* x_2^*), \\ s_2 &= \eta(h_2^* x_1^* - h_1^* x_2), \end{aligned} \quad (1)$$

where h_l indicates the fading coefficient of the link from the transmit antenna to the l th receive antenna, with $h_l \sim \mathcal{CN}(0, 1)$, while x_1 and x_2 are the conventional M_1 -ary and M_2 -ary quadrature-amplitude modulation (QAM) or phase-shift keying (PSK) symbols, respectively. Moreover, η is a normalization factor to conform

$$E|s_t|^2 = \sigma_x^2, t \in \{1, 2\}. \quad (2)$$

Thus, we have

$$\eta = \frac{1}{\sqrt{|h_1|^2 + |h_2|^2}}. \quad (3)$$

In the context of the above transmission process, the spectral efficiency of STLC can be calculated as

$$e_0 = \frac{\log_2 M_1 + \log_2 M_2}{2}. \quad (4)$$

Then, the received signals can be expressed as

$$\begin{bmatrix} r_{1,1} & r_{1,2} \\ r_{2,1} & r_{2,2} \end{bmatrix} = \begin{bmatrix} h_1 \\ h_2 \end{bmatrix} \begin{bmatrix} s_1 & s_2 \end{bmatrix} + \begin{bmatrix} n_{1,1} & n_{1,2} \\ n_{2,1} & n_{2,2} \end{bmatrix}, \quad (5)$$

where $r_{l,t}$ is the signal received at time t of rx antenna index l , and $n_{l,t}$ represents additive white Gaussian noise (AWGN), as $n_{l,t} \sim \mathcal{CN}(0, \sigma_n^2)$.

With the STLC process, the received signal can be rephrased as

$$\underbrace{\begin{bmatrix} r_{1,1} + r_{2,2}^* \\ r_{2,1}^* - r_{1,2} \end{bmatrix}}_{\mathbf{y}} = \frac{1}{\eta} \underbrace{\begin{bmatrix} x_1 \\ x_2 \end{bmatrix}}_{\mathbf{x}} + \underbrace{\begin{bmatrix} n_{1,1} + n_{2,2}^* \\ n_{2,1}^* - n_{1,2} \end{bmatrix}}_{\mathbf{n}}. \quad (6)$$

Then, according to the maximum likelihood (ML) criterion, the symbols after detection can be expressed as

$$\hat{\mathbf{x}} = \arg \min_{\mathbf{x} \in \mathbb{S}} \left\| \mathbf{y} - \frac{1}{\eta} \mathbf{x} \right\|_2^2, \quad (7)$$

where $\mathbb{S}$ is the set of all possible symbols.

It can be revealed from (7), in contrast to STBC [2], STLC only requires the effective channel gain $\sqrt{|h_1|^2 + |h_2|^2}$ rather than full CSI at the receiver for symbols detection, where $\sqrt{|h_1|^2 + |h_2|^2}$ can be estimated by the blind SNR estimation techniques [25]. Furthermore, for M -PSK symbols detection, effective channel gain is not desired.

B. SPACE-TIME LINE CODE AIDED OFFSET SPATIAL MODULATION

It was shown that the spectrum efficiency of the traditional STLC system is $e_0 = \frac{\log_2 M_1 + \log_2 M_2}{2}$ in (4). As a consequence, in the context of the fixed modulation order, the spectrum efficiency of conventional STLC systems is constant.

Motivated by the above analysis, we propose an STLC-OSM system as portrayed in Fig.2 to obtain higher spectral efficiency while maintaining a single RF structure. Specifically, for different RF switching scenarios, just like OSM, two modes of STLC-OSM are presented as follows.

1) STATIC STLC-OSM [21]

A randomly selected fixed antenna j is activated to convey the information during the transmission, i.e., the RF chain

is attached to a fixed antenna so that there is no need for RF switching at all.

2) DYNAMIC STLC-OSM [21]

The activated transmit antenna index is chosen by maximizing the effective channel gain, expressed as

$$q = \arg \max_{k \in \chi} \frac{1}{\eta_k}, \quad (8)$$

where χ denotes a predefined antenna index subset containing p elements, $1 \leq p \leq N_t$, and

$$\eta_k = \frac{1}{\sqrt{|h_{1,k}|^2 + |h_{2,k}|^2}}, \quad (9)$$

where $h_{1,k}$ and $h_{2,k}$ indicate the channel gain of the active link.

For STLC-OSM, the same antenna index is utilized to convey the symbols during two transmission cycles. Hence, the spectral efficiency can be calculated as

$$e_l = \frac{\log_2 M_1 + \log_2 M_2 + \log_2 N_t}{2}, \quad (10)$$

where $\log_2 N_t$ bits are used to choose spatial index i , while N_t is a power of two. Compared to the STLC system, the proposed STLC-OSM system provides an increment of $\frac{1}{2} \log_2 N_t$ bits.

After i and q are confirmed, for STLC-OSM, the super symbols at the two consecutive time slots can be given by

$$\begin{aligned} s_1^{i,q} &= \Omega_q s_{i,q,1}, \\ s_2^{i,q} &= \Omega_q s_{i,q,2}, \end{aligned} \quad (11)$$

where $s_{i,q,1}$ and $s_{i,q,2}$ are encoded STLC symbols represented as

$$\begin{aligned} s_{i,q,1} &= \beta \frac{\eta_q}{\eta_i} (h_{1,q}^* x_1 + h_{2,q}^* x_2^*), \\ s_{i,q,2} &= \beta \frac{\eta_q}{\eta_i} (h_{2,q}^* x_1^* - h_{1,q}^* x_2), \end{aligned} \quad (12)$$

where the corresponding normalization factor β satisfying $E|s_{i,q,1}|^2 = E|s_{i,q,2}|^2 = \sigma_x^2$ can be calculated as

$$\beta = \frac{1}{\sqrt{E \left(\frac{\eta_q^2}{\eta_i^2} \right)}}. \quad (13)$$

The specific calculation of (13) is provided in Appendix.

Similar to the STLC system, the reconstructed receive symbols of the STLC-OSM system can be given by

$$\underbrace{\begin{bmatrix} r_{1,1} + r_{2,2}^* \\ r_{2,1}^* - r_{1,2} \end{bmatrix}}_{\mathbf{y}} = \beta \frac{1}{\eta_i} \underbrace{\begin{bmatrix} x_1 \\ x_2 \end{bmatrix}}_{\mathbf{x}} + \underbrace{\begin{bmatrix} n_{1,1} + n_{2,2}^* \\ n_{2,1}^* - n_{1,2} \end{bmatrix}}_{\mathbf{n}}, \quad (14)$$

where $\mathbf{n}$ denotes the equivalent AWGN, $n_{1,1} + n_{2,2}^*$ and $n_{2,1}^* - n_{1,2}$ follow $\mathcal{CN}(0, 2\sigma_n^2)$.

Then, it can be inferred from (14) that the detected data symbols and antenna index using the optimum ML algorithm

are given by

$$(\hat{\mathbf{x}}, \hat{i}) = \arg \min \left\| \mathbf{y} - \beta \frac{1}{\eta_i} \mathbf{x} \right\|^2. \quad (15)$$

In general, for OSM, the spatial index is detected by distinguishing the channel characteristics at the receiver. However, as seen from (15), the spatial index is detected only by the channel gain for STLC-OSM, which deteriorates the overall system performance. Therefore, as a remedy, angle rotation is designed in the next portion to tackle the above challenge and further boost the system's performance.

III. ANGLE ROTATION AND PERFORMANCE ANALYSIS

A. ANGLE ROTATION DESIGN

Under the premise of maintaining the special structure of the STLC receiver, angle rotation is devised as follows.

With the aid of angle rotation, the encoded symbols in (12) can be reconstructed as

$$\begin{aligned} s_{i,q,1} &= \beta \frac{\eta_q}{\eta_i} (h_{1,q}^* x_1 e^{j\theta_{1,i}} + h_{2,q}^* x_2^* e^{-j\theta_{2,i}}), \\ s_{i,q,2} &= \beta \frac{\eta_q}{\eta_i} (h_{2,q}^* x_1^* e^{-j\theta_{1,i}} - h_{1,q}^* x_2 e^{j\theta_{2,i}}), \end{aligned} \quad (16)$$

where $\theta_{l,i}$ is the designed rotation angle associated with the spatial index i and modulation order M_l , $e^{j\theta_{l,i}} \in \Phi_l = [e^{j\theta_{l,1}}, \dots, e^{j\theta_{l,N_t}}]^T$, $l \in \{1, 2\}$.

Hence, at the receiver, the reconstructed symbols can be shown as

$$\underbrace{\begin{bmatrix} r_{1,1} + r_{2,2}^* \\ r_{2,1}^* - r_{1,2} \end{bmatrix}}_{\mathbf{y}} = \beta \frac{\Psi_i}{\eta_i} \underbrace{\begin{bmatrix} x_1 \\ x_2 \end{bmatrix}}_{\mathbf{x}} + \underbrace{\begin{bmatrix} n_{1,1} + n_{2,2}^* \\ n_{2,1}^* - n_{1,2} \end{bmatrix}}_{\mathbf{n}}, \quad (17)$$

where $\Psi_i = \text{diag}([e^{j\theta_{1,i}}, e^{j\theta_{2,i}}])$.

The corresponding detected data with ML can be given by

$$(\hat{\mathbf{x}}, \hat{i}) = \arg \min \left\| \mathbf{y} - \beta \frac{\Psi_i}{\eta_i} \mathbf{x} \right\|_2^2. \quad (18)$$

In comparison to (15), the rotation angle is supplemented to the detection of the spatial index in (18) without affecting the original structure.

B. ROTATION ANGLE OPTIMIZATION

It can be inferred from (18) that the instantaneous conditional pairwise error probability (PEP) can be calculated as

$$\begin{aligned} P\left(\mathbf{x} \rightarrow \hat{\mathbf{x}}, \lambda / \frac{1}{\eta}\right) &= P\left\{ \left\| \mathbf{y} - \beta \frac{\Psi_i}{\eta_i} \mathbf{x} \right\|^2 > \left\| \mathbf{y} - \beta \frac{\Psi_{\hat{i}}}{\eta_{\hat{i}}} \hat{\mathbf{x}} \right\|^2 \right\} \\ &= P\left\{ -\Re \left\{ \left(\beta \left(\frac{\Psi_i}{\eta_i} \mathbf{x} - \frac{\Psi_{\hat{i}}}{\eta_{\hat{i}}} \hat{\mathbf{x}} \right) \right)^H \mathbf{n} \right\} \right. \\ &\quad \left. > \frac{\left\| \beta \left(\frac{\Psi_i}{\eta_i} \mathbf{x} - \frac{\Psi_{\hat{i}}}{\eta_{\hat{i}}} \hat{\mathbf{x}} \right) \right\|^2}{2} \right\}, \end{aligned} \quad (19)$$

where $\lambda = \frac{\sigma_x^2}{\sigma_n^2}$.

Then, according to the distribution of the element of $\mathbf{n}$, we have

$$-\Re \left\{ \left(\beta \left(\frac{\Psi_i}{\eta_i} \mathbf{x} - \frac{\Psi_{\hat{i}}}{\eta_{\hat{i}}} \hat{\mathbf{x}} \right) \right)^H \mathbf{n} \right\} \sim \mathcal{N} \left(0, \sigma_n^2 \left\| \beta \left(\frac{\Psi_i}{\eta_i} \mathbf{x} - \frac{\Psi_{\hat{i}}}{\eta_{\hat{i}}} \hat{\mathbf{x}} \right) \right\|^2 \right). \quad (20)$$

Hence, with Q function, (19) can be restated as

$$P \left(\mathbf{x} \rightarrow \hat{\mathbf{x}}, \lambda / \frac{1}{\eta} \right) = Q \left(\sqrt{\frac{\lambda \beta^2 \left\| \left(\frac{\Psi_i}{\eta_i} \mathbf{x} - \frac{\Psi_{\hat{i}}}{\eta_{\hat{i}}} \hat{\mathbf{x}} \right) \right\|^2}{4}} \right) \leq Q \left(\sqrt{\frac{\lambda \beta^2 D_{\min}}{4}} \right), \quad (21)$$

where

$$D_{\min} = \min \left\| \left(\frac{\Psi_i}{\eta_i} \mathbf{x} - \frac{\Psi_{\hat{i}}}{\eta_{\hat{i}}} \hat{\mathbf{x}} \right) \right\|^2.$$

Furthermore, $D_{\min}$ can be restated as

$$D_{\min} = \left| \frac{e^{j\theta_{1,i}}}{\eta_i} x_1 - \frac{e^{j\theta_{1,\hat{i}}}}{\eta_{\hat{i}}} \hat{x}_1 \right|^2 + \left| \frac{e^{j\theta_{2,i}}}{\eta_i} x_2 - \frac{e^{j\theta_{2,\hat{i}}}}{\eta_{\hat{i}}} \hat{x}_2 \right|^2. \quad (22)$$

Without loss of generality, the rotation angle Φ_1 is optimized by maximizing

$$D_{1,\min} = \left| \frac{e^{j\theta_{1,i}}}{\eta_i} x_1 - \frac{e^{j\theta_{1,\hat{i}}}}{\eta_{\hat{i}}} \hat{x}_1 \right|^2,$$

which can be expressed as

$$(P1) : \max_{\Phi} D_{1,\min} \quad s.t. \quad 0 \leq \theta_{1,m} < 2\pi, m = 1, \dots, N_t. \quad (23)$$

Then, under various conditions, we solve the above problem as follows.

1) $N_T = 2$ AND M -PSK

In this case, without losing generality, the designed angle for the first antenna is set to 0 and $\theta_{1,1} = 0$. The optimal solution to problem (P1) can then be computed as follows

$$\Phi_{1,opt} = \arg \max \min \{d_1, d_2, d_3\} \quad 0 \leq \theta_{1,2} \leq \frac{2\pi}{M}, \quad (24)$$

where d_1 , d_2 and d_3 denote the distances between constellation points at the receiver, expressed as

$$d_1 = \frac{1}{\eta_1} (|x_1 - \hat{x}_1|^2), \quad d_2 = \frac{1}{\eta_2} (|x_1 - \hat{x}_1|^2), \quad (25)$$

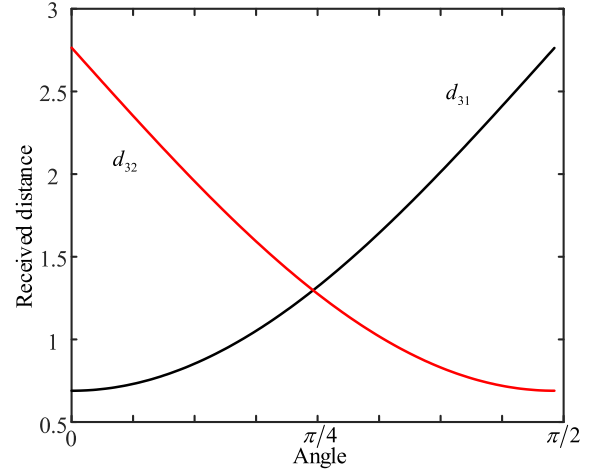


FIGURE 3. Received distance versus the designed rotation angle with QPSK modulation.

and

$$d_3 = \left| \frac{1}{\eta_1} x_1 - \frac{e^{j\theta_{1,2}}}{\eta_2} \hat{x}_1 \right|^2. \quad (26)$$

Since d_1 and d_2 are independent of $\theta_{1,2}$, (23) can be restated as

$$\Phi_{opt} = \arg \max \min \{d_3\} \quad s.t. \quad 0 \leq \theta_{1,2} \leq \frac{2\pi}{M}. \quad (27)$$

Due to the symmetry of M -PSK constellation points, we consider only the adjacent constellation points. The phase of the adjacent constellation points are $e^{j(2m\pi/M)}$ and $e^{j(2(m+1)\pi/M)}$. Thus, (26) can be split into

$$d_{31} = \left| \frac{1}{\eta_1} - \frac{e^{j\theta_{1,2}}}{\eta_2} \right|^2, \quad (28)$$

and

$$d_{32} = \left| \frac{e^{j(2\pi/M)}}{\eta_1} - \frac{e^{j\theta_{1,2}}}{\eta_2} \right|^2. \quad (29)$$

Then, (26) can be reformulated as

$$\Phi_{opt} = \arg \max \min \{d_{31}, d_{32}\} \quad s.t. \quad 0 \leq \theta_{1,2} \leq \frac{2\pi}{M}. \quad (30)$$

As can be seen from Fig. 3, the optimal solution of Φ is required to fulfill the restriction of $d_{31} = d_{32}$. Therefore, we arrive at

$$\left| \frac{1}{\eta_1} - \frac{e^{j\theta_{1,2}}}{\eta_{1,2}} \right|^2 = \left| \frac{e^{j(2\pi/M)}}{\eta_1} - \frac{e^{j\theta_{1,2}}}{\eta_2} \right|^2. \quad (31)$$

Thus, we can obtain

$$\theta_{1,2,opt} = \frac{\pi}{M}. \quad (32)$$

Finally, the optimal solution of Φ can be expressed as

$$\Phi_{1,opt} = [1 \quad e^{j\frac{\pi}{M}}]^T. \quad (33)$$

Algorithm 1 Additional Procedure for the SDR Method**Require:** $\mathbf{V}_{opt}$.**Ensure:** Φ_1

- 1: Initially, the eigenvalue decomposition of $\mathbf{V}_{opt}$ can be expressed as $\mathbf{V}_{opt} = \mathbf{U}\Sigma\mathbf{U}^H$, where $\mathbf{U}$ is a unitary matrix, and Σ is a diagonal matrix with the eigenvalues of $\mathbf{V}_{opt}$;
- 2: For problem (P2), the suboptimal solution Φ_{subopt} can be calculated as $\Phi_{subopt} = \mathbf{U}\Sigma^{1/2}\mathbf{r}$, where $\mathbf{r} \in \mathbb{C}^{N_t \times 1}$ denotes a random complex Gaussian vector, $\mathbf{r} \sim \mathcal{CN}(0, 1)$;
- 3: According to [27], a set $\mathbf{R}$ containing a sufficiently large number of realizations of $\mathbf{r}$ is generated to achieve a $\frac{\pi}{4}$ -approximation of the optimal solution of problem (P2), $\mathbf{R} = \{\mathbf{r}_1, \dots, \mathbf{r}_N\}$;
- 4: **for** $n = 1 : N$ **do**
- 5: $\Phi_{subopt}(n) = \mathbf{U}\Sigma^{1/2}\mathbf{R}(n)$;
- 6: $D_{1,\min}(n) = \min_{\mathbf{c}} |\mathbf{G}diag(\Phi_1)\mathbf{c}|^2$;
- 7: **end for**
- 8: $\hat{n} = \arg \max (D_{1,\min})$;
- 9: $\Phi_1 = \Phi_{subopt}(\hat{n})$.

2) $N_T > 2$ AND HIGHER MODULATION ORDERS

From (23), it is hard to achieve the optimal angle $\Phi_{1,opt}$ for large N_t and higher modulation orders. Therefore, we first apply the SDR method to obtain a sub-optimal solution. Then, an ES angle arrangement is applied to further reduce the complexity with satisfactory BER performance. The details are as follows.

a: THE SEMI-DEFINITE RELAXATION (SDR) METHOD

In this case, $D_{1,\min}$ can be restated as

$$D_{1,\min} = |\mathbf{G}diag(\Phi_1)\mathbf{c}|^2, \quad (34)$$

where $\mathbf{G} = \left[\frac{1}{\eta_1}, \dots, \frac{1}{\eta_{N_t}} \right] \in \mathbb{R}^{1 \times N_t}$, and $\mathbf{c} = \Omega_i \times x_1 - \Omega_{\hat{i}} \times \hat{x}_1$.

By incorporating an auxiliary variable t , (P1) can be restated as

$$\begin{aligned} (P2) : \max_{\Phi} t \\ \text{s.t. } |\mathbf{G}diag(\Phi_1)\mathbf{c}|^2 \geq t \quad \forall \mathbf{c} \\ 0 \leq \theta_{1,m} < 2\pi, m = 1, \dots, N_t. \end{aligned} \quad (35)$$

As a further advancement, $|\mathbf{G}diag(\Phi_1)\mathbf{c}|^2$ can be simplified upon using the knowledge of matrix theory as

$$\begin{aligned} |\mathbf{G}diag(\Phi)\mathbf{c}|^2 &= \left| \Phi^T diag(\mathbf{G})\mathbf{c} \right|^2 \\ &= \text{tr}(\mathbf{F}\Phi^*\Phi^T), \end{aligned} \quad (36)$$

where $\mathbf{F} = (diag(\mathbf{G})\mathbf{c})(diag(\mathbf{G})\mathbf{c})^H$.

Let $\mathbf{V} = \Phi^*\Phi^T$. Then, we have $\mathbf{V} \succeq 0$ with $\text{rank}(\mathbf{V}) = 1$. Since the non-convexity of the rank-one constraint, the SDR method is used to relax this constraint to transform

problem (P2) into a convex program. Consequently, problem (P2) can be reformed as

$$\begin{aligned} (P3) : \max_{\Phi} t \\ \text{s.t. } \text{tr}(\mathbf{F}\mathbf{V}) \geq t \quad \forall \mathbf{F} \\ \mathbf{V}(m, m) = 1, m = 1, \dots, N_t \\ \mathbf{V} \succeq 0. \end{aligned} \quad (37)$$

Furthermore, by available convex optimization solvers such as CVX [26], problem (P3) can be solved optimally. However, the solution for problem (P3) may not satisfy the restriction of $\text{rank}(\mathbf{V}) = 1$ in problem (P2), which means that the optimal solution of problem (P3) serves as an upper bound on that of the problem (P2). As a remedy, on the basis of the optimal solution $\mathbf{V}_{opt}$ of the problem (P3), additional procedures are supplemented to achieve a rank-one solution. The additional procedure is summarized in Algorithm 1.

b: THE EQUISPACED (ES) ANGLE ARRANGEMENT

Motivated by the optimal rotation angle $\Phi_{1,opt} = [1 e^{j\frac{\pi}{M}}]^T$ in (33) and the angle alignment in [24], we utilize an ES angle arrangement to enhance the minimum ED of STLC-OSM statistically as a trade-off between complexity and performance.

The angles are taken from an ES angle arrangement within $[0, 2\pi)$, which is given by

$$\theta_{1,m} = \frac{2\pi}{N_t M} (m-1), m = 1, \dots, N_t. \quad (38)$$

As will be shown in Section IV, compared to STLC-OSM-SDR, STLC-OSM-ES can deliver lower complexity while achieving satisfactory BER performance.

C. BER PERFORMANCE ANALYSIS FOR STLC-OSM-ES

Based on the above rotation angle optimization, it is difficult to obtain the theoretical BER performance for STLC-OSM with an optimal rotation angle since it is related to the instantaneous channel. Thus, we primarily concentrate on the performance analysis for STLC-OSM-ES.

With the assistance of [28], a BER upper bound of the STLC-OSM system can be quantified by

$$\begin{aligned} P_{bit} &\leq \frac{1}{k_2 \times 2^{k_2}} \times \sum_{\mathbf{x}} \sum_{\mathbf{x} \neq \hat{\mathbf{x}}} [d(\mathbf{x} \rightarrow \hat{\mathbf{x}}) \\ &\quad \times E \left\{ P \left(\mathbf{x} \rightarrow \mathbf{x}, \lambda / \frac{1}{\eta} \right) \right\}], \end{aligned} \quad (39)$$

where $k_2 = 2k_1$, and $d(\mathbf{x} \rightarrow \hat{\mathbf{x}})$ is the Hamming distance between $\mathbf{x}$ and its detected counterpart $\hat{\mathbf{x}}$.

According to the different types of symbol errors, $z = \left\| \left(\frac{\Psi_i}{\eta_i} \mathbf{x} - \frac{\Psi_{\hat{i}}}{\eta_{\hat{i}}} \hat{\mathbf{x}} \right) \right\|^2$ can be rewritten as

$$z = \begin{cases} \frac{1}{\eta^2} (|x_1 - \hat{x}_1|^2 + |x_2 - \hat{x}_2|^2) & \text{for } \varepsilon_1, \\ \left| \frac{e^{j\theta_{1,i}}}{\eta_i} x_1 - \frac{e^{j\theta_{1,\hat{i}}}}{\eta_{\hat{i}}} \hat{x}_1 \right|^2 + \left| \frac{e^{j\theta_{2,i}}}{\eta_i} x_2 - \frac{e^{j\theta_{2,\hat{i}}}}{\eta_{\hat{i}}} \hat{x}_2 \right|^2 & \text{for } \varepsilon_2, \end{cases} \quad (40)$$

where ε_1 denotes the case in which only the conventional symbol is incorrectly detected, and ε_2 denotes the case of incorrect index detection.

Observed in (39) and (40), the marginal probability density function (pdf) of $\frac{1}{\eta_i}$ and the joint pdf of each pair between $\frac{1}{\eta_i}$ and $\frac{1}{\eta_{\bar{i}}}$ are required for deriving the average PEP.

Since $h_{1,i}$ and $h_{1,\bar{i}}$ follow $\mathcal{CN}(0, 1)$, $\frac{1}{\eta}$ obeys the chi distribution with a degree of freedom 4. Hence, the PDF of $\frac{1}{\eta}$ is given by

$$f_{\frac{1}{\eta}}(x) = 2x^3 e^{-x^2}, x > 0. \quad (41)$$

Then, since $\frac{1}{\eta_i}$ and $\frac{1}{\eta_{\bar{i}}}$ are independent with the identical distribution, the joint pdf of each pair between $\frac{1}{\eta_i}$ and $\frac{1}{\eta_{\bar{i}}}$ can be computed by

$$f_{\frac{1}{\eta_i}, \frac{1}{\eta_{\bar{i}}}}(x, y) = 4x^3 y^3 e^{-x^2 - y^2}, x > 0, y > 0. \quad (42)$$

In the case of ε_1 , (23) can be rewritten as

$$P_{\varepsilon_1}(\mathbf{x} \rightarrow \hat{\mathbf{x}}, \lambda/\frac{1}{\eta}) = Q\left(\sqrt{\frac{\lambda\beta^2\delta}{4\eta^2}}\right), \quad (43)$$

where $\delta = (|x_1 - \hat{x}_1|^2 + |x_2 - \hat{x}_2|^2)$. In the context of $Q(x) \leq \frac{1}{12}e^{-x^2/2} + \frac{1}{4}e^{-2x^2/3}$ [29], the average PEP can be shown as

$$E\left\{P_{\varepsilon_1}\left(\mathbf{x} \rightarrow \hat{\mathbf{x}}, \lambda/\frac{1}{\eta}\right)\right\} \leq E_{\frac{1}{\eta}}\left[\frac{1}{12}e^{-\lambda\beta^2\delta/(8\eta^2)}\right] + E_{\frac{1}{\eta}}\left[\frac{1}{4}e^{-\lambda\beta^2\delta/(6\eta^2)}\right]. \quad (44)$$

Therefore, the right side of (44) can be computed by

$$\begin{aligned} & E_{\frac{1}{\eta}}\left[\frac{1}{12}e^{-\lambda\beta^2\delta/(8\eta^2)}\right] + E_{\frac{1}{\eta}}\left[\frac{1}{4}e^{-\lambda\beta^2\delta/(6\eta^2)}\right] \\ &= \int_0^{+\infty} \frac{1}{6}x^3 e^{-(1+\lambda\beta^2\delta/8)x^2} dx + \int_0^{+\infty} \frac{1}{2}x^3 e^{-(1+\lambda\beta^2\delta/6)x^2} dx \\ &= \frac{1}{12(1+\lambda\beta^2\delta/8)^2} + \frac{1}{4(1+\lambda\beta^2\delta/6)^2}. \end{aligned} \quad (45)$$

On the other hand, for the case of ε_2 , the average PEP of STLC-OSM can be bounded by

$$\begin{aligned} & E\left\{P_{\varepsilon_2}\left(\mathbf{x} \rightarrow \hat{\mathbf{x}}, \lambda/\frac{1}{\eta}\right)\right\} \\ & \leq E_{\frac{1}{\eta}}\left[\frac{1}{12}e^{-\frac{\lambda\beta^2\xi}{8}}\right] + E_{\frac{1}{\eta}}\left[\frac{1}{4}e^{-\frac{\lambda\beta^2\xi}{6}}\right], \end{aligned} \quad (46)$$

where

$$\xi = \left|\frac{e^{j\theta_{1,i}}}{\eta_i}x_1 - \frac{e^{j\theta_{1,\hat{i}}}}{\eta_{\hat{i}}}\hat{x}_1\right|^2 + \left|\frac{e^{j\theta_{2,i}}}{\eta_i}x_2 - \frac{e^{j\theta_{2,\hat{i}}}}{\eta_{\hat{i}}}\hat{x}_2\right|^2. \quad (47)$$

For simplicity, (46) can be restated as

$$E\left\{P_{\varepsilon_2}\left(\mathbf{x} \rightarrow \hat{\mathbf{x}}, \lambda/\frac{1}{\eta}\right)\right\} \leq g\left(\frac{1}{12}, \frac{1}{2}\right) + g\left(\frac{1}{4}, \frac{2}{3}\right), \quad (48)$$

where

$$\begin{aligned} & g(\alpha_1, \alpha_2) \\ &= E_{\frac{1}{\eta}}\left[\alpha_1 e^{-\frac{\alpha_2\lambda\beta^2}{4}\xi}\right] \\ &= \int_0^{+\infty} \int_0^{+\infty} 4\alpha_1 x^3 y^3 \times e^{-x^2 - y^2 - \frac{\alpha_2\lambda\beta^2}{4}\xi} dx dy. \end{aligned} \quad (49)$$

Let $a_1 = e^{j\theta_{1,i}}x_1$, $a_2 = e^{j\theta_{2,i}}x_2$, $b_1 = e^{j\theta_{1,\hat{i}}}\hat{x}_1$, and $b_2 = e^{j\theta_{2,\hat{i}}}\hat{x}_2$. Then, $g(\alpha_1, \alpha_2)$ can be calculated as

$$\begin{aligned} & g(\alpha_1, \alpha_2) \\ &= \int_0^{+\infty} \int_0^{+\infty} 4\alpha_1 x^3 y^3 \\ & \times e^{-x^2 - y^2 - \frac{\alpha_2\lambda\beta^2}{4}(|a_1x - b_1y|^2 + |a_2x - b_2y|^2)} dx dy. \end{aligned} \quad (50)$$

Let $x = r \cos(\theta)$ and $y = r \sin(\theta)$. Using straightforward mathematical operations, (50) can be rewritten as

$$\begin{aligned} & g(\alpha_1, \alpha_2) = \int_{\theta=0}^{\pi/2} \int_{r=0}^{+\infty} \frac{1}{2}\alpha_1 r^7 \sin^3(2\theta) \\ & \times e^{-r^2(1+c_1\cos^2(\theta)+c_2\sin^2(\theta)-c_3\sin(2\theta))} dr d\theta, \end{aligned} \quad (51)$$

where

$$c_1 = \frac{\alpha_2\lambda\beta^2}{4}(|a_1|^2 + |a_2|^2), \quad (52)$$

$$c_2 = \frac{\alpha_2\lambda\beta^2}{4}(|b_1|^2 + |b_2|^2), \quad (53)$$

and

$$c_3 = \frac{\alpha_2\lambda\beta^2}{4}(\Re(a_1b_1^*) + \Re(a_2b_2^*)). \quad (54)$$

Furthermore, for the sake of evaluating (51), the integral table in [30] is utilized, yielding

$$\int_0^{\infty} x^m e^{-\mu x^n} dx = \frac{\Gamma(v)}{n\mu^v} \quad v = \frac{m+1}{n} \quad (55)$$

$$\left[\Re(\mu) > 0 \Re(m) > 0 \Re(n) > 0\right],$$

where $\Gamma(\cdot)$ is the gamma function.

Then, we arrive at

$$g(\alpha_1, \alpha_2) = \frac{1}{2}\alpha_1 \int_{\theta=0}^{\pi/2} \frac{\sin^3(2\theta) \Gamma(4)}{2c^4} d\theta, \quad (56)$$

where $c = (1 + c_1\cos^2(\theta) + c_2\sin^2(\theta) - c_3\sin(2\theta))$.

For the integral in (56), we employ numerical methods to calculate its value.

i) *M - PSK*: For the case of both x_1 and x_2 being PSK symbols, (50) can be rewritten as

$$g(\alpha_1, \alpha_2) = \int_{\theta=0}^{\pi/2} \int_{r=0}^{+\infty} \frac{1}{2}\alpha_1 r^7 \sin^3(2\theta) \times e^{-r^2(u-v\sin(2\theta))} dx dy, \quad (57)$$

where

$$u = 1 + \frac{\alpha_2 \lambda \beta^2}{2}, v = \frac{\alpha_2 \lambda \beta^2}{4} (\Re(a_1 b_1^*) + \Re(a_2 b_2^*)).$$

Subsequently, we can obtain

$$\begin{aligned} g(\alpha_1, \alpha_2) &= \int_{\theta=0}^{\pi/2} \frac{1}{2} \alpha_1 \sin^3(2\theta) \frac{\Gamma(4)}{2(u - v \sin(2\theta))^4} d\theta \\ &= \frac{\alpha_1}{4(u^2 - v^2)^3} \left[4u^2 + 11v^2 \right. \\ &\quad \left. + \frac{3v(3u^2 + 2v^2)}{\sqrt{u^2 - v^2}} \left(\frac{\pi}{2} + \tan^{-1} \left(\frac{v}{\sqrt{u^2 - v^2}} \right) \right) \right]. \end{aligned} \quad (58)$$

Upon substituting (46) and (44) into (39), a BER upper bound of the STLC-OSM-ES system can be obtained.

Compared to a single RF chain OSM system with a transmit diversity of 1, the proposed STLC-OSM-ES introduces an amplitude-phase diversity. Moreover, in contrast to STBC-SM with transmit diversity of 2, STLC-OSM-ES can obtain an additional phase diversity, due to the existence of ES rotation angle [24]. Specifically, based on the BER analysis of STLC-OSM-ES in the previous subsection, we can obtain the range of diversity gain that can be achieved by STLC-OSM-ES.

In the case of ε_1 , it can be observed from (45) that the average PEP of STLC-OSM-ES varies as λ^{-2} at high SNRs. However, for the case of ε_2 , it can be inferred from (58) that the average PEP varies approximately at the λ^{-4} rate by using the approximation of $u \approx \frac{\alpha_2 \lambda \beta^2}{2}$ at high SNRs. This is a consequence of the fact that when the index is detected correctly, it remains the same as the STLC system diversity gain, and when the index is detected incorrectly, the additional phase rotation further boosts the system's diversity gain.

IV. PRACTICAL CHANNEL ESTIMATION MODEL

In a practical scenario, with the presence of channel estimation errors, the estimated channel available to both of the transmitter and receiver can be expressed as

$$\begin{aligned} \tilde{h}_{u,v} &= h_{u,v} + e_{u,v}, \\ u &\in \{1, 2\}, v \in \{1, 2, \dots, N_t\}, \end{aligned} \quad (59)$$

where $e_{u,v}$ is the channel error with complex Gaussian distribution, $\mathcal{CN}(0, \sigma_e^2)$.

In this case, with the estimated channel $\tilde{\mathbf{H}}$, the STLC-OSM symbols can be expressed as

$$\begin{aligned} s_{i,q,1} &= \tilde{\beta} \frac{\tilde{\eta}_q^2}{\tilde{\eta}_i} (\tilde{h}_{1,q}^* x_1 e^{j\theta_{1,i}} + \tilde{h}_{2,q}^* x_2 e^{-j\theta_{2,i}}), \\ s_{i,q,2} &= \tilde{\beta} \frac{\tilde{\eta}_q^2}{\tilde{\eta}_i} (\tilde{h}_{2,q}^* x_1 e^{-j\theta_{1,i}} - \tilde{h}_{1,q}^* x_2 e^{j\theta_{2,i}}), \end{aligned} \quad (60)$$

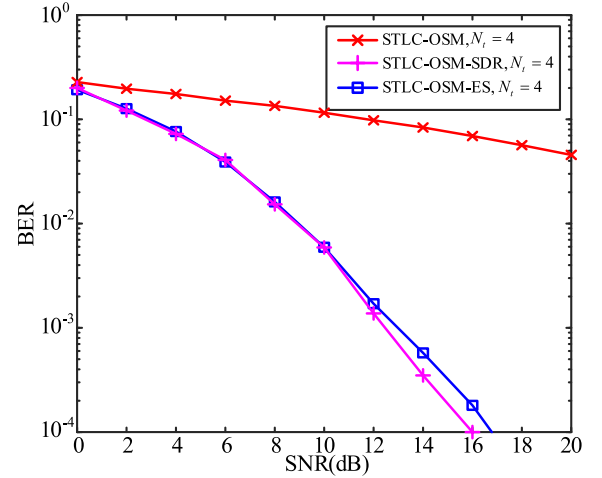


FIGURE 4. Performance comparison of STLC-OSM, STLC-OSM-SDR, and STLC-OSM-ES under BPSK transmission.

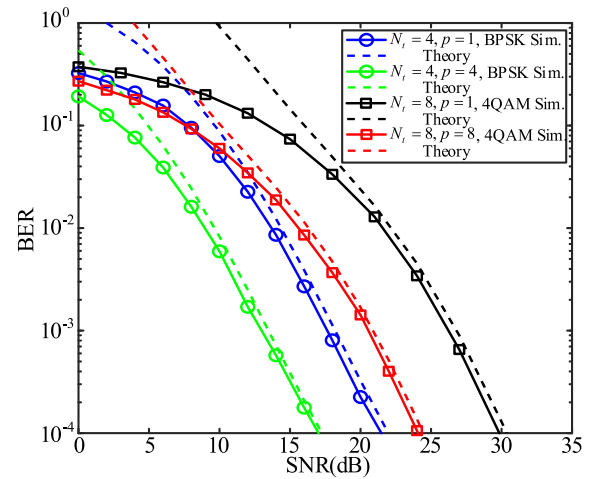


FIGURE 5. Simulation and theoretical results of static STLC-OSM-ES ($p=1$) and dynamic STLC-OSM-ES ($p=4$ and $p=8$).

where

$$\tilde{\eta}_k = \frac{1}{\sqrt{|\tilde{h}_{1,k}|^2 + |\tilde{h}_{2,k}|^2}}, \quad (61)$$

and

$$\tilde{\beta} = \frac{1}{\sqrt{E \left(\frac{\tilde{\eta}_q^2}{\tilde{\eta}_i^2} \right)}}. \quad (62)$$

The specific calculation of (62) is provided in Appendix. At the receiver, the corresponding recombination signal is

$$\underbrace{\begin{bmatrix} r_{1,1} + r_{2,2}^* \\ r_{2,1}^* - r_{1,2} \end{bmatrix}}_{\mathbf{y}} = \tilde{\beta} \frac{\Psi_i}{\tilde{\eta}_i} \underbrace{\begin{bmatrix} x_1 \\ x_2 \end{bmatrix}}_{\mathbf{x}} + \mathbf{U} + \underbrace{\begin{bmatrix} n_{1,1} + n_{2,2}^* \\ n_{2,1}^* - n_{1,2} \end{bmatrix}}_{\mathbf{n}}, \quad (63)$$

where

$$\mathbf{U} = \begin{bmatrix} -e_{1,q} s_{i,q,1} - e_{2,j}^* s_{i,q,2}^* \\ -e_{2,q}^* s_{i,q,1}^* + e_{1,j} s_{i,q,2} \end{bmatrix}. \quad (64)$$

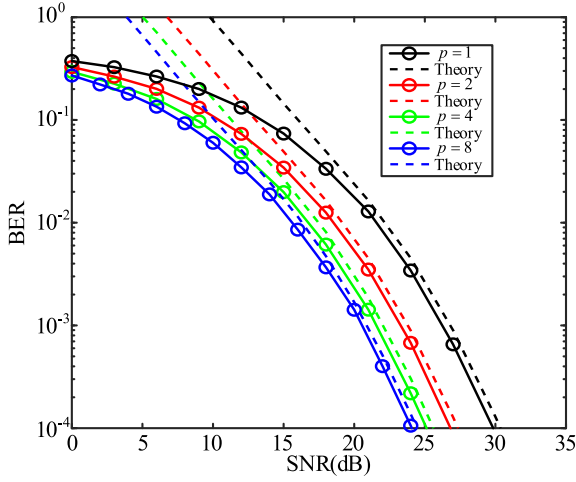


FIGURE 6. BER bound of STLC-OSM-ES with different p , $N_t = 8$, and 4QAM modulation.

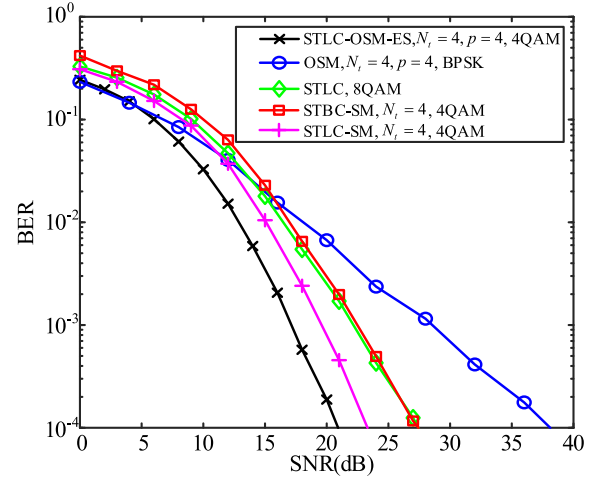


FIGURE 8. BER performance at 3bits/s/Hz for STLC-OSM-ES, STLC-SM, STBC-SM, OSM, and STLC.

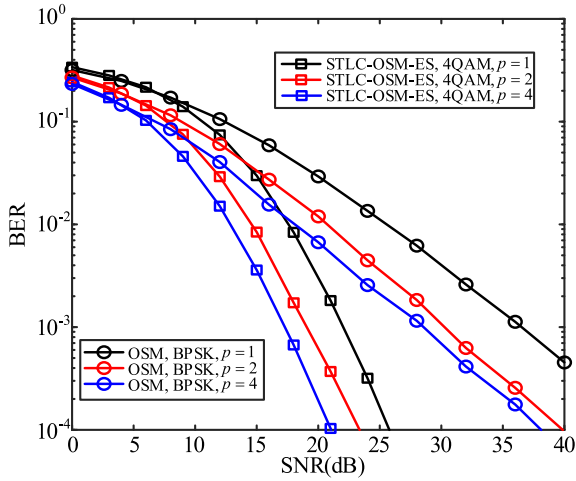


FIGURE 7. BER performance comparison of STLC-OSM-ES and OSM with different p and $N_t = 4$.

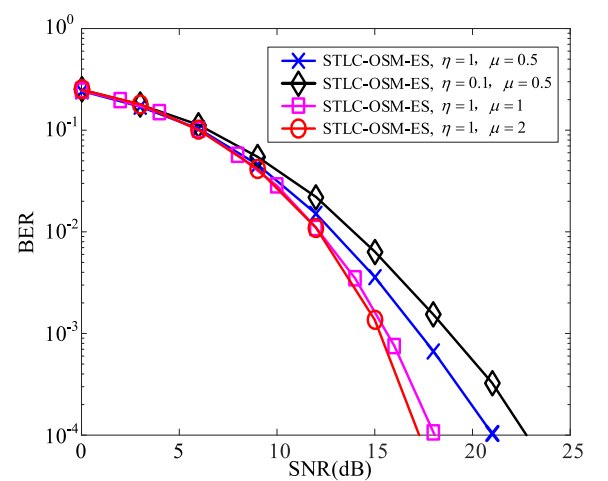


FIGURE 9. Performance comparison of STLC-OSM-ES under the α - η - μ channel fading with $N_t = 4$, 4QAM, $p = 4$ and $\alpha = 2$.

Finally, the detection symbols and antenna index can be detected as

$$(\hat{\mathbf{x}}, \hat{i}) = \arg \min \left\| \mathbf{y} - \tilde{\beta} \frac{\Psi_i}{\tilde{\eta}_i} \mathbf{x} \right\|_2^2. \quad (65)$$

V. SIMULATION RESULTS AND COMPARISONS

In this section, first, the superiority of angle rotation for STLC-OSM is evaluated by comparing the BER performance of STLC-OSM with and without rotation angle. Then, the simulation results of STLC-OSM with the SDR algorithm are presented and compared in conjunction with the ES angle arrangement. Afterward, the derived BER upper bounds for the static and dynamic STLC-OSM-ES are evaluated and compared in conjunction with link-level simulation results. Furthermore, considering conventional STLC [11], OSM [21], [22], STLC-SM [19], and STBC-SM [17] as benchmarks, the performance of the proposed STLC-OSM system with angle rotation is evaluated.

Finally, under imperfect CSI, the comparative performance of STLC-OSM-ES and OSM is evaluated. Note that $M_1 = M_2$ and $\text{SNR} = \gamma$ are used for the following simulation results.

Fig. 4 explores the BER comparisons among STLC-OSM, STLC-OSM-SDR, and STLC-OSM-ES. Observed in Fig. 4, both STLC-OSM-ES and STLC-OSM-SDR have a major SNR advantage over the STLC-OSM scheme without angel rotation. This is attributable to the enhancement of the minimum ED by angel rotation. Moreover, in comparison to STLC-OSM-SDR, STLC-OSM-ES has similar performance at low SNR region but with performance loss at high SNR, e.g., about 2 dB at $\text{BER} = 10^{-4}$, while significantly reducing the computational complexity.

The theoretical BER upper bounds of the proposed static and dynamic STLC-OSM-ES under different parameters are evaluated in Fig. 5, respectively. As can be observed in Fig. 5, dynamic STLC-OSM-ES has a significant

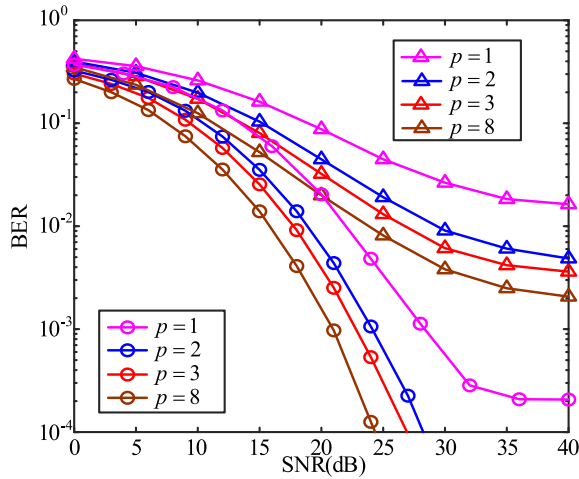


FIGURE 10. Performance comparison of STLC-OSM-ES and OSM with $N_t = 8$, 4QAM, and $\sigma_e^2 = 0.001$.

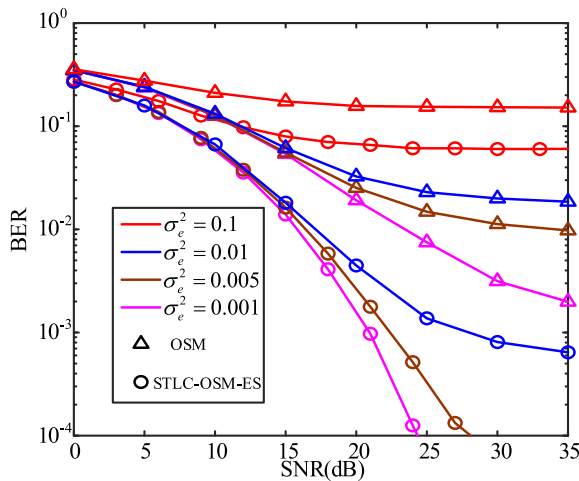


FIGURE 11. BER versus SNR in STLC-OSM-ES and OSM with $N_t = 8$, 4QAM, and $p = 8$.

performance improvement relative to static STLC-OSM-ES due to the activation of the antenna with maximum channel gain in the antenna subset. Moreover, in the high SNR region, the theoretical results approximately coincide with the simulation ones, which confirms the accuracy of the theoretical derivation.

In Fig. 6, we focus on the impact of different p on the BER performance of the proposed STLC-OSM-ES system. It can be seen from the results that the value of p is positively correlated with the BER performance, and thus a compromise between performance and RF switching can be achieved by setting the value of p . Moreover, the theoretical upper bounds roughly coincide the simulated counterpart, confirming the reliability of the mathematical derivation.

Fig. 7 presents a comparison of the BER performance of OSM and the proposed STLC-OSM-ES at different p for the same spectral efficiency. As expected, in comparison

to conventional OSM, the proposed STLC-OSM-ES can achieve higher diversity gain and better performance. More specifically, at $\text{BER} = 10^{-3}$, STLC-OSM-ES can obtain about 12dB SNR gains over OSM with $p = 4$.

Fig. 8 exhibits the performance of the proposed STLC-OSM-ES system in comparison to the counterparts of STLC, OSM, STLC-SM, and STBC-SM at 3 bits/s/Hz. As can be seen in Fig. 8, the combination of STLC and OSM further enhances the BER performance relative to the standard STLC. In addition, with the aid of CSI at the transmitter, STLC-OSM-ES provides a performance improvement in contrast to STBC-SM and STLC-SM. From the results we observe furthermore that, due to statistically significant improvement in minimum ED by the introduction of angle optimization, STLC-OSM-ES yields a higher diversity gain while reducing RF switching over STBC-SM at the same spectral efficiency. More specifically, when $\text{BER} = 10^{-4}$, the proposed STLC-OSM-ES system provides about 6dB SNR gain over STBC-SM, 2dB SNR gain over STLC-SM and 17dB SNR gain over OSM.

Fig. 9 exhibits the performance of the proposed STLC-OSM-ES system under the α - η - μ channel fading [32]. As can be seen in Fig. 9, the adaptability of the STLC-OSM-ES system to different channel environments can be visually appreciated. Specifically, for fixed values of α and η , the STLC-OSM-ES performs better as the value of μ increases, due to the fact that larger values of μ indicate smoother fading and less signal fluctuation. In addition, when $\mu = 0.5$, $\eta = 1$, the STLC-OSM-ES performance is the same in Fig. 8 and Fig. 9, which also implies that the α - η - μ channel transforms into a Rayleigh channel, which laterally verifies the accuracy of the simulation.

Fig. 10 and 11 characterize the BER performance comparison of STLC-OSM-ES and OSM with different parameter configurations under imperfect channel estimation. In Fig. 10, we focus on the effect of different p on the performance gap of STLC-OSM-ES and OSM. It can be observed that STLC-OSM-ES has a significant improvement over OSM. This is due to the increased resilience against fading channels by STLC and the further improvement of the minimum ED by angle rotation. In addition, Fig. 11 shows the considerable performance improvements of STLC-OSM-ES in comparison to OSM from the perspective of different σ_e^2 .

VI. CONCLUSION

A novel single-RF MIMO system, dubbed STLC-OSM, was implemented to integrate the merits of space-time line code (STLC) and offset spatial modulation (OSM). In order to boost the performance of the proposed STLC-OSM scheme, we applied angle rotation for various transmit conditions to enhance the minimum Euclidean distance without compromising the desirable low-complexity decoding structure. In the case of an equispaced (ES) angle arrangement, the

theoretical BER upper bound of STLC-OSM-ES was derived with the aid of an approximate equation of the Q function. By comparing the simulation results, both the proposed STLC-OSM-SDR and STLC-OSM-ES schemes were found to significantly increase the BER performance relative to OSM and STLC. In summary, compared to space-time block coded spatial modulation (STBC-SM), STLC-OSM exhibits better performance due to the advantages of OSM over SM and the introduction of ES arrangement. On the other hand, compared to STLC-SM, STLC-OSM provides a performance improvement while reducing RF switching. Furthermore, in the presence of imperfect channel estimation, STLC-OSM-ES offers more robust performance over original OSM.

Due to the characteristics of STLC-OSM, it may be suitable for future implementation in communications between high performance transmitters and low complexity receivers, such as Internet of Things (IoT) and wearable devices. Furthermore, in the future work, we will further consider the performance analysis of the proposed STLC-OSM under generalised channel fading. Moreover, We plan to conduct a more detailed analysis of the practical implementation of the STLC-OSM scheme, including an examination of the specific hardware requirements, computational complexity and potential interoperability issues that may arise.

APPENDIX FOOTNOTES

Here, for static and dynamic STLC-OSM, the exact β is calculated under perfect and imperfect channel estimation, respectively, as follows.

A. PERFECT CHANNEL ESTIMATION

1) β FOR STATIC STLC-OSM

In this case, the randomly selected fixed antenna j is irrelevant with the spatial index i , that is, $\frac{1}{\eta_i}$ and $\frac{1}{\eta_j}$ are independent. Therefore, we have

$$\begin{aligned} E\left(\frac{\eta_j^2}{\eta_i^2}\right) &= \frac{1}{N_t} + \frac{N_t - 1}{N_t} \int_{x=0}^{+\infty} \int_{y=0}^{+\infty} \frac{x^2}{y^2} f_{\frac{1}{\eta_i}}(x) f_{\frac{1}{\eta_q}}(y) dx dy \\ &= \frac{1}{N_t} + \frac{N_t - 1}{N_t} \int_{x=0}^{+\infty} \int_{y=0}^{+\infty} 4x^5 e^{-x^2} y e^{-y^2} dx dy \\ &= \frac{2N_t - 1}{N_t}, \end{aligned} \quad (66)$$

where $\frac{1}{N_t}$ and $\frac{N_t - 1}{N_t}$ denote the probabilities for the cases $i = j$ and $i \neq j$, respectively.

Therefore, β can be calculated as

$$\beta = \frac{1}{\sqrt{E\left(\frac{\eta_j^2}{\eta_i^2}\right)}} = \frac{1}{\sqrt{\frac{2N_t - 1}{N_t}}}. \quad (67)$$

2) β FOR DYNAMIC STLC-OSM

The q th antenna selected from a predetermined antenna set χ including p elements is assigned to transmit STLC-OSM symbols. Thus, the expectation of $\frac{\eta_q^2}{\eta_i^2}$ can be computed as

$$E\left(\frac{\eta_q^2}{\eta_i^2}\right) = \frac{p}{N_t} E_{in}\left(\frac{\eta_q^2}{\eta_i^2}\right) + \frac{N_t - p}{N_t} E_{out}\left(\frac{\eta_q^2}{\eta_i^2}\right), \quad (68)$$

where $\frac{p}{N_t}$ represents the probability of the case $i \in \chi$, E_{in} and E_{out} are the expectation of $\frac{\eta_q^2}{\eta_i^2}$ when $i \in \chi$ and $i \notin \chi$, respectively.

When $i \in \chi$, $\frac{1}{\eta_i}$ and $\frac{1}{\eta_q}$ are correlated variables. Therefore, we obtain

$$E_{in}\left(\frac{\eta_q^2}{\eta_i^2}\right) = \int_0^\infty \int_x^\infty \frac{x^2}{y^2} f_{\frac{1}{\eta_i}, \frac{1}{\eta_q}}(x, y) dx dy, \quad (69)$$

where $f_{\frac{1}{\eta_i}, \frac{1}{\eta_q}}(x, y)$ is the joint PDF of $\frac{1}{\eta_i}$ and $\frac{1}{\eta_q}$.

To obtain $f_{\frac{1}{\eta_i}, \frac{1}{\eta_q}}(x, y)$, we sort the elements of χ as $[o_1, \dots, o_p]$, $o_1 \leq o_2 \leq \dots \leq o_p$. Furthermore, according to [31], the joint PDF of any two elements o_u and o_v can be expressed as

$$\begin{aligned} f(o_u, o_v) &= \frac{p!}{(u-1)!(v-u-1)!(p-v)!} [1 - F(o_v)]^{p-v} \\ &\quad \cdot [F(o_v) - F(o_u)]^{v-u-1} F(o_u)^{u-1} f(o_u) f(o_v) \\ &\quad u < v. \end{aligned} \quad (70)$$

It can be inferred from (41) that $f(o_u) = 2o_u^3 e^{-o_u^2}$ and $F(o_u) = 1 - (1 + o_u^2) e^{-o_u^2}$. In dynamic STLC-OSM, we have $\frac{1}{\eta_q} = o_p$ and $\frac{1}{\eta_i} = o_u$. Therefore, for $\frac{1}{\eta_i} < \frac{1}{\eta_q}$, we can obtain the joint PDF of $\frac{1}{\eta_i}$ and $\frac{1}{\eta_q}$ as

$$\begin{aligned} f_{\frac{1}{\eta_i}, \frac{1}{\eta_q}}(x, y) &= \delta_{u,p} \left[\left(1 + x^2\right) e^{-x^2} - \left(1 + y^2\right) e^{-y^2} \right]^{p-u-1} \\ &\quad \cdot \left(1 - \left(1 + x^2\right) e^{-x^2}\right)^{u-1} x^3 e^{-x^2} y^3 e^{-y^2}, \end{aligned} \quad (71)$$

where $\delta_{u,p} = \frac{4p!}{(u-1)!(p-u-1)!}$.

By using the binomial expansion of the exponential term, (71) can be restated as

$$\begin{aligned} f_{\frac{1}{\eta_i}, \frac{1}{\eta_q}}(x, y) &= \delta_{u,p} \sum_{r=0}^{u-1} \sum_{m=0}^{p-u-1} \sum_{n=0}^m \sum_{l=0}^{p-u-m+r-1} (-1)^{m+r} \\ &\quad \times \varepsilon_1 y^{2n+3} e^{-(m+1)y^2} x^{2l+3} e^{-(p-u-m+r)x^2}, \end{aligned} \quad (72)$$

where

$$\begin{aligned} \varepsilon_1 &= \binom{u-1}{r} \binom{p-u-1}{m} \\ &\quad \times \binom{m}{n} \binom{p-u-m+r-1}{r}. \end{aligned} \quad (73)$$

Substituting (72) into (69), we arrive at

$$E_{in}\left(\frac{\eta_q^2}{\eta_i^2}\right) = \frac{1}{p} + \frac{1}{p} \sum_{u=1}^{p-1} \delta_{u,p} \sum_{r=0}^{u-1} \sum_{m=0}^{p-u-1} \sum_{n=0}^m \sum_{l=0}^{p-u-m+r-1} \times (-1)^{m+r} \varepsilon_1 \frac{1}{2(m+1)^{n+1}} R_1, \quad (74)$$

where

$$R_1 = \int_0^\infty \Gamma(n+1, (m+1)x^2) x^{2l+5} e^{-(p-u-m+r)x^2} dx. \quad (75)$$

Due to the difficulty in calculating the closed-form expression for the integral in (75), we obtain R_1 by a numerical method.

For the case of $i \notin \chi$, $\frac{1}{\eta_i}$ and $\frac{1}{\eta_q}$ are independent variables. Thus, we have

$$E_{out}\left(\frac{\eta_q^2}{\eta_i^2}\right) = \int_0^\infty \int_0^\infty \frac{x^2}{y^2} f_{\frac{1}{\eta_i}}(x) f_{\frac{1}{\eta_q}}(y) dx dy. \quad (76)$$

Since $\frac{1}{\eta_q}$ is the maximal element in the set χ , the PDF of $\frac{1}{\eta_q}$ is given by

$$f_{\frac{1}{\eta_q}}(y) = 2p \left(1 - (1+y^2)e^{-y^2}\right)^{p-1} y^3 e^{-y^2}. \quad (77)$$

Substituting (77) into (76), we arrive at

$$E_{out}\left(\frac{\eta_q^2}{\eta_i^2}\right) = p \sum_{m=0}^{p-1} \sum_{n=0}^m (-1)^m \varepsilon_2 \frac{\Gamma(n+1) \Gamma(3)}{(m+1)^{n+1}}, \quad (78)$$

where $\varepsilon_2 = \binom{p-1}{m} \binom{m}{n}$.

Then, by substituting (74) and (78) into (68), we get the expectation of $\frac{\eta_q^2}{\eta_i^2}$. Finally, β can be obtained.

B. IMPERFECT CHANNEL ESTIMATION

1) $\tilde{\beta}$ FOR STATIC STLC-OSM

In this case, according to (59), we have $\tilde{\mathbf{H}} \sim \mathcal{CN}(0, \sigma_e^2)$. Thus, the PDF of $\frac{1}{\eta_i}$ can be expressed as

$$f_{\frac{1}{\eta_i}}(x) = \frac{2}{(1+\sigma_e^2)^2} x^3 e^{-\frac{x^2}{1+\sigma_e^2}}. \quad (79)$$

Then, similar with Section VI, we have

$$E\left(\frac{\tilde{\eta}_j^2}{\tilde{\eta}_i^2}\right) = \frac{1}{N_t} + \frac{N_t-1}{N_t} \int_0^\infty \int_0^\infty \frac{x^2}{y^2} f_{\frac{1}{\eta_i}}(x) f_{\frac{1}{\eta_j}}(y) dx dy = \frac{2N_t-1}{N_t}. \quad (80)$$

Therefore, β can be calculated as

$$\beta = \frac{1}{\sqrt{E\left(\frac{\tilde{\eta}_j^2}{\tilde{\eta}_i^2}\right)}} = \frac{1}{\sqrt{\frac{2N_t-1}{N_t}}}. \quad (81)$$

It can be inferred from (67) and (81) that whether the channel estimation is perfect does not affect the value of β .

2) $\tilde{\beta}$ FOR DYNAMIC STLC-OSM

In the case, for $i \in \chi$, we can obtain the joint PDF of $\frac{1}{\eta_i}$ and $\frac{1}{\eta_q}$ as

$$f_{\frac{1}{\eta_i}, \frac{1}{\eta_q}}(x) = \tilde{\delta}_{u,p} \left[(1+\xi x^2) e^{-\xi x^2} - (1+\xi y^2) e^{-\xi y^2} \right]^{p-u-1} \cdot \left(1 - (1+\xi x^2) e^{-\xi x^2} \right)^{u-1} x^3 y^3 e^{-\xi x^2} e^{-\xi y^2}, \quad (82)$$

where $\xi = \frac{1}{(1+\sigma_e^2)}$ and $\tilde{\delta}_{u,p} = \frac{4p! \xi^4}{(u-1)!(p-u-1)!}$.

Furthermore, (82) can be recomputed as

$$f_{\frac{1}{\eta_i}, \frac{1}{\eta_q}}(x) = \tilde{\delta}_{u,p} \sum_{r=0}^{u-1} \sum_{m=0}^{p-u-1} \sum_{n=0}^m \sum_{l=0}^{p-u-m+r-1} (-1)^{m+r} \times \varepsilon_1 \xi^{n+l} y^{2n+3} x^{2l+3} e^{-(m+1)\xi y^2} e^{-(p-u-m+r)\xi x^2}. \quad (83)$$

Substituting (83) into (69), we can obtain

$$E_{in}\left(\frac{\tilde{\eta}_q^2}{\tilde{\eta}_i^2}\right) = \frac{1}{p} + \frac{1}{p} \sum_{u=1}^{p-1} \alpha_u \sum_{r=0}^{u-1} \sum_{m=0}^{p-u-1} \sum_{n=0}^m \sum_{l=0}^{p-u-m+r-1} \times (-1)^{m+r} \varepsilon_1 \xi^{l-1} \frac{1}{2(m+1)^{n+1}} R_2, \quad (84)$$

where

$$R_2 = \int_0^\infty \Gamma(n+1, (m+1)\xi x^2) x^{2l+5} e^{-(p-u-m+r)\xi x^2} dx. \quad (85)$$

Again, we obtain R_2 numerically due to the difficulty of calculating the closed expression for the integral in (85).

When $i \notin \chi$, the PDF of the maximal element $\frac{1}{\eta_q}$ in the set χ is given by

$$f_{\frac{1}{\eta_q}}(y) = 2p \xi^2 \left(1 - (1+\xi y^2) e^{-\xi y^2}\right)^{p-1} y^3 e^{-\xi y^2}. \quad (86)$$

Then, we have

$$E_{out}\left(\frac{\tilde{\eta}_q^2}{\tilde{\eta}_i^2}\right) = p \sum_{m=0}^{p-1} \sum_{n=0}^m (-1)^m \varepsilon_2 \frac{\Gamma(n+1) \Gamma(3)}{(m+1)^{n+1}}. \quad (87)$$

Finally, combining the above results we can get the value of $\tilde{\beta}$.

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YASHAN PANG received the Ph.D. degree from the National Key Laboratory of Wireless Communications, University of Electronic Science and Technology of China, Chengdu, China, in 2022.

He is currently an Engineer with the National Key Laboratory of Advanced Communication Networks, Academy of Network and Communications of CETC, Shijiazhuang, Hebei, China. His research interests include signal processing and wireless communications, with emphasis

on MIMO communication technology and practical issues in 5G and 6G systems.



YUE XIAO (Member, IEEE) received the Ph.D. degree in communication and information systems from the University of Electronic Science and Technology of China (UESTC), in 2007.

He is currently a Professor with the National Key Laboratory of Wireless Communications, UESTC. He has published more than 100 international journals and has been in charge of more than 20 projects in the area of Chinese 3G/4G/5G wireless communication systems. He is

an inventor of more than 50 Chinese and PCT patents on wireless systems. His research interests include system design and signal processing toward future wireless communication systems. He also serves as a Senior Associate Editor for *IEEE COMMUNICATIONS LETTERS* and *IEEE OPEN JOURNAL OF THE COMMUNICATIONS SOCIETY*.



XIA LEI (Member, IEEE) received the Ph.D. degree in communication and information systems from the University of Electronic Science and Technology of China, Chengdu, China, in 2005.

She is currently a Professor with the University of Electronic Science and Technology of China, where she is also the Supervisor of the Ph.D. degree candidates. She has published more than 40 international journals and has been in several projects. Her research interest includes wireless broadband communication systems.



YONG LIANG GUAN (Senior Member, IEEE) received the B.Eng. degree (Hons.) from the National University of Singapore, Singapore, and the Ph.D. degree from Imperial College London, London, U.K.

He is currently a Professor of communication engineering with the School of Electrical and Electronic Engineering, Nanyang Technological University, Singapore, where he leads the Continental-NTU Corporate Laboratory and the successful deployment of the campus-wide NTU-NXP V2X test bed. His current research interests include coding and signal processing for communication systems and data storage systems. He is also the Associate Vice President of NTU and a Distinguished Lecturer of the IEEE Vehicular Technology Society. He is an Editor of IEEE TRANSACTIONS ON VEHICULAR TECHNOLOGY.



JINGON JOUNG (Senior Member, IEEE) received the B.S. degree in radio communication engineering from Yonsei University, Seoul, South Korea, in 2001, and the M.S. and Ph.D. degrees in electrical engineering and computer science from KAIST, Daejeon, South Korea, in 2003 and 2007, respectively. He was a Postdoctoral Fellow with KAIST, in 2007; and UCLA, Los Angeles, CA, USA, in 2008. He was a Scientist with the Institute for Infocomm

Research (I²R), Agency for Science, Technology and Research (A*STAR), Singapore, from 2009 to 2015. He joined Chung-Ang University (CAU), Seoul, in 2016, as a Faculty Member, where he is currently a Professor with the School of Electrical and Electronics Engineering and the Principal Investigator of the Intelligent Wireless Systems Laboratory. His research interests include wireless communication signal processing, numerical analysis, algorithms, and machine learning. He was a recipient of the First Prize in the Intel-ITRC Student Paper Contest, in 2006, and the Research Pioneer Award of the Institute of Electronics and Information Engineering (IEIE), South Korea, in 2022. He was recognized as an Exemplary Reviewer of the IEEE COMMUNICATIONS LETTERS (2012) and the IEEE WIRELESS COMMUNICATIONS LETTERS (2012–2014 and 2019). He served as the Guest Editor/Associate Editor for several journals, including IEEE ACCESS (2016), the *Electronics* (2019), the *APSIPA Transactions on Signal and Information Processing* (2014–2019), the *Sensors* (2020), and IEEE TRANSACTIONS ON VEHICULAR TECHNOLOGY (2018–2023). He is an Inventor of a space-time line code (STLC), a fully symmetric scheme to a space-time block code.

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