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# Space–Time Line Code

**JINGON JOUNG, (Senior Member, IEEE)**
School of Electrical and Electronics Engineering, Chung-Ang University, Seoul 06974, South Korea
(jgjoung@cau.ac.kr)

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**ABSTRACT** This paper characterizes rate-one (i.e., full rate) full-spatial-diversity-achieving communication schemes based on the channel state information (CSI) availability and antenna configurations, i.e., CSI at a transmitter (CSIT) or CSI at a receiver (CSIR) and the numbers of transmit and receive antennas $M$ and $N$ (denoted by $M \times N$), respectively. The maximum ratio combining (MRC), maximum ratio transmission (MRT), and space–time block code (STBC) schemes are rate-one full-spatial-diversity-achieving method facilitated for communication systems with: 1) $1 \times N$ and CSIR; 2) $M \times 1$ and CSIT; and 3) $M \times 1$ and CSIR, respectively. A novel space–time line code (STLC) is then introduced for a $1 \times 2$ system with CSIT, and it is extended to an $M \times 2$ STLC. The proposed STLC uses CSI for encoding at the transmitter and enables the receiver to decode the STLC symbols without CSI. Also, the STLC encoding matrices with various code rates and decoding (combining) schemes are designed for the $M \times 3$ and $M \times 4$ STLC systems: A code rate of 3/4, 1/2, and 3/7 for the $M \times 3$ systems and a code rate of 3/4, 4/7, and 1/2 for the $M \times 4$ systems. For each STLC scheme, a full-diversity achieving STLC decoding method is designed. Based on analyses and numerical results, we verify that the proposed STLC scheme achieves a full diversity order, i.e., $MN$, and is robust against CSI uncertainty. It is also shown that the array processing gain is inversely proportional to the code rate. To verify the merit of STLC, we introduce a joint operation with STBC and STLC schemes, called an STBLC system. The STBLC system achieves full-spatial-diversity gain in both uplink and downlink communications. The new STLC achieving full-spatial diversity is scalable for various code rates and expected to be applied to various wireless communication systems along with MRC, MRT, and STBC.

**INDEX TERMS** Space–time code, space–time block code, space–time line code, spatial diversity gain, multiple antennas.

## I. INTRODUCTION

This study began as a fundamental attempt to look for a counter part to a well-known spatial diversity scheme known as the space-time block code (STBC). When a multiple-antenna transmitter does not have any channel state information (CSI), a space-time code (STC) such as the STBC technique, can provide full-spatial-diversity gain [1]–[4]. The full-spatial-diversity gain can be achieved through various techniques using multiple antennas depending on what is known about the CSI at the transmitter and/or receiver [5], [6]. A classical maximum-ratio-combining (MRC) technique achieves full diversity when, for example, a *receiver* with multiple antennas knows the CSI [7]–[11]. As a counterpart of the MRC, a maximum-ratio-transmission (MRT) technique was established that also achieves full-spatial-diversity gain when a *transmitter* with multiple antennas knows the CSI [12]–[14]. For clear and simple comparison, we assume $\alpha$-channel systems, in which $\alpha$-spatial channels are involved in communications, where $\alpha \geq 2$. A system with $M$-transmit and $N$-receive antennas has $\alpha = MN$ channels, and its configuration is denoted by $M \times N$. The existing $\alpha$(full)-spatial-diversity-achieving schemes can be categorized according to their system configurations into the following three schemes (see also Figs. 1(a)–(c), where $\alpha = 2$):

- MRC: $1 \times 2$ CSI is available at the receiver (Rx) only.
- MRT: $2 \times 1$ CSI is available at the transmitter (Tx) only.
- STBC: $2 \times 1$ CSI is available at the Rx only.

Here, one question arises: How do we achieve full-spatial diversity gain for a $1 \times 2$ multiple-input multiple-output (MIMO) system if CSI is available at the Tx only? This fundamental design question in full-spatial-diversity-achieving schemes motivates us to attempt to investigate a simple STC scheme with CSI at the transmitter as a counter part to STBC.

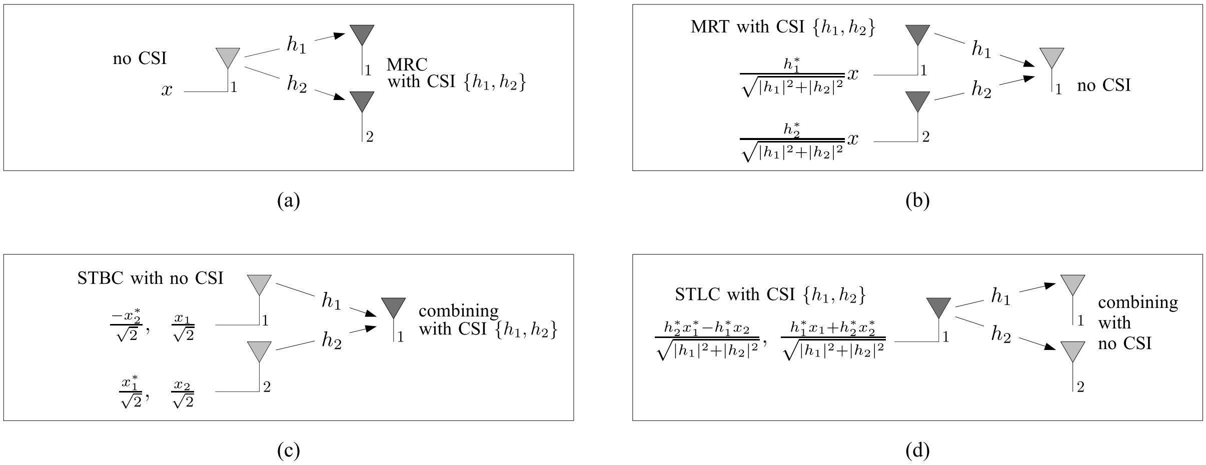


**FIGURE 1. Full-spatial-diversity-achieving schemes with two channels (i.e., $M = 2$). (a) MRC. (b) MRT. (c) STBC. (d) STLC.**

Consequently, we devise a new STC scheme referred to as space-time line code (STLC).

Precisely, in this study, we first devise a novel rate-one full-spatial-diversity-achieving STC scheme with CSI at the transmitter. In the proposed scheme, two information symbols are encoded using the multiple channel gains (*space*) and are transmitted consecutively (*time*). Given that the coded symbols are transmitted sequentially through a single transmit antenna, they are a *line*-shaped compared to the block shape of STBC, justifying the name of this new STC scheme as *space–time line code (STLC)*. The STLC can be directly extended to a system with multiple transmit antennas. In this case, multiple independent STLCs are implemented in parallel lines. Full diversity can be achieved by a simple decoding scheme, i.e., a direct combining scheme of the received signals at two receive antennas in two symbol times. An STLC receiver does not need full CSI for the decoding. Just the sum of all channel gain, which is a single real-value variable, is required; however, even this is not necessary for the detection of phase-shift keying (PSK) constellation signals. The full diversity of STLC is verified by a signal-to-noise ratio (SNR) analysis and by an uncoded bit error rate (BER) simulation. The new full-spatial-diversity-achieving STLC technique completes a missing part of the list above (see Fig. 1(d), where $\alpha = 2$):

- STLC: $1 \times 2$ CSI is available at the Tx only.

Further analytical and numerical results show that the proposed STLC is robust against CSI uncertainty as in STBC. Because the STLC is a full-spatial-diversity-achieving STC scheme, it has the same design properties, e.g., the orthogonality of encoding matrix and maximum-rate-achieving full-spatial-diversity gain, and advantages identical to those of the STBC, such as improved the error performance, data rates, capacities, and coverage areas.

Next, the STLC is further studied for a system with more than two receive antennas. The STLC encoding matrix $\boldsymbol{C}$ with code rate-$K/T$ is a $T$-by-$K$ complex-valued matrix, where $T$ is the number of symbol transmissions and $K$ is the length of information symbols. The designed STLC encoding matrix consists of the channel gains, and it fulfills the orthogonal property, i.e., $\boldsymbol{C}^H\boldsymbol{C}$ is a diagonal matrix. Herein, six STLC encoding matrices are designed. Three of them have code rates $3/4$, $1/2$, and $3/7$ and they are for a system with three receive antennas. The other three STLCs having code rates $3/4$, $4/7$, and $1/2$ are for a system with four receive antennas. For each of the designed STLC encoding scheme, an STLC decoding scheme that combines the received STLC signals is proposed in order to achieve full diversity order $\alpha$. This is verified by analytically showing that the instantaneous received SNR has a scale of $\gamma_\alpha \sigma_x^2/\sigma_z^2$, where $\gamma_\alpha$ is the sum of all channel gains involved in the communication and $\sigma_x^2/\sigma_z^2$ is the SNR of the single transmit and receive antenna system. Here, namely $\sigma_x^2$ is the transmitted symbol energy and $\sigma_z^2$ is the variance of additive white Gaussian noise (AWGN) at the receiver. Furthermore, the BER performance of each designed code is rigorously evaluated and compared to show the diversity gain and the achieved diversity order. From the results, it is generally observed that it is better to achieve the same data rate using more receive antennas.

Since the full CSI is required solely for a transmitter, the STLC scheme is a relevant strategy for communications between high-capable (complex) transmitters and minimal-function (simple) receivers. For example, internet-of-things (IoT) and wearable devices are the applications of the simple receivers, for which low cost, low complexity, and low power consumption are required [15]. The new full-spatial-diversity achieving structure, in which full CSI is available only at the complex device, relieves the simple device from frequent channel estimations and complex decoding and enables minimal-function operation by jointly operating the STLC and STBC schemes, which is called an STBLC. The proposed STLC scheme in this study is expected to be applied to various applications of the MIMO systems.

The rest of the paper is organized as follows. Section II briefly describes the three existing full-spatial-diversity-achieving schemes, i.e., MRC, MRC, and STBC.

In Section III, we propose STLC, the new full-spatial-diversity-achieving scheme, and present SNR analyses and BER simulation results in order to verify the performance of STLC. In Section IV, the STLC with various code rates is designed for a system with three and four receive antennas. Section V provides an application system, and Section VI concludes this paper.

*Notation:* Superscripts $T$, $H$ and $*$ denote transposition, Hermitian transposition, and complex conjugate, respectively, for any scalar, vector, or matrix. Further, E stands for the expectation of random variable $x$; for any scalar $x$, vector $\boldsymbol{x}$, and matrix $\boldsymbol{X}$, the notations $|x|$, $\|\boldsymbol{x}\|$, and $\|\boldsymbol{X}\|_F$ denote the absolute value of $x$, the 2-norm of $\boldsymbol{x}$, and the Frobenius-norm of $\boldsymbol{X}$, respectively; $\boldsymbol{I}_x$ represents an $x$-by-$x$ identity matrix; and $x \sim \mathcal{CN}(0, \sigma^2)$ means that a complex random variable $x$ conforms to a normal distribution with a zero mean and variance $\sigma^2$.

## II. EXISTING FULL-SPATIAL-DIVERSITY-ACHIEVING SCHEMES: MTC, MRT, AND STBC

We first briefly review the existing rate-one full-spatial-diversity-achieving schemes. Throughout the paper, a channel gain the $m$th transmit antenna to the $n$th receive antenna is denoted by $h_{(m-1)N+n}$, where subscripts $m = 1, \ldots, M$ and $n = 1, \ldots, N$ are indices for the transmit and receive antennas, respectively. Here, $MN$ spatial channels exist. Defining $\alpha = MN$, the sum of $\alpha$-spatial channel gains is then defined as follows:

$$\gamma_\alpha = \sum_{m=1}^{M}\sum_{n=1}^{N} |h_{(m-1)N+n}|^2. \quad (1)$$

For simplicity, we describe two-channel diversity systems with $\alpha = 2$, specifically the $1 \times 2$ MRC, $2 \times 1$ MRT, and $2 \times 1$ STBC systems, as depicted in Figs. 1(a), 1(b), and 1(c), respectively.

### A. MRC

Let $x$ be an information symbol that conforms to a complex normal distribution with $\mathrm{E}[|x|^2] = \sigma_x^2$, i.e., $\mathcal{CN}(0, \sigma_x^2)$. The two Rayleigh-fading channels from the transmit antenna to receive antenna $n$ are denoted by $h_n$, i.e., $h_1 \sim \mathcal{CN}(0, 1)$ and $h_2 \sim \mathcal{CN}(0, 1)$. The transmitter has no CSI, and thus, it transmits $x$ without precoding to the receiver. The received signal $r_{n,t}$ at the receive antenna $n$ at time $t$ is expressed as

$$r_{n,1} = h_n x + z_{n,1}, \quad n \in \{1, 2\}, \quad (2)$$

where, without a loss of generality (w.l.o.g.), we assume that the total transmit power is limited by the symbol power $\sigma_x^2$, and $z_{n,t}$ is AWGN at the $n$th receive antenna at time $t$ with zero mean and $\sigma_z^2$ variance; i.e., $z_{n,t} \sim \mathcal{CN}(0, \sigma_z^2)$.

Because the optimal receive combining (postprocessing) weights to maximize the receive SNR is $h_n^*/\sigma_z$ at the $n$th receive antenna [7], using the weights, the MRC output signal is derived from (2) as follows [7]:

$$\frac{h_1^*}{\sigma_z} r_{1,1} + \frac{h_2^*}{\sigma_z} r_{2,1} = \frac{\gamma_2}{\sigma_z} x + \frac{1}{\sigma_z}(h_1^* z_{1,1} + h_2^* z_{2,1}). \quad (3)$$

**TABLE 1. Encoding and transmit sequence for the STBC scheme.**

| | Tx antenna 1 | Tx antenna 2 |
|---|---|---|
| Tx time $t = 1$ | $x_1$ | $x_2$ |
| Tx time $t = 2$ | $-x_2^*$ | $x_1^*$ |

The combined signal in (3) is the input for a maximum likelihood (ML) detector achieving full-spatial-diversity gain, and the maximized received SNR is obtained by

$$\mathsf{SNR}_{\mathrm{MRC}}(\gamma_2) = \frac{\gamma_2 \sigma_x^2}{\sigma_z^2}. \quad (4)$$

From (4), it is clear that MRC obtains full-spatial-diversity gain, i.e., two. The transmit rate is obviously one.

### B. MRT

A transmitter has two antennas and CSI, while a receiver has a single antenna without CSI, as shown in Fig. 1(b). From the CSI, the transmitter can obtain the optimally weighted (precoded) symbol for transmit antenna $m$ that maximizes the received SNR, as follows [12]:

$$s_m = \frac{h_m^*}{\sqrt{\gamma_2}} x, \quad m \in \{1, 2\}, \quad (5)$$

where $h_m$ denotes the channel gain from transmit antenna $m$ to the receive antenna. In (5), the denominator follows the transmit power constraint, i.e., w.l.o.g., the total transmit power is limited by the symbol power $\sigma_x^2$ as an MRC system.

The weighed signals are transmitted to the receiver, simultaneously through two antennas, and the received signal is derived as follows:

$$r_{1,1} = h_1 s_1 + h_2 s_2 + z_{1,1} = \sqrt{\gamma_2} x + z_{1,1}. \quad (6)$$

From (6), we can readily derive the received SNR of MRT as

$$\mathsf{SNR}_{\mathrm{MRT}}(\gamma_2) = \frac{\gamma_2 \sigma_x^2}{\sigma_z^2}, \quad (7)$$

and clearly observe that the MRT achieves full-spatial-diversity gain identical to the maximum SNR of MRC in (4).

### C. STBC

Suppose a transmitter has two antennas without CSI, while a receiver has one antenna and CSI, as depicted in Fig. 1(c). Though the transmitter with multiple antennas has no CSI, the receiver can achieve full-spatial-diversity gain using STBC in Table 1 [2], where $x_k$ is the $k$th information symbol. Two consecutive STBC symbols are transmitted through two transmit antennas at two consecutive symbol times, $t = 1$ and $t = 2$. The receive signals are then written as follows:

$$r_{1,1} = h_1 \frac{x_1}{\sqrt{2}} + h_2 \frac{x_2}{\sqrt{2}} + z_{1,1},$$
$$r_{1,2} = -h_1 \frac{x_2^*}{\sqrt{2}} + h_2 \frac{x_1^*}{\sqrt{2}} + z_{1,2}, \quad (8)$$

respectively, where $1/\sqrt{2}$ is used for power constraint $\sigma_x^2$ at each transmission time. The receiver rearranges $r_{1,1}$ and $r_{1,2}$ and forms vector $\boldsymbol{r} = [r_{1,1}\ r_{1,2}^*]^T$, as expressed by

$$\boldsymbol{r} = \frac{1}{\sqrt{2}}\begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix}\begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} z_{1,1} \\ z_{1,2}^* \end{bmatrix} \triangleq \frac{1}{\sqrt{2}}\boldsymbol{H}_{(1,2)}\boldsymbol{x} + \boldsymbol{z}, \tag{9}$$

where $\boldsymbol{H}_{(1,2)}$ is the effective channel matrix of STBC and the subscript (1, 2) represents the channel indices associated with $h_1$ and $h_2$.

Noting that $\boldsymbol{H}_{(1,2)}$ satisfies an orthogonal property, i.e., $\boldsymbol{H}_{(1,2)}^H\boldsymbol{H}_{(1,2)} = \boldsymbol{I}_2$, for optimal decoding or combining, the receiver multiplies $\boldsymbol{H}_{(1,2)}^H$ by $\boldsymbol{r}$. The decoded signals are then derived as follows:

$$\boldsymbol{H}_{(1,2)}^H\boldsymbol{r} = \frac{\gamma_2}{\sqrt{2}}\boldsymbol{I}_2\boldsymbol{x} + \boldsymbol{z}', \tag{10}$$

where $\boldsymbol{z}' = \boldsymbol{H}_{(1,2)}^H\boldsymbol{z}$ is a complex Gaussian noise vector with a zero mean and covariance matrix $\mathrm{E}[\boldsymbol{z}'(\boldsymbol{z}')^H] = \gamma_2\sigma_z^2\boldsymbol{I}_2$. From (10), we note that $x_1$ and $x_2$ can be estimated separately by an ML detector, with the received SNR derived as

$$\mathsf{SNR}_{\mathrm{STBC}}(\gamma_2) = \frac{\gamma_2\sigma_x^2}{2\sigma_z^2}. \tag{11}$$

From (11), we also observe that the STBC achieves full-spatial-diversity gain, i.e., two. Note that there is no array processing gain (i.e., the array gain is one) because $x_k$ is transmitted using half of the total symbol energy $\sigma_x^2$. In general, $M$ array processing gain is achieved by a 2-by-$M$ STBC scheme.

## III. NEW FULL-SPATIAL-DIVERSITY-ACHIEVING SCHEME: STLC

### A. SINGLE TRANSMIT ANTENNA AND TWO RECEIVE ANTENNAS

Consider a system with one transmit and two receive antennas, as depicted in Figs. 1(d). Channel gains $h_1$ and $h_2$ represent the independent channel gains from the transmit antenna to the receive antennas 1 and 2, respectively. The transmitter has CSI, yet the receiver does not.

#### 1) ENCODING AND TRANSMISSION SEQUENCE

Denote the STLC symbol transmitted at time $t$ by $s_t$. Two information symbols $x_1$ and $x_2$ are encoded to two STLC symbols $s_1$ and $s_2$, as follows:

$$\begin{bmatrix} s_1^* \\ s_2 \end{bmatrix} = \boldsymbol{C}_{(1,2)}\begin{bmatrix} x_1^* \\ x_2 \end{bmatrix} = \begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix}\begin{bmatrix} x_1^* \\ x_2 \end{bmatrix}, \tag{12}$$

where $\boldsymbol{C}_{(1,2)}$ is referred to as an STLC encoding matrix with channels $h_1$ and $h_2$. In this example, the STLC encoding matrix $\boldsymbol{C}_{(1,2)}$ is designed to be identical to STBC effective channel matrix $\boldsymbol{H}_{(1,2)}$ in (9). The two STLC symbols $s_1$ and $s_2$ are consecutively transmitted during the first and second symbol periods; they are expressed from (12) as follows (Table 2):

$$s_1 = h_1^*x_1 + h_2^*x_2^*, \tag{13a}$$

$$s_2 = h_2^*x_1^* - h_1^*x_2. \tag{13b}$$

**TABLE 2. Encoding and transmit sequence for the STLC scheme with one transmit antenna.**

| | Tx time $t=1$ | Tx time $t=2$ |
|---|---|---|
| Tx antenna 1 | $s_1 = h_1^*x_1 + h_2^*x_2^*$ | $s_2 = h_2^*x_1^* - h_1^*x_2$ |

**TABLE 3. Notations for the STLC received signals.**

| | Rx time $t=1$ | Rx time $t=2$ |
|---|---|---|
| Rx antenna 1 | $r_{1,1}$ | $r_{1,2}$ |
| Rx antenna 2 | $r_{2,1}$ | $r_{2,2}$ |

To satisfy the transmit power constraint $\sigma_x^2$, the transmitter normalizes $s_1$ and $s_2$ with $\eta$ and transmits them consecutively. The normalization factor $\eta$ can be readily obtained as $\eta = 1/\sqrt{\gamma_2}$ such that $\mathrm{E}[|\eta s_t|^2] = \sigma_x^2$.

The four received symbols are then written as follows (see Table 3):

$$\begin{bmatrix} r_{1,1} & r_{1,2} \\ r_{2,1} & r_{2,2} \end{bmatrix} = \begin{bmatrix} h_1 \\ h_2 \end{bmatrix}\frac{1}{\sqrt{\gamma_2}}\underbrace{\begin{bmatrix} s_1 & s_2 \end{bmatrix}}_{STLC} + \begin{bmatrix} z_{1,1} & z_{1,2} \\ z_{2,1} & z_{2,2} \end{bmatrix}. \tag{14}$$

Substituting $s_t$ in (13a) into (14), each of the received signals, $r_{n,t}$, is written as follows:

$$r_{1,1} = \frac{1}{\sqrt{\gamma_2}}h_1(h_1^*x_1 + h_2^*x_2^*) + z_{1,1}, \tag{15a}$$

$$r_{1,2} = \frac{1}{\sqrt{\gamma_2}}h_1(h_2^*x_1^* - h_1^*x_2) + z_{1,2}, \tag{15b}$$

$$r_{2,1} = \frac{1}{\sqrt{\gamma_2}}h_2(h_1^*x_1 + h_2^*x_2^*) + z_{2,1}, \tag{15c}$$

$$r_{2,2} = \frac{1}{\sqrt{\gamma_2}}h_2(h_2^*x_1^* - h_1^*x_2) + z_{2,2}. \tag{15d}$$

#### 2) DECODING SCHEME

The four received symbols in (15a) are combined to decode the STLC symbols so that full-spatial diversity is achieved as shown in Fig. 2. Directly combining $\{r_{n,t}\}$ in (15a), the receiver can decode the STLC symbols as follows:

$$r_{1,1} + r_{2,2}^* = \sqrt{\gamma_2}x_1 + z_{1,1} + z_{2,2}^* \tag{16a}$$

$$r_{2,1}^* - r_{1,2} = \sqrt{\gamma_2}x_2 + z_{2,1}^* - z_{1,2}. \tag{16b}$$

Note that (16a) is only a function of $x_1$ and (16b) is only a function of $x_2$. Thus, two separate ML detections of $x_1$ and $x_2$ are possible, as in an STBC decoder. The effective channel gain $\sqrt{\gamma_2}$ is needed during the subsequent ML detection process. In contrast to MRC in (3), the STLC receiver does not need full CSI. Instead, to combine the received signals in (16a), only the effective channel gain is required and it can be estimated by using the blind SNR estimation techniques (see [16] and references therein). Thus, (16a) is called a *blind combining* technique. Note that even the partial CSI $\sqrt{\gamma_2}$ is not required for PSK symbol detection.

An STBC decoding scheme which requires six operations (four multiplications and two additions) of the complex values. On top of this, the STBC receiver needs to estimate the

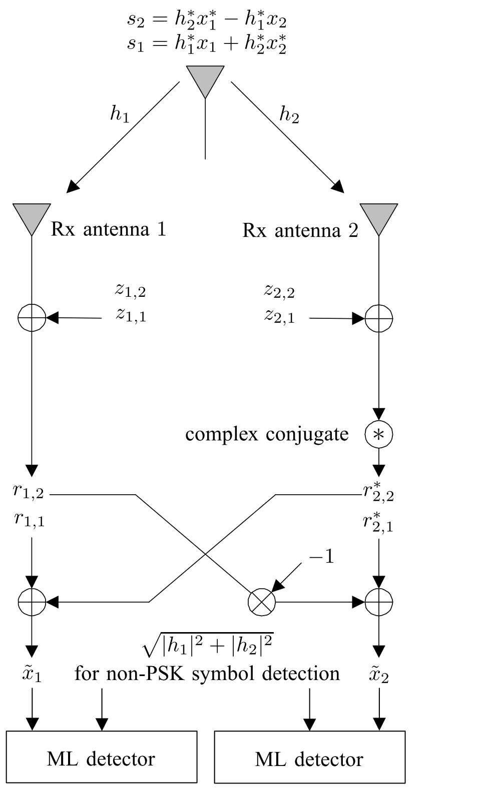


**FIGURE 2.** Example of a new full-spatial-diversity-achieving 1 × 2 STLC system.

**TABLE 4.** Rate-One STLC encoding and decoding structures for a system with two receive antennas.

| | STLC encoding matrices | Decoding function: $f(a,b)=a+b$ for $\tilde{x}_1$ | for $\tilde{x}_2$ |
|---|---|---|---|
| $\boldsymbol{C}_{(1,2)}$ | $\boldsymbol{C}^a_{(1,2)} = \begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix}$ | $f(r^*_{1,1}, r_{2,2})$ | $f(r^*_{2,1}, -r_{1,2})$ |
| | $\boldsymbol{C}^b_{(1,2)} = \begin{bmatrix} h_1 & h_2 \\ -h_2^* & h_1^* \end{bmatrix}$ | $f(r^*_{1,1}, -r_{2,2})$ | $f(r^*_{2,1}, r_{1,2})$ |
| | $\boldsymbol{C}^c_{(1,2)} = \begin{bmatrix} h_1 & -h_2 \\ h_2^* & h_1^* \end{bmatrix}$ | $f(r^*_{1,1}, r_{2,2})$ | $f(-r^*_{2,1}, r_{1,2})$ |
| | $\boldsymbol{C}^d_{(1,2)} = \begin{bmatrix} -h_1 & h_2 \\ h_2^* & h_1^* \end{bmatrix}$ | $f(-r^*_{1,1}, r_{2,2})$ | $f(r^*_{2,1}, r_{1,2})$ |
| Type | Encoding $s_1$ and $s_2$ | Decoding for $\tilde{x}_1$ | Decoding for $\tilde{x}_2$ |
| 1 | $\begin{bmatrix} s_1^* \\ s_2 \end{bmatrix} = \boldsymbol{C}_{(1,2)} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix}$ | $f(\cdot,\cdot)$ | $f(\cdot,\cdot)$ |
| 2 | $\begin{bmatrix} s_1^* \\ s_2 \end{bmatrix} = \boldsymbol{C}_{(1,2)} \begin{bmatrix} x_1^* \\ x_2 \end{bmatrix}$ | $f^*(\cdot,\cdot)$ | $f(\cdot,\cdot)$ |
| 3 | $\begin{bmatrix} s_1^* \\ s_2 \end{bmatrix} = \boldsymbol{C}_{(1,2)} \begin{bmatrix} x_1 \\ x_2^* \end{bmatrix}$ | $f(\cdot,\cdot)$ | $f^*(\cdot,\cdot)$ |
| 4 | $\begin{bmatrix} s_1^* \\ s_2 \end{bmatrix} = \boldsymbol{C}_{(1,2)} \begin{bmatrix} x_1^* \\ x_2^* \end{bmatrix}$ | $f^*(\cdot,\cdot)$ | $f^*(\cdot,\cdot)$ |
| 5 | $\begin{bmatrix} s_1 \\ s_2^* \end{bmatrix} = \boldsymbol{C}^*_{(1,2)} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix}$ | $f^*(\cdot,\cdot)$ | $f^*(\cdot,\cdot)$ |
| 6 | $\begin{bmatrix} s_1 \\ s_2^* \end{bmatrix} = \boldsymbol{C}^*_{(1,2)} \begin{bmatrix} x_1^* \\ x_2 \end{bmatrix}$ | $f(\cdot,\cdot)$ | $f^*(\cdot,\cdot)$ |
| 7 | $\begin{bmatrix} s_1 \\ s_2^* \end{bmatrix} = \boldsymbol{C}^*_{(1,2)} \begin{bmatrix} x_1 \\ x_2^* \end{bmatrix}$ | $f^*(\cdot,\cdot)$ | $f(\cdot,\cdot)$ |
| 8 | $\begin{bmatrix} s_1 \\ s_2^* \end{bmatrix} = \boldsymbol{C}^*_{(1,2)} \begin{bmatrix} x_1^* \\ x_2^* \end{bmatrix}$ | $f(\cdot,\cdot)$ | $f(\cdot,\cdot)$ |
| 9 | $\begin{bmatrix} s_1^* \\ s_2 \end{bmatrix} = -\boldsymbol{C}_{(1,2)} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix}$ | $-f(\cdot,\cdot)$ | $-f(\cdot,\cdot)$ |
| 10 | $\begin{bmatrix} s_1^* \\ s_2 \end{bmatrix} = -\boldsymbol{C}_{(1,2)} \begin{bmatrix} x_1^* \\ x_2 \end{bmatrix}$ | $-f^*(\cdot,\cdot)$ | $-f(\cdot,\cdot)$ |
| 11 | $\begin{bmatrix} s_1^* \\ s_2 \end{bmatrix} = -\boldsymbol{C}_{(1,2)} \begin{bmatrix} x_1 \\ x_2^* \end{bmatrix}$ | $-f(\cdot,\cdot)$ | $-f^*(\cdot,\cdot)$ |
| 12 | $\begin{bmatrix} s_1^* \\ s_2 \end{bmatrix} = -\boldsymbol{C}_{(1,2)} \begin{bmatrix} x_1^* \\ x_2^* \end{bmatrix}$ | $-f^*(\cdot,\cdot)$ | $-f^*(\cdot,\cdot)$ |
| 13 | $\begin{bmatrix} s_1 \\ s_2^* \end{bmatrix} = -\boldsymbol{C}^*_{(1,2)} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix}$ | $-f^*(\cdot,\cdot)$ | $-f(\cdot,\cdot)^*$ |
| 14 | $\begin{bmatrix} s_1 \\ s_2^* \end{bmatrix} = -\boldsymbol{C}^*_{(1,2)} \begin{bmatrix} x_1^* \\ x_2 \end{bmatrix}$ | $-f(\cdot,\cdot)$ | $-f^*(\cdot,\cdot)$ |
| 15 | $\begin{bmatrix} s_1 \\ s_2^* \end{bmatrix} = -\boldsymbol{C}^*_{(1,2)} \begin{bmatrix} x_1 \\ x_2^* \end{bmatrix}$ | $-f^*(\cdot,\cdot)$ | $-f(\cdot,\cdot)$ |
| 16 | $\begin{bmatrix} s_1 \\ s_2^* \end{bmatrix} = -\boldsymbol{C}^*_{(1,2)} \begin{bmatrix} x_1^* \\ x_2^* \end{bmatrix}$ | $-f(\cdot,\cdot)$ | $-f(\cdot,\cdot)$ |

channels which requires two operations if a simple linear estimator is employed. On the other hand, since the proposed STLC decoder performs only two additions of complex values for the decoding, the decoding complexity per symbol is reduced roughly by 75%.

#### 3) RECEIVED SNR OF STLC

Because the sum of two independent AWGNs is also an AWGN, $(z_{1,1} + z^*_{2,2})$ and $(z^*_{2,1} - z_{1,2})$ in (16a) are AWGN with a zero mean and $2\sigma_z^2$ variance; specifically, $(z^*_{1,1} + z_{2,2}) \sim \mathcal{CN}(0, 2\sigma_z^2)$ and $(-z_{2,1} + z^*_{1,2}) \sim \mathcal{CN}(0, 2\sigma_z^2)$. Thus, the resulting instantaneous received SNR after the blind combining is readily derived from (16a) as

$$\mathsf{SNR}_{\mathsf{STLC}}(\gamma_2) = \frac{\gamma_\alpha \sigma_x^2}{2\sigma_z^2} = \frac{\gamma_2 \sigma_x^2}{2\sigma_z^2}, \tag{17}$$

which is identical to the SNR of STBC in (11). From (17), we verify that STLC achieves performance identical to that of STBC in terms of the diversity gain and array processing gain. The factor of two stems from the fact that the two AWGNs are directly combined at the receiver, as shown in (16a).

#### 4) ENCODING AND COMBINING STRUCTURE

The encoding and decoding structures in (12) and (16a) are not unique. Examples of possible STLC encoding and decoding structures are listed in Table 4. In (12) and (16a), we use an STLC matrix $\boldsymbol{C}^a_{(1,2)}$ and Type-2 STLC encoding-and-decoding structure. Here, $f(\cdot,\cdot)$ is a decoding function defined as $f(a,b) = a + b$. Note that all the STLC encoding matrices in Table 4 fulfill an orthogonal property, i.e., $\boldsymbol{C}^H\boldsymbol{C} = \gamma_2 \boldsymbol{I}_2$, and their rank is two, which provides the full-spatial-diversity order two [1], [17]. It is apparent that the decoding scheme is determined according to the encoding scheme. All of the encoding-decoding pairs show identical performance capabilities, specifically, rate-one and full-spatial-diversity gain; however, the requirements for implementation may not be equal. For example, a type-2 STLC scheme requires conjugate operation at only one radio frequency (RF) chain (e.g., Rx antenna 2 in Fig. 2), while a Type-1 STLC scheme requires conjugate operations at both RF chains. Hence, we can provide a design policy of the STLC structure depending on the receiver capability.

### B. MULTIPLE TRANSMIT ANTENNAS AND TWO RECEIVE ANTENNAS

The STLC encoding proposed in Section III-A can be generalized to a system with $M$ multiple transmit antennas, where $M \geq 2$. The STLC encoding for $M$ transmit antennas can be written as follows:

$$\left[s^*_{1,1}\ s_{1,2} \cdots s^*_{m,1}\ s_{m,2}\ \cdots s^*_{M,1}\ s_{M,2}\right]^T = \boldsymbol{C}_{(1:2M)} \begin{bmatrix} x_1^* \\ x_2 \end{bmatrix}, \tag{18}$$

**TABLE 5.** Definitions of channels between the transmit and receive antennas.

| | Tx antenna 1 | Tx antenna 2 |
|---|---|---|
| Rx antenna 1 | $h_1$ | $h_3$ |
| Rx antenna 2 | $h_2$ | $h_4$ |

**TABLE 6.** Encoding and transmit sequence for the STLC scheme with two transmit antennas.

| | Tx time $t = 1$ | Tx time $t = 2$ |
|---|---|---|
| Tx antenna 1 | $s_{1,1} = h_1^* x_1 + h_2^* x_2^*$ | $s_{1,2} = h_2^* x_1^* - h_1^* x_2$ |
| Tx antenna 2 | $s_{2,1} = h_3^* x_1 + h_4^* x_2^*$ | $s_{2,2} = h_4^* x_1^* - h_3^* x_2$ |

where $s_{m,t}$ is the transmitted symbol through the $m$th transmit antenna at time $t$, and the STLC encoding matrix $\boldsymbol{C}_{(1:2M)} \in \mathbb{C}^{2M\times 2}$ is constructed as

$$\boldsymbol{C}_{(1:2M)} = \left[\boldsymbol{C}_{(1,2)}^T \cdots \boldsymbol{C}_{(2m-1,2m)}^T \cdots \boldsymbol{C}_{(2M-1,2M)}^T\right]^T. \tag{19}$$

Here, $\boldsymbol{C}_{(2m-1,2m)}$ is an STLC encoding matrix that consists of $h_{2m-1}$ and $h_{2m}$. In (19), generally, $\boldsymbol{C}_{(2m-1,2m)}$ can conform to any STLC structure in Table 4. For this reason, $\boldsymbol{C}_{(1:2M)}$ fulfills an orthogonality property as

$$\boldsymbol{C}_{(1:2M)}^H \boldsymbol{C}_{(1:2M)} = \sum_{m=1}^{M} \boldsymbol{C}_{(2m-1,2m)}^H \boldsymbol{C}_{(2m-1,2m)} = \gamma_{2M} \boldsymbol{I}_2.$$

The decoding structure depends on the STLC encoding structure, as shown in Table 4, similar to the case of a single receive antenna system. The details are introduced with an example of a 2 × 2 STLC system, as depicted in Fig. 3. The notations for the channel gains for a 2 × 2 STLC are shown in Table 5. Channels $h_3$ and $h_4$ represent independent channel gains from transmit antenna 2 to receive antennas 1 and 2, respectively.

### 1) ENCODING AND TRANSMISSION SEQUENCE

For STLC with two transmit antennas, w.l.o.g., we apply an STLC encoding matrix $\boldsymbol{C}_{(1,2)}^a$ and a Type-2 structure in Table 4 to each transmit antenna, i.e., $\boldsymbol{C}_{(2m-1,2m)} = \boldsymbol{C}_{(2m-1,2m)}^a$ for all $m \in \{1, 2\}$, with decoding functions $f^*(\cdot,\cdot)$ and $f(\cdot,\cdot)$ are used for estimating $x_1$ and $x_2$, respectively. Thus, the encoding with $x_1$ and $x_2$ is written as follows:

$$\begin{bmatrix} s_{1,1}^* \\ s_{1,2} \\ s_{2,1}^* \\ s_{2,2} \end{bmatrix} = \begin{bmatrix} \boldsymbol{C}_{(1,2)}^a \\ \boldsymbol{C}_{(3,4)}^a \end{bmatrix} \begin{bmatrix} x_1^* \\ x_2 \end{bmatrix} = \begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \\ h_3 & h_4 \\ h_4^* & -h_3^* \end{bmatrix} \begin{bmatrix} x_1^* \\ x_2 \end{bmatrix}. \tag{20}$$

The resultant STLC symbols are shown in Table 6.

To satisfy transmit power constraint $\sigma_x^2$, the transmitter normalizes $s_{1,t}$ and $s_{2,t}$, according to $\eta$. The normalization factor $\eta$ can be readily derived as $\eta = 1/\sqrt{\gamma_4}$ such that $\mathrm{E}\,|\eta s_{1,t}|^2 + \mathrm{E}\,|\eta s_{2,t}|^2 = \sigma_x^2$ for all $t$. The transmitter then transmits $\eta s_{1,t}$ and $\eta s_{2,t}$ through transmit antennas 1 and 2, respectively, simultaneously at time $t$. Concurrently,

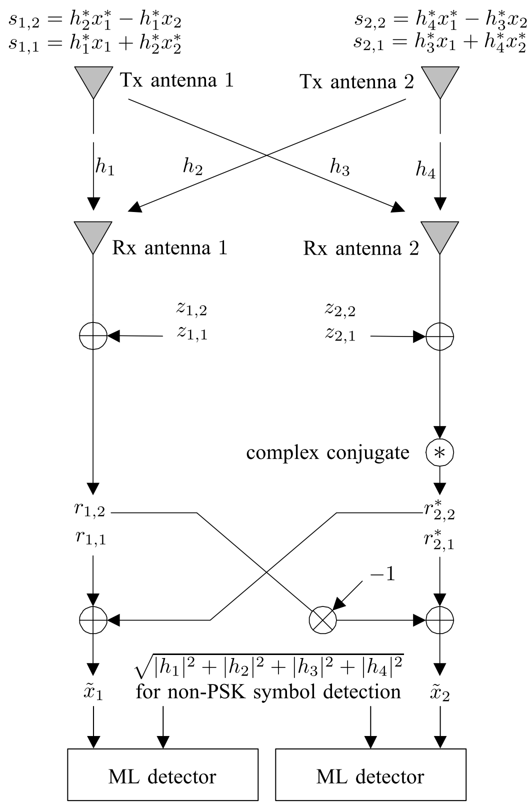


**FIGURE 3.** Example of a new full-spatial-diversity-achieving 2 × 2 STLC system.

the receive symbols defined in Table 3 can be expressed as follows:

$$\begin{bmatrix} r_{1,1} & r_{1,2} \\ r_{2,1} & r_{2,2} \end{bmatrix} = \begin{bmatrix} h_1 & h_3 \\ h_2 & h_4 \end{bmatrix} \frac{1}{\sqrt{\gamma_4}} \begin{bmatrix} \underbrace{s_{1,1}\ s_{1,2}}_{STLC1} \\ \underbrace{s_{2,1}\ s_{2,2}}_{STLC2} \end{bmatrix} + \begin{bmatrix} z_{1,1} & z_{1,2} \\ z_{2,1} & z_{2,2} \end{bmatrix}. \tag{21}$$

### 2) DECODING SCHEME

Because an STLC decoding structure follows the encoding structure, specifically STLC encoding matrix $\boldsymbol{C}_{(1,2)}^a$ with a Type-2 structure, from Table 4, the decoding scheme is readily determined as follows:

$$f^*(r_{1,1}^*, r_{2,2}) = r_{1,1} + r_{2,2}^* = \sqrt{\gamma_4} x_1 + z_{1,1} + z_{2,2}^*, \tag{22a}$$
$$f(r_{2,1}^*, -r_{1,2}) = r_{2,1}^* - r_{1,2} = \sqrt{\gamma_4} x_2 + z_{2,1}^* - z_{1,2}. \tag{22b}$$

The combiner is shown in Fig. 3. Here, we reemphasize that the receiver does not need full CSI to combine the received signals, yet it requires the effective channel gain $\sqrt{\gamma_4}$ for the ML detection of non-PSK symbols in the sequel.

### 3) RECEIVED SNR OF STLC

The resulting SNR after blind combining is readily derived from (22a) as follows:

$$\mathrm{SNR}_{\mathrm{STLC}}(\gamma_4) = \frac{\gamma_4 \sigma_x^2}{2\sigma_z^2}. \tag{23}$$

From (23), we verify that STLC with four channels undoubtedly achieves a diversity order of four (full-spatial-diversity gain) and an array processing gain of two. Directly extending the STLC encoding matrix in (20) for $M$ transmit antennas, and following (21) and the decoding scheme in (22a), we can derive the received SNR of a system with $M$ transmit and two receive antennas as follows:

$$\mathrm{SNR}_{\mathrm{STLC}}(\gamma_{2M}) = \frac{\gamma_\alpha \sigma_x^2}{2\sigma_z^2} = \frac{\gamma_{2M}\sigma_x^2}{2\sigma_z^2}. \tag{24}$$

From (24), we find that an $M \times 2$ STLC system can achieve a diversity order of $2M$ and array gain of $M$, identical to the diversity order and array gain of a $2 \times M$ STBC system.

The multiple transmit antenna STLC scheme designed in this section can be applied to a massive MIMO system. In [18], it was shown that the $M$-by-2 STLC system asymptotically achieves optimal (maximum) SNR as $M$ increases, and it achieves stable SNR, regardless of the spatial correlation, and considerable robustness against channel uncertainty at the transmitter. Furthermore, the massive MIMO STLC was applied to a system that supports multiple users (i.e., receivers). In this system, the transmit antennas are allocated to each user in order to improve the average signal-to-interference-plus-noise ratio (SINR). In the STLC antenna allocation system, each user achieves a full-spatial diversity order from the allocated transmit antennas. The STLC was also exploited to improve secure MIMO communications [19].

### C. PERFORMANCE VERIFICATION OF STLC

We present the Monte Carlo simulation results of the BER performance for the four rate-one full-spatial-diversity-achieving schemes, specifically MRC, MRT, STBC, and STLC. For comparison purposes, no diversity scheme having single transmit and receive antennas is included. Channels are of the Rayleigh-fading type, i.e., $h \sim \mathcal{CN}(0, 1)$. Uncoded coherent binary PSK (BPSK) is considered in the simulation.

In Fig. 4, two spatial channels, i.e., $\alpha = 2$, are considered: $1 \times 2$ MRC, $2 \times 1$ MRT, $2 \times 1$ STBC, and $1 \times 2$ STLC. In Fig. 5, four spatial channels, i.e., $\alpha = 4$, are considered: $1\times4$ MRC, $4\times1$ MRT, $2\times2$ STBC, and $2\times2$ STLC. As shown in Figs. 4 and 5, the BER results demonstrate clearly that the new diversity scheme STLC achieves performance identical to that of the STBC, and that all schemes achieve full-spatial-diversity gain. There is degradation of 3 dB SNR for STBC and STLC compared to MRC and MRT. As stated in Section III, the reason for the 3-dB penalty in the STBC case is the half-power transmission on each antenna, while that of STLC is caused by the direct combining of received signals, which directly combines the noise factors.

### D. ROBUSTNESS AGAINST CSI UNCERTAINTY

The CSI at the STLC transmitter could be outdated in practice, as the CSI is obtained from channel estimation in the previous receiving mode; i.e., a time-varying channel in a time-division duplex (TDD) system. In addition, considering the estimation error and the mismatch of channel calibration

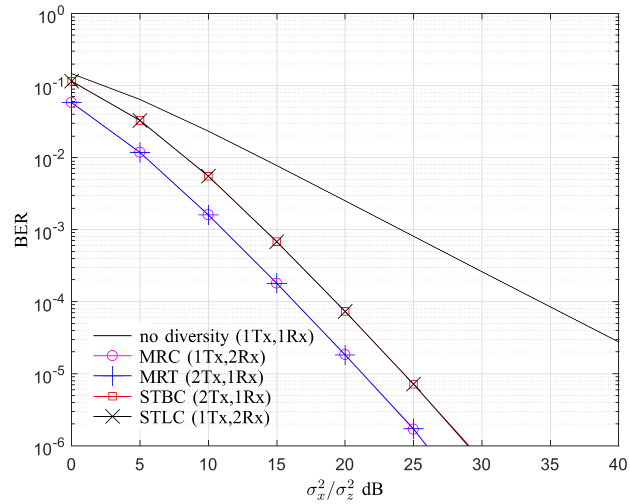


**FIGURE 4.** BER performance comparison of coherent BPSK with $1 \times 2$ MRC, $2 \times 1$ MRT, $2 \times 1$ STBC, and $1 \times 2$ STLC during Rayleigh fading.

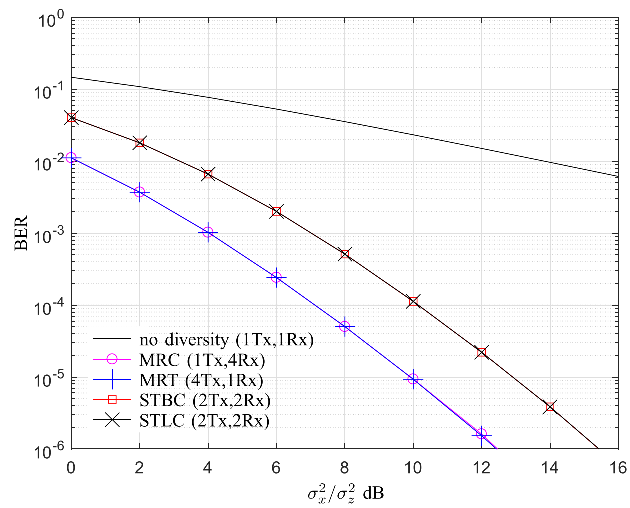


**FIGURE 5.** BER performance comparison of coherent BPSK with $1 \times 4$ MRC, $4 \times 1$ MRT, $2 \times 2$ STBC, and $2 \times 2$ STLC during Rayleigh fading.

in the TDD systems, CSI uncertainty from the estimation error is inevitable.

Denote an estimated CSI as $\tilde{h} \triangleq h + \epsilon$, where $h$ is the actual channel and $\epsilon$ is the estimation error. Assuming that the estimation errors of all channels are independent of one another and that they conform to a normal distribution with a mean of zero and variance of $\sigma_\epsilon^2$, we represent the mean-squared error (MSE) of the estimation by $\sigma_\epsilon^2$, i.e., $\mathrm{E}\,|h - \tilde{h}|^2 = \sigma_\epsilon^2$. Under CSI uncertainty, the SNRs of STLC and STBC are now analyzed and compared. As a result, it is analytically and numerically shown that STBC and STLC provide identical performance levels even when the CSI contains uncertainty.

#### 1) SNR ANALYSIS OF $1 \times 2$ STLC UNDER CSI UNCERTAINTY ($\alpha = 2$)

Applying the uncertain CSI model, i.e., $\tilde{h}_n = h_n + \epsilon_n$ and $\epsilon_n \sim \mathcal{CN}(0, \sigma_\epsilon^2)$, to the encoding of the STLC symbols in (13a), the received signals in (15a) are rewritten as follows:

$$r_{1,1} = \frac{1}{\sqrt{\tilde{\gamma}_2}} h_1(\tilde{h}_1^* x_1 + \tilde{h}_2^* x_2^*) + z_{1,1}, \quad (25a)$$

$$r_{1,2} = \frac{1}{\sqrt{\tilde{\gamma}_2}} h_1(\tilde{h}_2^* x_1^* - \tilde{h}_1^* x_2) + z_{1,2}, \quad (25b)$$

$$r_{2,1} = \frac{1}{\sqrt{\tilde{\gamma}_2}} h_2(\tilde{h}_1^* x_1 + \tilde{h}_2^* x_2^*) + z_{2,1}, \quad (25c)$$

$$r_{2,2} = \frac{1}{\sqrt{\tilde{\gamma}_2}} h_2(\tilde{h}_2^* x_1^* - \tilde{h}_1^* x_2) + z_{2,2}, \quad (25d)$$

where $\tilde{\gamma}_2 = |\tilde{h}_1|^2 + |\tilde{h}_2|^2$. Using (16a), the receiver decodes STLC from (25a) as

$$r_{1,1} + r_{2,2}^* = \frac{\gamma_2}{\sqrt{\tilde{\gamma}_2}} x_1 + \frac{(h_1\epsilon_1^* - h_2^*\epsilon_2)\, x_1}{\sqrt{\tilde{\gamma}_2}} + \frac{(h_1\epsilon_2^* + h_2^*\epsilon_1)\, x_2^*}{\sqrt{\tilde{\gamma}_2}} + z_{1,1} - z_{2,2}^*, \quad (26a)$$

$$r_{2,1}^* - r_{1,2}^* = \frac{\gamma_2}{\sqrt{\tilde{\gamma}_2}} x_2 + \frac{(h_1\epsilon_1^* + h_2^*\epsilon_2)\, x_2}{\sqrt{\tilde{\gamma}_2}} + \frac{(h_2^*\epsilon_1 - h_1\epsilon_2^*)\, x_1^*}{\sqrt{\tilde{\gamma}_2}} + z_{2,1}^* - z_{1,2}. \quad (26b)$$

Assuming that the receiver knows the effective channel gain (not full CSI) $\gamma_2/\sqrt{\tilde{\gamma}_2}$ for equalization and that the interference terms, i.e., the second and third terms on the right-hand side of (26a) and (26b), are AWGN, the SNR of $1 \times 2$ STLC under CSI uncertainty is derived from (26a) as follows:

$$\mathsf{SNR}_{\mathsf{STLC}}(\gamma_2, \sigma_\epsilon^2) = \frac{\gamma_2^2 \sigma_x^2}{2(\gamma_2 \sigma_x^2 \sigma_\epsilon^2 + (\gamma_2 + 2\sigma_\epsilon^2)\sigma_z^2)}. \quad (27)$$

In (27), we also assumed that the estimation error $\epsilon_n$, data symbol $x_n$, and noise $z_n$ are independent of one another (the proof is tedious and is omitted in this paper). As expected, when there is no channel uncertainty, i.e., $\sigma_\epsilon^2 = 0$, (27) is reduced to (17).

Similarly, under the CSI uncertainty, the decoded symbols of STBC in (10) are written as $\tilde{\boldsymbol{H}}_{(1,2)}^H(\boldsymbol{H}_{(1,2)}\boldsymbol{x}+\boldsymbol{z})$. Hence, the estimates of $x_1$ and $x_2$ of STBC can be derived as follows:

$$\tilde{x}_1 = x_1 + \frac{(h_1\epsilon_1^* + h_2^*\epsilon_2)\, x_1}{\gamma_2} + \frac{(h_2\epsilon_1^* - h_1^*\epsilon_2)\, x_2}{\gamma_2} + \frac{\sqrt{2}\left((h_1^* + \epsilon_1^*)z_{1,1} + (h_2 + \epsilon_2)z_{1,2}^*\right)}{\gamma_2}, \quad (28a)$$

$$\tilde{x}_2 = x_2 + \frac{(h_1^*\epsilon_1 + h_2\epsilon_2^*)\, x_2}{\gamma_2} + \frac{(h_1\epsilon_2^* - h_2^*\epsilon_1)\, x_1}{\gamma_2} + \frac{\sqrt{2}\left((h_2^* + \epsilon_2^*)z_{1,1} - (h_1 + \epsilon_1)z_{1,2}^*\right)}{\gamma_2}, \quad (28b)$$

where the effective channel gain $\gamma_2$ is assumed to be known at the STBC receiver. From (28a), we can derive the SNR of a $2 \times 1$ STBC system and show that it is identical to that of a $1 \times 2$ STLC system, i.e.,

$$\mathsf{SNR}_{\mathsf{STBC}}(\gamma_2, \sigma_\epsilon^2) = \mathsf{SNR}_{\mathsf{STLC}}(\gamma_2, \sigma_\epsilon^2). \quad (29)$$

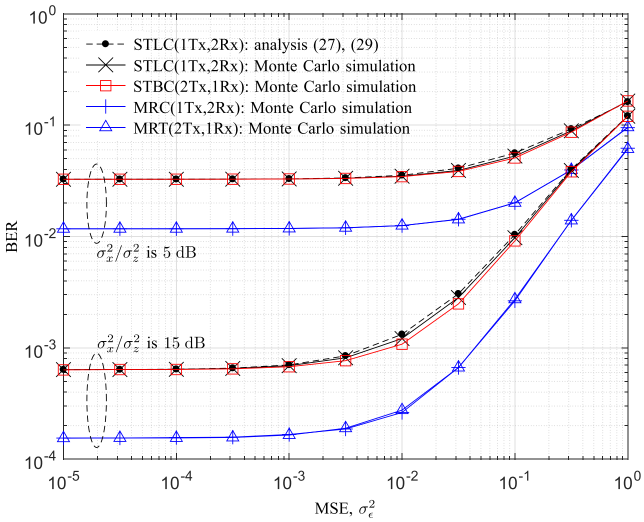


**FIGURE 6. BER performance over the MSE of CSI estimation for two-channel systems.**

The analysis in (29) shows that the SNRs of two-channel STBC and STLC are identical even under an uncertain CSI environment, which is also numerically verified in Fig. 6, where BER with BPSK modulation is evaluated for a different channel uncertainty of $\sigma_\epsilon^2$, i.e., the MSE of channel estimation. The analytical BER performance is determined as BER $= Q(\sqrt{2\mathsf{SNR}})$ for BPSK [20], where the SNR is the analytic SNRs in (27) and (29) and $Q(\cdot)$ is a Q-function, i.e., $Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty \exp\left(\frac{-u^2}{2}\right) du$. The analyses in (27) and (29) are in good agreement with the numerical results. It is observed that STLC and STBC are tolerant against CSI uncertainty up to approximately $\sigma_\epsilon^2 = 10^{-3}$. For comparison purposes, we include the BER performances of MRT and MRC when applying the same CSI uncertainty model. As in STBC and STLC systems, which provide identical performance capabilities regardless of the channel uncertainty, MRC and MRT systems perform similarly given different levels of CSI uncertainty. For the case of MRC and MRT, as in the cases of the STBC and STLC systems, the effective channel gains, i.e., $\gamma_2$ and/or $\tilde{\gamma}_2$, are assumed to be known at the receivers.

### 2) SNR ANALYSIS OF $2 \times 2$ STLC UNDER CSI UNCERTAINTY ($\alpha = 4$)

Applying the channel uncertainty model to (18)–(22a) with the same assumptions applied in the case when $\alpha = 2$, we can derive the estimates of $x_1$ and $x_2$ for a $2 \times 2$ STLC system as follows:

$$\tilde{x}_1 = x_1 + \frac{(h_1\epsilon_1^* + h_2^*\epsilon_2 + h_3\epsilon_4^* - h_4^*\epsilon_3)\, x_1}{\gamma_4} + \frac{(h_1\epsilon_2^* - h_2^*\epsilon_1 + h_3\epsilon_3^* + h_4^*\epsilon_4)\, x_2}{\gamma_4} + \frac{\sqrt{\tilde{\gamma}_4}(z_{1,1} + z_{2,2}^*)}{\gamma_4}, \quad (30a)$$

$$\tilde{x}_2 = x_2 + \frac{\left(h_1\epsilon_1^* + h_2^*\epsilon_2 + h_3\epsilon_3^* + h_4^*\epsilon_4\right)x_2}{\gamma_4} + \frac{\left(-h_1\epsilon_2^* + h_2^*\epsilon_1 - h_3\epsilon_4^* + h_4^*\epsilon_3\right)x_1^*}{\gamma_4} + \frac{\sqrt{\tilde{\gamma}_4}(z_{2,1} - z_{1,2}^*)}{\gamma_4}. \tag{30b}$$

From (30a), we can readily derive the SNR of a 2 × 2 STLC system as

$$\mathrm{SNR}_{\mathrm{STLC}}(\gamma_4, \sigma_\epsilon^2) = \frac{\gamma_4^2\sigma_x^2}{2(\gamma_4\sigma_x^2\sigma_\epsilon^2 + (\gamma_4 + 4\sigma_\epsilon^2)\sigma_z^2)}. \tag{31}$$

As expected, (31) is reduced to (23) when $\sigma_\epsilon^2 = 0$.

Similarly, the estimates of $x_1$ and $x_2$ for a 2 × 2 STBC system are derived as follows:

$$\begin{aligned}\tilde{x}_1 = x_1 &+ \frac{1}{\gamma_4}\left(h_1\epsilon_1^* + h_2^*\epsilon_2 + h_3\epsilon_3^* + h_4^*\epsilon_4\right)x_1 \\ &+ \frac{1}{\gamma_4}\left(-h_1^*\epsilon_2 + h_2\epsilon_1^* - h_3^*\epsilon_4 + h_4\epsilon_3^*\right)x_2 \\ &+ \frac{\sqrt{2}(h_1^* + \epsilon_1^*)z_{1,1}}{\gamma_4} + \frac{\sqrt{2}(h_2 + \epsilon_2)z_{1,2}^*}{\gamma_4} \\ &+ \frac{\sqrt{2}(h_3^* + \epsilon_3^*)z_{2,1}}{\gamma_4} + \frac{\sqrt{2}(h_4 + \epsilon_4)z_{2,2}^*}{\gamma_4},\end{aligned} \tag{32a}$$

$$\begin{aligned}\tilde{x}_2 = x_2 &+ \frac{1}{\gamma_4}\left(h_1^*\epsilon_1 + h_2\epsilon_2^* + h_3^*\epsilon_3 + h_4\epsilon_4^*\right)x_2 \\ &+ \frac{1}{\gamma_4}\left(h_1\epsilon_2^* - h_2^*\epsilon_1 + h_3\epsilon_4^* - h_4^*\epsilon_3\right)x_1 \\ &+ \frac{\sqrt{2}(h_2^* + \epsilon_2^*)z_{1,1}}{\gamma_4} - \frac{\sqrt{2}(h_1 + \epsilon_1)z_{1,2}^*}{\gamma_4} \\ &+ \frac{\sqrt{2}(h_4^* + \epsilon_4^*)z_{2,1}}{\gamma_4} - \frac{\sqrt{2}(h_3 + \epsilon_3)z_{2,2}^*}{\gamma_4}.\end{aligned} \tag{32b}$$

From (32a), we can derive the SNR of a 2 × 2 STBC system and show that it is identical to that of a 2 × 2 STLC system, as follows:

$$\mathrm{SNR}_{\mathrm{STBC}}(\gamma_4, \sigma_\epsilon^2) = \mathrm{SNR}_{\mathrm{STLC}}(\gamma_4, \sigma_\epsilon^2). \tag{33}$$

In Fig. 7, four-channel STBC and STLC systems, i.e., $\alpha = 4$, are evaluated in terms of BER performance with BPSK modulation for a different MSEs of channel estimation. For the $\alpha = 4$ configuration, 1 × 4 MRC and 4 × 1 MRT systems are compared. The results verify the analyses in (31) and (33). As shown in (33), the new full-spatial-diversity scheme STLC is robust against CSI uncertainty such as STBC. As expected, the MRC and MRT schemes perform identically to each other with respect to the CSI uncertainty.

## IV. STLC DESIGNS FOR SYSTEMS WITH THREE AND FOUR RECEIVE ANTENNAS

In this section, we propose encoding and decoding schemes for complex orthogonal STLC for systems with three and four receive antennas. For brevity, we first introduce an encoding model with the encoding matrix and the transmitted

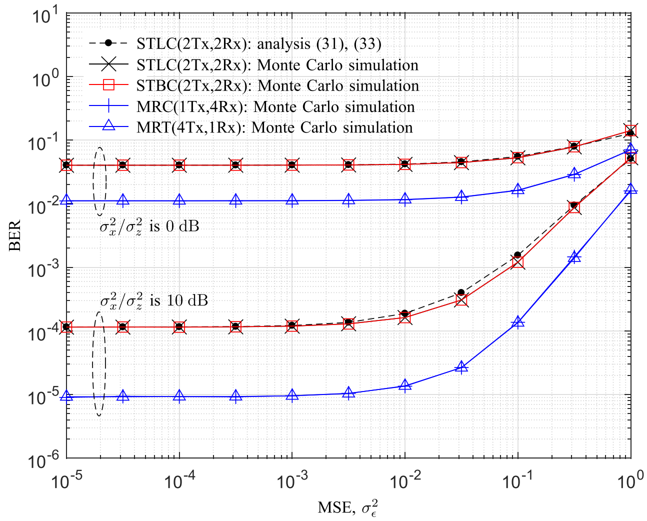


**FIGURE 7. BER performance over the MSE of CSI estimation for four-channel systems.**

information symbol vector $\boldsymbol{x} = [x_1 \cdots x_K]^T$. After showing the normalized STLC symbol vector $\bar{\boldsymbol{s}} = [\bar{s}_1 \cdots \bar{s}_T]^T$, whose code rate is $K/T$, the decoding model is provided to obtain $\tilde{\boldsymbol{x}} = [\tilde{x}_1 \cdots \tilde{x}_K]^T$, which is the input of an ML detector. The STLC, which is designed here, for a system with one transmit antenna and three receive antennas can be directly extended to a system that has multiple transmit antennas.

### A. RATE-$\frac{3}{4}$ STLC DESIGN FOR THREE RECEIVE ANTENNAS

The encoding procedure for rate-$\frac{3}{4}$ STLC is as follows:

$$\begin{bmatrix} s_1^* \\ s_2 \\ s_3^* \\ s_4^* \end{bmatrix} = \boldsymbol{C}_{3/4}^3\boldsymbol{x} = \begin{bmatrix} h_1 & h_2 & 0 \\ h_2^* & -h_1^* & h_3^* \\ 0 & h_3 & h_1 \\ h_3 & 0 & -h_2 \end{bmatrix}\begin{bmatrix} x_1 \\ x_2 \\ x_3 \end{bmatrix}, \tag{34}$$

where $\boldsymbol{C}_{K/T}^N$ is an STLC encoding matrix with rate-$K/T$ for a system with $N$ receive antennas and the two overlapped 2 × 2 STLC encoding matrices are highlighted by a dashed-line boxed. In (34), we use $\boldsymbol{C}_{2/2}^1$ and $-(\boldsymbol{C}_{2/2}^1)^*$ for constructing $\boldsymbol{C}_{3/4}^3$. Other 2 × 2 STLC encoding matrices introduced in Table 4 can also be used, and $\boldsymbol{C}_{3/4}^3$ in (IV-A) is not unique. For example, by extracting the first four rows and the firs three columns from $\boldsymbol{C}_{4/8}^3$ in (40), we can construct $\boldsymbol{C}_{3/4}^3$.

After the normalization of $\boldsymbol{s}$ in (IV-A), the transmitted STLC symbols are represented as follows:

$$\bar{s}_1 = \eta\left(h_1^*x_1^* + h_2^*x_2^*\right), \tag{35a}$$

$$\bar{s}_2 = \eta\left(h_2^*x_1 - h_1^*x_2 + h_3^*x_3\right), \tag{35b}$$

$$\bar{s}_3 = \eta\left(h_3^*x_2^* + h_1^*x_3^*\right), \tag{35c}$$

$$\bar{s}_4 = \eta\left(h_3^*x_1^* - h_2^*x_3^*\right), \tag{35d}$$

where the normalization factor $\eta = \sqrt{\frac{4}{3\gamma_3}}$, and it fulfills the average transmission power constraint as shown below:

$$\frac{1}{T}\sum_{t=1}^{t=T} |\eta s_t|^2 = \sigma_x^2. \quad (36)$$

Then, the STLC transmitter sends the STLC symbols $\bar{s}_1 \cdots \bar{s}_4$, sequentially, to the receiver. The received signals through the three receive antennas are written as follows:

$$\begin{bmatrix} r_{1,1} & \cdots & r_{1,4} \\ r_{2,1} & \cdots & r_{2,4} \\ r_{3,1} & \cdots & r_{3,4} \end{bmatrix} = \begin{bmatrix} h_1 \\ h_2 \\ h_3 \end{bmatrix} \begin{bmatrix} \bar{s}_1 & \cdots \bar{s}_4 \end{bmatrix} + \begin{bmatrix} z_{1,1} & \cdots & z_{1,4} \\ z_{2,1} & \cdots & z_{2,4} \\ z_{3,1} & \cdots & z_{3,4} \end{bmatrix}. \quad (37)$$

Now, we introduce a decoding scheme. By simply combining the received signals in (37), without using full CSI, we can decode the STLC signals as follows:

$$\tilde{x}_1 = r_{1,1}^* + r_{2,2} + r_{3,4}^* = gx_1 + z_{1,1}^* + z_{2,2} + z_{3,4}^*, \quad (38a)$$
$$\tilde{x}_2 = -r_{1,2} + r_{2,1}^* + r_{3,3}^* = gx_2 - z_{1,2} + z_{2,1}^* + z_{3,2}^*, \quad (38b)$$
$$\tilde{x}_3 = r_{1,3}^* - r_{2,4}^* + r_{3,2} = gx_3 + z_{1,3}^* - z_{2,4}^* + z_{3,2}, \quad (38c)$$

where the effective channel gain $g = \sqrt{\frac{4\gamma_3}{3}}$. Certainly, the decoded signal $\tilde{x}_k$ is a function of only $x_k$, which enables an independent and simple ML detection. It should be noted that even though only $\gamma_3$ is required for the ML detection at the receiver, it is not required for PSK symbol detection.

From (38a), we can derive the received SNR for $x_k$ as follows:

$$\mathsf{SNR}_k = \frac{4\gamma_3\sigma_x^2}{9\sigma_z^2}. \quad (39)$$

Here, it can be seen that the achieved diversity order is three, which is full, and the array gain is $4/3$.

### B. RATE-$\frac{1}{2}$ STLC DESIGN FOR THREE RECEIVE ANTENNAS

Here, we introduce the encoding and decoding structures for rate-$\frac{1}{2}$ STLC scheme. By decreasing the code rate from $3/4$ to $1/2$, we can construct the STLC encoding matrix without overlapping the $2 \times 2$ STLC matrices as follows:

$$\begin{bmatrix} s_1^* \\ s_2 \\ s_3 \\ s_4^* \\ s_5^* \\ s_6 \\ s_7 \\ s_8^* \end{bmatrix} = \boldsymbol{C}_{4/8}^3 \boldsymbol{x} = \begin{bmatrix} h_1 & h_2 & h_3 & 0 \\ h_2^* & -h_1^* & 0 & -h_3^* \\ h_3^* & 0 & -h_1^* & -h_2^* \\ 0 & -h_3 & -h_2 & h_1 \\ h_1 & h_2 & -h_3 & 0 \\ h_2^* & -h_1^* & 0 & h_3^* \\ h_3^* & 0 & h_1^* & h_2^* \\ 0 & -h_3 & h_2 & -h_1 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{bmatrix}, \quad (40)$$

where we use four $2 \times 2$ STLC encoding matrices. Any of the $2 \times 2$ STLC encoding matrices introduced in Table 4 can be used to construct a rate-$1/2$ STLC matrix $\boldsymbol{C}_{4/8}^3$ in (40).

After the normalization of $\boldsymbol{s}$ in (40), the transmitted STLC symbols are represented as follows:

$$\bar{s}_1 = \eta \left(h_1^* x_1^* + h_2^* x_2^* + h_3^* x_3^*\right), \quad (41a)$$
$$\bar{s}_2 = \eta \left(h_2^* x_1 - h_1^* x_2 - h_3^* x_4\right), \quad (41b)$$
$$\bar{s}_3 = \eta \left(h_3^* x_1 - h_1^* x_3 - h_2^* x_4\right), \quad (41c)$$
$$\bar{s}_4 = \eta \left(-h_3^* x_2^* - h_2^* x_3^* + h_1^* x_4^*\right), \quad (41d)$$
$$\bar{s}_5 = \eta \left(h_1^* x_1^* + h_2^* x_2^* - h_3^* x_3^*\right), \quad (41e)$$
$$\bar{s}_6 = \eta \left(h_2^* x_1 - h_1^* x_2 + h_3^* x_4\right), \quad (41f)$$
$$\bar{s}_7 = \eta \left(h_3^* x_1 + h_1^* x_3 + h_2^* x_4\right), \quad (41g)$$
$$\bar{s}_8 = \eta \left(-h_3^* x_2^* + h_2^* x_3^* - h_1^* x_4^*\right), \quad (41h)$$

where the power normalization factor $\eta = \frac{1}{\sqrt{\gamma_3}}$. Then, the STLC transmitter sends the STLC symbols $\bar{s}_1 \cdots \bar{s}_8$, sequentially, to the receiver, and the receiver combines the received signals to decode the STLC signals as follows:

$$\begin{aligned}\tilde{x}_1 &= r_{1,1}^* + r_{1,5}^* + r_{2,2} + r_{2,6} + r_{3,3} + r_{3,7} \\ &= gx_1 + z_{1,1}^* + z_{1,5}^* + z_{2,2} + z_{2,6} + z_{3,3} + z_{3,7}, \quad (42a)\end{aligned}$$
$$\begin{aligned}\tilde{x}_2 &= -r_{1,2} - r_{1,6} + r_{2,1}^* + r_{2,5}^* - r_{3,4}^* - r_{3,8}^* \\ &= gx_2 - z_{1,2} - z_{1,6} + z_{2,1}^* + z_{2,5}^* - z_{3,4}^* - z_{3,8}^*, \quad (42b)\end{aligned}$$
$$\begin{aligned}\tilde{x}_3 &= -r_{1,3} + r_{1,7} - r_{2,4}^* + r_{2,8}^* + r_{3,1}^* - r_{3,5}^* \\ &= gx_3 - z_{1,3} + z_{1,7} - z_{2,4}^* + z_{2,8}^* + z_{3,1}^* - z_{3,5}^*, \quad (42c)\end{aligned}$$
$$\begin{aligned}\tilde{x}_4 &= r_{1,4}^* - r_{1,8}^* - r_{2,3} + r_{2,7} - r_{3,2} + r_{3,6} \\ &= gx_4 + z_{1,4}^* - z_{1,8}^* - z_{2,3} + z_{2,7} - z_{3,2} + z_{3,6}. \quad (42d)\end{aligned}$$

where the effective channel gain $g = 2\sqrt{\gamma_3}$. The decoded signal $\tilde{x}_k$ is a function of only $x_k$, which enables an independent and simple ML detection. Again, it should be noted that though $\gamma_3$ is required for the ML detection at the receiver, it is not required for PSK symbol detection.

From (42a), we can derive the received SNR for $x_k$ as follows:

$$\mathsf{SNR}_k = \frac{2\gamma_3\sigma_x^2}{3\sigma_z^2}, \quad (43)$$

where it can be seen that the achieved diversity order is three and the array gain is two.

### C. RATE-$\frac{3}{7}$ STLC DESIGN FOR THREE RECEIVE ANTENNAS

Designing a low-code-rate (less than $1/2$) STLC encoding matrix from a high-code-rate STLC encoding matrix is straight forward. For example, the encoding procedure for rate-$\frac{3}{7}$ STLC is as follows:

$$\begin{bmatrix} s_1^* \\ s_2 \\ s_3 \\ s_4 \\ s_5 \\ s_6 \\ s_7 \end{bmatrix} = \boldsymbol{C}_{3/7}^{3}\boldsymbol{x} = \begin{bmatrix} h_1 & h_2 & 0 \\ h_2^* & -h_1^* & 0 \\ h_3^* & 0 & 0 \\ 0 & h_3^* & 0 \\ 0 & 0 & h_3^* \\ 0 & 0 & h_2^* \\ 0 & 0 & h_1^* \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \end{bmatrix}. \tag{44}$$

After the normalization of $\boldsymbol{s}$ in (44), the transmitted STLC symbols are represented as follows:

$$\bar{s}_1 = \eta\left(h_1^* x_1^* + h_2^* x_2^*\right), \tag{45a}$$
$$\bar{s}_2 = \eta\left(h_2^* x_1 - h_1^* x_2\right), \tag{45b}$$
$$\bar{s}_3 = \eta h_3^* x_1, \tag{45c}$$
$$\bar{s}_4 = \eta h_3^* x_2, \tag{45d}$$
$$\bar{s}_5 = \eta h_3^* x_3, \tag{45e}$$
$$\bar{s}_6 = \eta h_2^* x_3, \tag{45f}$$
$$\bar{s}_7 = \eta h_1^* x_3, \tag{45g}$$

where the power normalization factor $\eta = \sqrt{\frac{7}{3\gamma_3}}$.

Then, the STLC transmitter sends the STLC symbols $\bar{s}_1 \cdots \bar{s}_7$, sequentially, to the receiver, and the receiver combines the 21 received signals to decode the STLC signals as follows:

$$\tilde{x}_1 = r_{1,1}^* + r_{2,2} + r_{3,3} = g x_1 + z_{1,1}^* + z_{2,2} + z_{3,3}, \tag{46a}$$
$$\tilde{x}_2 = -r_{1,2} + r_{2,1}^* + r_{3,4} = g x_2 - z_{1,2} + z_{2,1}^* + r_{3,4}, \tag{46b}$$
$$\tilde{x}_3 = r_{1,5}^* + r_{2,6} + r_{3,7} = g x_3 + z_{1,5}^* + z_{2,6} + z_{3,7}, \tag{46c}$$

where $g = \sqrt{\frac{7\gamma_3}{3}}$. The decoded signal $\tilde{x}_k$ is a function of only $x_k$, which enables an independent and simple ML detection.

From (46a), we can derive the received SNR for $x_k$ as

$$\mathsf{SNR}_k = \frac{7\gamma_3\sigma_x^2}{9\sigma_z^2}. \tag{47}$$

Here, it can be seen that the achieved diversity order is three and the array processing gain is $7/3$.

### D. RATE-$\frac{3}{4}$ STLC DESIGN FOR FOUR RECEIVE ANTENNAS

The encoding procedure for rate-$\frac{3}{4}$ STLC is as follows:

$$\begin{bmatrix} s_1^* \\ s_2 \\ s_3^* \\ s_4 \\ s_5^* \\ s_6 \\ s_7^* \\ s_8 \end{bmatrix} = \boldsymbol{C}_{3/4}^{4}\boldsymbol{x} = \begin{bmatrix} h_1 & h_2 & 0 & h_4 & 0 & 0 \\ h_2^* & -h_1^* & h_3^* & 0 & 0 & 0 \\ 0 & h_3 & h_1 & 0 & h_4 & 0 \\ 0 & -h_4^* & 0 & h_2^* & h_3^* & 0 \\ 0 & 0 & 0 & h_3 & -h_2 & h_1 \\ 0 & 0 & -h_4* & 0 & h_1^* & h_2^* \\ -h_3 & 0 & h_2 & 0 & 0 & h_4 \\ h_4^* & 0 & 0 & -h_1^* & 0 & h_3^* \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \\ x_5 \\ x_6 \end{bmatrix}, \tag{48}$$

where four $2 \times 2$ STLC matrices are used with overlapping.

After the normalization of $\boldsymbol{s}$ in (48), the transmitted STLC symbols are represented as follows:

$$\bar{s}_1 = \eta\left(h_1^* x_1^* + h_2^* x_2^* + h_4^* x_4^*\right), \tag{49a}$$
$$\bar{s}_2 = \eta\left(h_2^* x_1 - h_1^* x_2 + h_3^* x_3\right), \tag{49b}$$
$$\bar{s}_3 = \eta\left(h_3^* x_2^* + h_1^* x_3^* + h_4^* x_5^*\right), \tag{49c}$$
$$\bar{s}_4 = \eta\left(-h_4^* x_2 + h_2^* x_4 + h_3^* x_5\right), \tag{49d}$$
$$\bar{s}_5 = \eta\left(h_3^* x_4^* - h_2^* x_5^* + h_1^* x_6^*\right), \tag{49e}$$
$$\bar{s}_6 = \eta\left(-h_4^* x_3 + h_1^* x_5 + h_2^* x_6\right), \tag{49f}$$
$$\bar{s}_7 = \eta\left(-h_3^* x_1^* + h_2^* x_3^* + h_4^* x_6^*\right), \tag{49g}$$
$$\bar{s}_8 = \eta\left(h_4^* x_1 - h_1^* x_4 + h_3^* x_6\right), \tag{49h}$$

where the power normalization factor $\eta = \sqrt{\frac{4}{3\gamma_4}}$.

Then, the STLC transmitter sends the STLC symbols $\bar{s}_1 \cdots \bar{s}_8$, sequentially, to the receiver, and the receiver combines the received signals to decode the STLC signals as follows:

$$\tilde{x}_1 = r_{1,1}^* + r_{2,2} - r_{3,7}^* + r_{4,8}$$
$$= g x_1 + z_{1,1}^* + z_{2,2} - z_{3,7}^* + z_{4,8}, \tag{50a}$$
$$\tilde{x}_2 = -r_{1,2} + r_{2,1}^* + r_{3,3}^* - r_{4,4}$$
$$= g x_2 - z_{1,2} + z_{2,1}^* + z_{3,3}^* - z_{4,4}, \tag{50b}$$
$$\tilde{x}_3 = r_{1,3}^* + r_{2,7}^* + r_{3,2} - r_{4,6}$$
$$= g x_3 + z_{1,3}^* + z_{2,7}^* + z_{3,2} - z_{4,6}, \tag{50c}$$
$$\tilde{x}_4 = -r_{1,8} + r_{2,4} + r_{3,5}^* + r_{4,1}^*$$
$$= g x_4 - z_{1,8} + z_{2,4} + z_{3,5}^* + z_{4,1}^*, \tag{50d}$$
$$\tilde{x}_5 = r_{1,6} - r_{2,5}^* + r_{3,4} + r_{4,3}^*$$
$$= g x_4 + z_{1,6} - z_{2,5}^* + z_{3,4} + z_{4,3}^*, \tag{50e}$$
$$\tilde{x}_6 = r_{1,5}^* + r_{2,6} + r_{3,8} + r_{4,7}^*$$
$$= g x_4 + z_{1,5}^* + z_{2,6} + z_{3,8} + z_{4,7}^*, \tag{50f}$$

where $g = \sqrt{\frac{4\gamma_4}{3}}$. Independent parallel ML detections of $x_k$'s are possible with a knowledge of $\gamma_4$. The value of $\gamma_4$ is not required for PSK symbol detection.

From (50), we can derive the received SNR for $x_k$ as follows:

$$\mathsf{SNR}_k = \frac{\gamma_4\sigma_x^2}{3\sigma_z^2}. \tag{51}$$

It can be seen that the achieved diversity order is four and the array processing gain is $4/3$.

### E. RATE-$\frac{4}{7}$ STLC DESIGN FOR FOUR RECEIVE ANTENNAS

The encoding procedure for rate-$\frac{4}{7}$ STLC is as follows:

$$\begin{bmatrix} s_1^* \\ s_2 \\ s_3 \\ s_4 \\ s_5 \\ s_6 \\ s_7 \end{bmatrix} = \boldsymbol{C}_{4/7}^4 \boldsymbol{x} = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & 0 & 0 \\ h_3^* & 0 & -h_1^* & 0 \\ h_4^* & 0 & 0 & -h_1^* \\ 0 & h_3^* & -h_2^* & 0 \\ 0 & h_4^* & 0 & -h_2^* \\ 0 & 0 & h_4^* & -h_3^* \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{bmatrix}. \tag{52}$$

After the normalization of $\boldsymbol{s}$, the transmitted STLC symbols are represented as follows:

$$\bar{s}_1 = \eta \left( h_1^* x_1^* + h_2^* x_2^* + h_3^* x_3^* + h_4^* x_4^* \right), \tag{53a}$$
$$\bar{s}_2 = \eta \left( h_2^* x_1 - h_1^* x_2 \right), \tag{53b}$$
$$\bar{s}_3 = \eta \left( h_3^* x_1 - h_1^* x_3 \right), \tag{53c}$$
$$\bar{s}_4 = \eta \left( h_4^* x_1 - h_1^* x_4 \right), \tag{53d}$$
$$\bar{s}_5 = \eta \left( h_3^* x_2 - h_2^* x_3 \right), \tag{53e}$$
$$\bar{s}_6 = \eta \left( h_4^* x_2 - h_2^* x_4 \right), \tag{53f}$$
$$\bar{s}_7 = \eta \left( h_4^* x_3 - h_3^* x_4 \right), \tag{53g}$$

where the power normalization factor $\eta = \sqrt{\frac{7}{4\gamma_4}}$.

Then, the STLC transmitter sends the STLC symbols $\bar{s}_1 \cdots \bar{s}_7$, sequentially, to the receiver, and the receiver combines the received signals to decode the STLC signals as follows:

$$\begin{aligned} \tilde{x}_1 &= r_{1,1}^* + r_{2,2} + r_{3,3} + r_{4,4} \\ &= g x_1 + z_{1,1}^* + z_{2,2} + z_{3,3} + z_{4,4}, \end{aligned} \tag{54a}$$
$$\begin{aligned} \tilde{x}_2 &= -r_{1,2} + r_{2,1}^* + r_{3,5} + r_{4,6} \\ &= g x_2 - z_{1,2} + z_{2,1}^* + z_{3,5} + z_{4,6}, \end{aligned} \tag{54b}$$
$$\begin{aligned} \tilde{x}_3 &= -r_{1,3} + r_{2,5} + r_{3,1}^* + r_{4,7} \\ &= g x_3 - z_{1,3} + z_{2,5} + z_{3,1}^* + z_{4,7}, \end{aligned} \tag{54c}$$
$$\begin{aligned} \tilde{x}_4 &= -r_{1,4}^* - r_{2,6} - r_{3,7} + r_{4,1}^* \\ &= g x_4 - z_{1,4}^* - z_{2,6} - z_{3,7} + z_{4,1}^*, \end{aligned} \tag{54d}$$

where $g = \sqrt{\frac{7}{4}\gamma_4}$. Clearly, with a knowledge of $\gamma_4$, an independent and parallel ML detection for each $x_k$ is possible, as the decoded signal $\tilde{x}_k$ is a function of only $x_k$. The value of $\gamma_4$ is not required for PSK symbol detection.

From (54a), we can derive the received SNR for $x_k$ as follows:

$$\mathsf{SNR}_k = \frac{7\gamma_4 \sigma_x^2}{16\sigma_z^2}. \tag{55}$$

It can be seen that the achieved diversity order is four and the array processing gain is 7/4.

### F. RATE-$\frac{1}{2}$ STLC DESIGN FOR FOUR RECEIVE ANTENNAS

The encoding procedure for rate-$\frac{1}{2}$ STLC is as follows:

$$\begin{bmatrix} s_1^* \\ s_2 \\ s_3^* \\ s_4 \end{bmatrix} = \boldsymbol{C}_{2/4}^4 \boldsymbol{x} = \begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \\ h_3 & h_4 \\ h_4^* & -h_3^* \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix}. \tag{56}$$

After the normalization of $\boldsymbol{s}$ in (56), the transmitted STLC symbols are represented as follows:

$$\bar{s}_1 = \eta \left( h_1^* x_1^* + h_2^* x_2^* \right), \tag{57a}$$
$$\bar{s}_2 = \eta \left( h_2^* x_1 - h_1^* x_2 \right), \tag{57b}$$
$$\bar{s}_3 = \eta \left( h_3^* x_1 + h_4^* x_2 \right), \tag{57c}$$
$$\bar{s}_4 = \eta \left( h_4^* x_1^* - h_3^* x_2^* \right), \tag{57d}$$

where the power normalization factor $\eta = \frac{2}{\sqrt{\gamma_4}}$.

Then, the STLC transmitter sends the STLC symbols $\bar{s}_1 \cdots \bar{s}_4$, sequentially, to the receiver, and the receiver combines the received signals to decode the STLC signals as follows:

$$\begin{aligned} \tilde{x}_1 &= r_{1,1}^* + r_{2,2} + r_{3,3}^* + r_{4,4} \\ &= g x_1 + z_{1,1}^* + z_{2,2} + z_{3,3}^* + z_{4,4}, \end{aligned} \tag{58a}$$
$$\begin{aligned} \tilde{x}_2 &= -r_{1,2} + r_{2,1}^* - r_{3,4} + r_{4,3}^* \\ &= g x_2 - z_{1,2} + z_{2,1}^* - z_{3,4} + z_{4,3}^*, \end{aligned} \tag{58b}$$

where $g = \sqrt{2\gamma_4}$.

The decoded signal $\tilde{x}_k$ is a function of only $x_k$, which enables an independent and simple ML detection. It should be noted that though the channel gain $\gamma_4$ is required for the ML detection at the receiver, it is not required for PSK symbol detection.

From (58a), we can derive the received SNR for $x_k$ as follows:

$$\mathsf{SNR}_k = \frac{\gamma_4 \sigma_x^2}{2\sigma_z^2}. \tag{59}$$

It can be seen that the achieved diversity order is four and the array processing gain is two.

### G. PERFORMANCE COMPARISON

The STLC schemes are denoted by **C1** to **C7**, and are summarized in Table 7. The STLC **C1** was designed in Section III, and STLCs **C2**–**C7** were designed in this Section. The designed STLC with rate-$K/T$ achieves an array processing gain of $MT/K$, i.e., $M$ divided by the code rate, and a diversity order of $MN$, i.e., full diversity order $\alpha$. For a given transmission data rate, the coding gain obtained by reducing the constellation size increases. However, since the array processing gain is inversely proportional to the code rate, it decreases. Therefore, for a given system configuration, i.e., $M$ and $N$, there is a tradeoff between coding gain and array processing gain with the same diversity order. This analysis

**TABLE 7.** Examples of orthogonal space–time line codes (STLC).

| Tx $M$ | Rx $N$ | Code | $K$ | $T$ | Code rate $(K/T)$ | Instantaneous SNR | Array gain $(M/\text{code rate})$ | Diversity order $(MN$: full) | Sections |
|---|---|---|---|---|---|---|---|---|---|
| $M$ | 2 | C1 | 2 | 2 | 1 | $(1/2)\gamma_{2M}\sigma_x^2/\sigma_z^2$ | $M$ | $2M$ | Section III |
| | 3 | C2 | 3 | 4 | 3/4 | $(4/9)\gamma_{3M}\sigma_x^2/\sigma_z^2$ | $4M/3$ | $3M$ | Section IV.A |
| | | C3 | 4 | 8 | 1/2 | $(2/3)\gamma_{3M}\sigma_x^2/\sigma_z^2$ | $2M$ | | Section IV.B |
| | | C4 | 3 | 7 | 3/7 | $(7/9)\gamma_{3M}\sigma_x^2/\sigma_z^2$ | $7M/3$ | | Section IV.C |
| | 4 | C5 | 6 | 8 | 3/4 | $(1/3)\gamma_{4M}\sigma_x^2/\sigma_z^2$ | $4M/3$ | $4M$ | Section IV.D |
| | | C6 | 4 | 7 | 4/7 | $(7/16)\gamma_{4M}\sigma_x^2/\sigma_z^2$ | $7M/4$ | | Section IV.E |
| | | C7 | 4 | 8 | 1/2 | $(1/2)\gamma_{4M}\sigma_x^2/\sigma_z^2$ | $2M$ | | Section IV.F |

**TABLE 8.** Examples of modulation schemes for BER comparison at transmission rate 3 bits/s/Hz and 1 bit/s/Hz.

| Code | Figs. 8 and 9 Data rate (bits/s/Hz) | Figs. 8 and 9 Modulation scheme for $K$ symbols | Figs. 10 and 11 Data rate (bits/s/Hz) | Figs. 10 and 11 Modulation scheme for $K$ symbols |
|---|---|---|---|---|
| C1 | 3.0 | two 8-PSK | 1.0 | two BPSK |
| C2 | 3.0 | three 16-QAM | 1.0 | one QPSK, two BPSK |
| C3 | 3.0 | four 64-QAM | 1.0 | four QPSK |
| C4 | 3.1 | two 256-QAM, one 64-QAM | 1.0 | one 16-QAM, one QPSK, one BPSK |
| C5 | 3.0 | six 16-QAM | 1.0 | two QPSK, four BPSK |
| C6 | 3.0 | three 64-QAM, one 8-PSK | 1.0 | one 16-QAM, three BPSK |
| C7 | 3.0 | four 64-QAM | 1.0 | four QPSK |

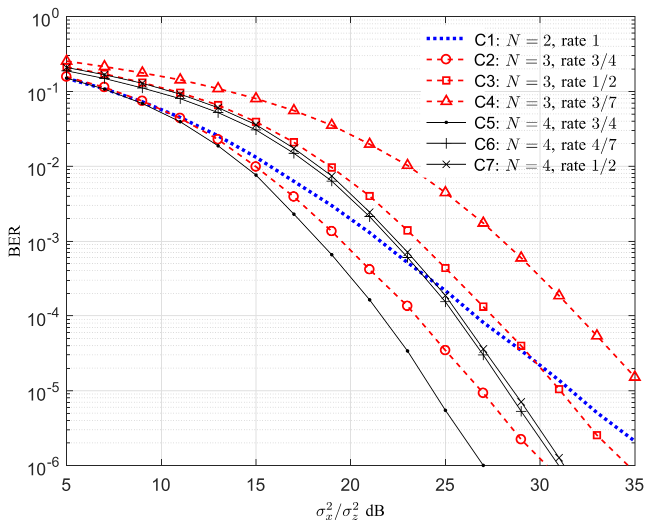


**FIGURE 8.** BER performance versus $\sigma_x^2/\sigma_z^2$ for space–time line codes at 3-bits/s/Hz; one transmit antenna, i.e., $M = 1$.

is verified by comparing the BER performances of the STLC schemes designed in this paper.

Fig. 8 shows the BER at a transmission data rate of 3 bits/s/Hz for STLC using one transmit antenna, i.e., $M = 1$. The modulation scheme used for each STLC is determined according to the code rate to achieve the transmission of 3 bits/s/Hz. For example, 8-PSK is used for **C1**. Since the STLC **C1** transmits two 8-PSK symbols (i.e., $K = 2$) for a symbol time of two (i.e., $T = 2$), the data rate is 3 bits/s/Hz. On the other hand, for STLC **C4** with a code rate of 3/7, three symbols are transmitted for a symbol time of seven. Here, two symbols are modulated by 256-quadrature amplitude modulation (QAM) to carry 16 bits and one symbol is modulated by 64-QAM to carry 6 bits. Thus, 22-bit information is transmitted with a symbol time of seven, resulting in a data rate of $22/7 \approx 3.1$ bits/s/Hz. Similarly, by adapting the modulation schemes, 3 bits/s/Hz are transmitted by **C2**, **C3**, **C5**, **C6**, and **C7** as listed in Table 8. Depending on the code rates, BPSK and quadrature PSK (QPSK) are involved, as well as 8-PSK, 16-QAM, and 64-QAM. Comparing the slope of the BER curve in the high signal-to-noise region, we see that the diversity orders of **C1**, **C2**–**C3**, and **C4**–**C7** are two, three, and four, respectively, i.e., $MN$, which is a full diversity order [21]. For the case of data transmission at 3 bits/s/Hz, a higher code rate provides better BER performance. For example, at a BER of $10^{-5}$, rate-3/4 **C2** gives about 4 dB gain over the use of rate-1/2 **C3**; and rate-3/4 **C5** gives about 4.2 dB gain over the use of rate-1/2 **C7**. From this, we can surmise that coding gain is more critical than array gain for high date rate transmission (compared to the results of low data rate transmissions, shown in Figs. 10 and 11). Similarly, it is seen that at a BER of $10^{-5}$, the rate-3/4 **C5** gives about 7.4 dB gain from the high diversity gain over the use of rate-1 **C1**.

In Fig. 9, we evaluate the BER performance of STLC using two transmit antennas, i.e., $M = 2$. The STLC scheme is determined according to the number of receive antennas $N$, as shown in Table 8. The same STLC scheme, which is used as the STLC scheme in a single receive antenna system, is

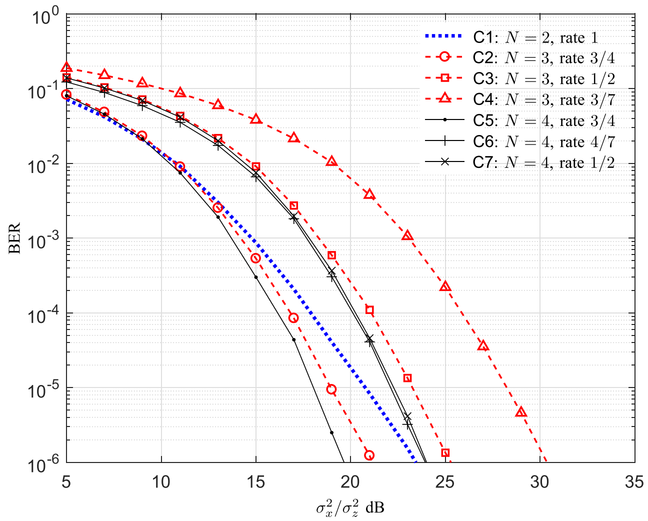


**FIGURE 9.** BER performance versus $\sigma_x^2/\sigma_z^2$ for space–time line codes at 3-bits/s/Hz using two transmit antennas, i.e., $M = 2$.

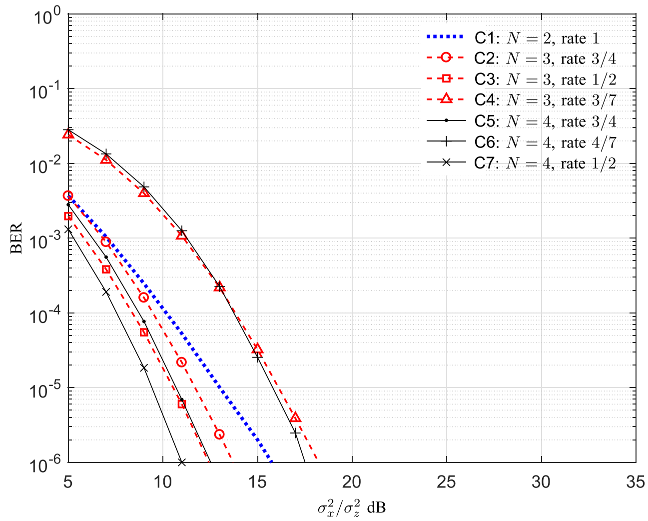


**FIGURE 11.** BER performance versus $\sigma_x^2/\sigma_z^2$ for space–time line codes at 1-bits/s/Hz using two transmit antennas, i.e., $M = 2$.

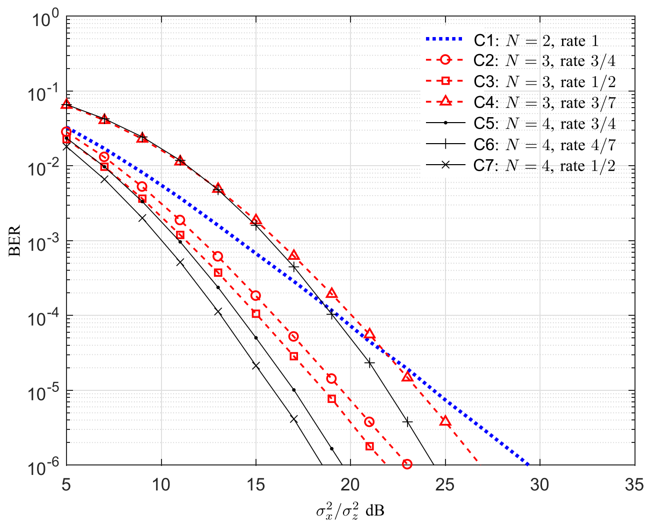


**FIGURE 10.** BER performance versus $\sigma_x^2/\sigma_z^2$ for space–time line codes at 1-bits/s/Hz using one transmit antenna, i.e., $M = 1$.

applied to each transmit antenna, and accordingly, the diversity order becomes double. Similar to the case when $M = 1$, a higher code rate provides better BER performance for 3 bits/s/Hz data transmission. For example, at a BER of $10^{-5}$, rate-3/4 **C2** and **C5** give about 4 dB gain over the use of rate-1/2 **C3** and **C7**, respectively. It is seen that at a BER of $10^{-5}$, the gain of the rate-3/4 **C5** over the use of rate-1 **C1** is reduced from 7.4 dB (Fig. 8) to 2.7 dB (Fig. 9), as the number of transmit antennas is increased from one to two. The reason is that much of the diversity gain is already achieved using the receive antennas.

The simulation results in Figs. 8 and 9 demonstrate that significant gains can be achieved by increasing the number of receive/transmit antennas with very little decoding complexity. This observation is valid for the data rate of 1 bit/s/Hz.

In Figs. 10 and 11, the BER performance at a data rate of 1 bit/s/Hz is evaluated for systems with one and two transmit antennas, respectively. For each designed STLC scheme, the modulation constellation is determined to transmit 1 bits/s/Hz as listed in Table 8. Comparing Figs. 8 and 10, we see that low data rate transmission achieves better BER performance, which is the expected result.

For a low data transmission rate of 1 bit/s/Hz, rate-4/7 **C6** outperforms rate-3/4 **C5**, which is the best code for the 3 bits/s/Hz data rate, as shown in Fig. 8. This implies that array processing gain is more critical than coding gain for low-rate data transmission. The diversity orders remain the same regardless of the data rate. The same observation is obtained for systems with two transmit antennas, as shown in Fig. 11.

## V. APPLICATION OF STLC

The proposed STLC is combined with STBC to support minimal-function lightweight devices, such as wearable devices, sensors, and many IoT devices, for which low cost, low complexity, and low power consumption are required [15]. To this end, using the reciprocity of the STBC and STLC schemes, we design a protocol-efficient STBC-and-STLC (STBLC) system, in which two devices, denoted by A and B, communicate with each other using TDD, as shown in Fig. 12. Note that a noncoherent detection causes significant performance degradation, and thus, the channel estimation is necessary at the receiver for coherent detection [22]. However, minimal-functional lightweight devices may have the limited essential functions including time and frequency synchronization, direct current offset estimation, and channel estimation just before the data transmission. Therefore, it is difficult to support such device without significant performance degradation if no CSI is available. To resolve this issue, the STBLC reifies and verifies STLC's benefits: i) the STLC enables the minimal operation of the lightweight devices and ii) STBLC operating in a TDD mode

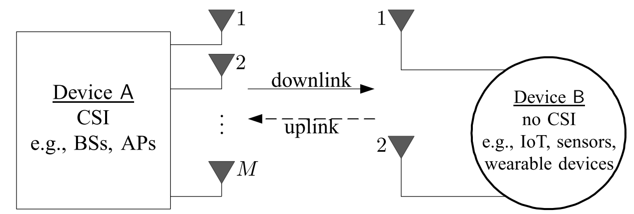


**FIGURE 12.** System model considered in this study. Device A has full CSI, while device B has no CSI. Downlink and uplink are directions from device A to device B and from device B to device A, respectively.

requires less frequent channel estimations while sustaining full-diversity gain.

We first design a $1 \times 2$ STBLC system and then extend it to an $M \times 2$ STBLS system. Using the various rates in Section IV, the proposed STBLC can be readily extended to $M \times 3$ and $M \times 4$ STBLS systems. Numerical results verify that the proposed STBLC outperforms an existing scheme. The proposed STBLC framework would have various potential applications in communication systems.

### A. STBLC SYSTEM MODEL

For convenience, the direction of communication from device A to B is denoted by downlink and that from device B to A by uplink. Device A has $M$ antennas and CSI, while device B has two antennas without CSI. In other words, device A has a CSI estimation function and more RF chains than device B. Base stations (BSs) and access points (APs) are the example applications of device A, while UE, wearable devices, and simple IoT devices are the example applications of minimally functional device B. For simple description, a single antenna, i.e., $M = 1$, is assumed at device A, as shown in Fig. 13. The system and results in this section can be directly extended to the case of $M > 1$. Now, we introduce the details of the proposed STBLC system in uplink and downlink communications.

#### 1) UPLINK COMMUNICATIONS USING STBC (PHASE 1)

First, device B transmits a pilot symbol $p_1$ using the first transmit antenna and then the next pilot symbol $p_2$ using the second transmit antenna so that device A estimates the channels $h_1$ and $h_2$. Without loss of generality, the pilot symbols are set to an arbitrary real value, i.e., $p_t = \sqrt{P}$, where $P$ is the transmit power constraint of device B.

The signal received at device A is then expressed as follows:

$$\begin{bmatrix} r_{A,1} & r_{A,2} \end{bmatrix} = \begin{bmatrix} h_1 & h_2 \end{bmatrix} \begin{bmatrix} p_1 & 0 \\ 0 & p_2 \end{bmatrix} + \begin{bmatrix} z_{A,1} & z_{A,2} \end{bmatrix}. \tag{60}$$

From (60), the minimum MSE (MMSE)-based estimates of channels are obtained as

$$\tilde{h}_n = \frac{1}{p_n} r_{A,n} = h_n + \epsilon_n, \tag{61}$$

where the channel estimation error $\epsilon_n$ is defined as $\epsilon_n \triangleq z_{A,n}/\sqrt{P}$. With no boosting of the pilot signal, w.l.o.g., we set $P = \sigma_x^2 = 1$, and accordingly, $\epsilon_n$ conforms to a normal distribution with a variance $\sigma_z^2$, i.e., $\epsilon_n \sim \mathcal{CN}(0, \sigma_z^2)$, which is the same as the AWGN.

After transmitting two pilot symbols, the device B sends information symbols, denoted by $b_1$ and $b_2$, and for simple description, $b_t$ is assumed to be modulated by a PSK symbol, where $\mathrm{E}\,|b_t|^2 = \sigma_x^2 = 1$. Device B encodes $b_1$ and $b_2$ to a $2 \times 2$ STBC symbol matrix and transmits it through two antennas for two symbol times. Substituting $\sigma_x^2 = 1$ and $\sigma_\epsilon^2 = \sigma_z^2$ to analyses in (25a)–(29), the received SNR at device A is derived as

$$\mathsf{SNR}_A^{\mathrm{STBC}} = \frac{\gamma_2^2}{4\sigma^2(\gamma_2 + \sigma^2)}. \tag{62}$$

#### 2) DOWNLINK COMMUNICATIONS USING STLC (PHASE 2)

For the downlink communications, i.e., direction from device A to device B, device A transmits PSK modulated symbols $a_1$ and $a_2$ to device B. To achieve full-diversity gain using a single transmit antenna at device A, device A encodes $a_1$ and $a_2$ by using the estimated CSI, namely $\tilde{h}_1$ and $\tilde{h}_2$, during the previous uplink communications, i.e., STLC encoding. From (29), we can derive the received SNR of device B as follows:

$$\mathsf{SNR}_B^{\mathrm{STLC}} = \mathsf{SNR}_A^{\mathrm{STBC}} = \frac{\gamma_2^2}{4\sigma^2(\gamma_2 + \sigma^2)}. \tag{63}$$

#### 3) EXTENSION TO DEVICE A WITH MULTIPLE ANTENNAS

For the STLC case, the $M$ STLC encodings are independent of one another, and the corresponding STLC decoding at device B is the same as the single-antenna STLC decoding as stated in Section III-B. Similarly, for the STBC decoding case, each receive antenna performs the STBC decoding in (28a), and all the estimates obtained from the multiple receive antennas are directly combined [2]. Of course, the multiple receive antennas at device A do not require any modification of device B.

### B. MERITS OF AN STBLC SCHEME

The STBLC system has three important merits: i) minimal-function device design, ii) less frequent channel estimations, and iii) easier channel estimation.

First, owing to the new diversity scheme STLC, we can design a device for dedicated and minimal capabilities. As assumed in this study, device A has $M$ antennas with CSI, while device B has two antennas without CSI, i.e., the minimal functions. Even with the minimal functions, full-diversity gain can be achieved for both directions of communications, namely during the uplink and downlink communications. If two devices are communicating and one is superior to the other in terms of hardware and functional capability, they can be considered as devices A and B, respectively, and achieve full diversity by using an STBLC scheme.

Second, channel estimation is less frequently required. In typical communications, both devices involved in communications always require the CSI as depicted in Fig. 14.

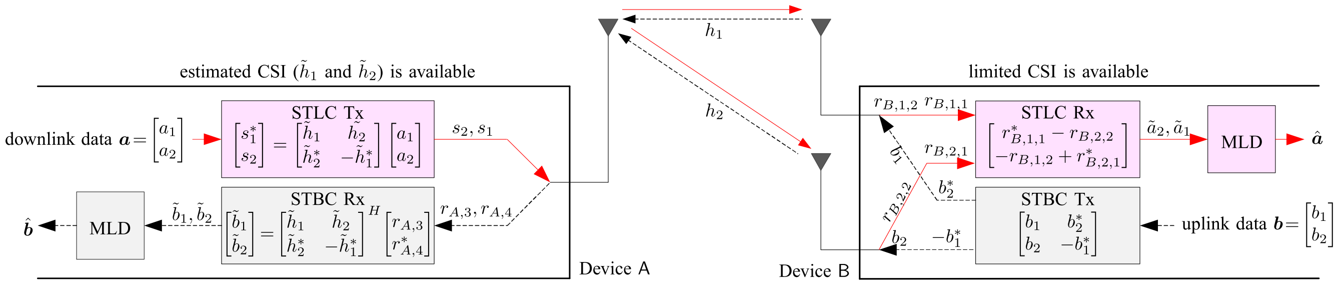


**FIGURE 13.** **Proposed STBLC system achieving full-spatial diversity with STBC and STLC methods. Device A has full CSI, while device B has no CSI. Uplink and downlink are directions from device B to device A and from device A to device B, respectively.**

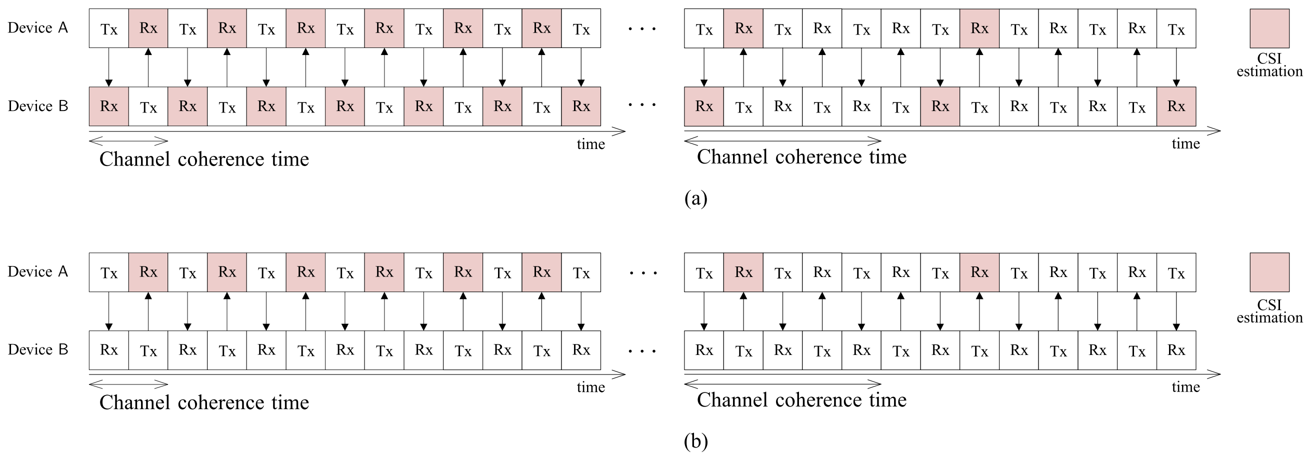


**FIGURE 14.** **Communication protocols. (a) A typical coherent detection system. (b) An STBLC system.**

However, the CSI is required at device A only in the STBLC framework. Thus, if the channel variation is not severe, as in a block fading channel, the CSI, which is used for STBC symbol decoding at device A, can be reused for the consecutive STLC encoding. Once the device A obtains the CSI, it can use the CSI for STBC decoding and also STLC encoding, consecutively, during channel coherence time. Roughly, the STBLC system needs CSI estimations at half the frequency of the estimations of a conventional coherent system, and this is true as the channel coherence time increases. Furthermore, as we mentioned previously, the CSI estimation is performed at device A only.

The third merit is the ease of channel estimation. Device A can have a large number of antennas, e.g., a few hundreds of antennas at a massive MIMO system [23], while device B has only two antennas. Therefore, the CSI can be readily estimated at device A by using only two orthogonal pilot symbols and/or training sequences that are transmitted from device B [24]. However, at least $M$-orthogonal pilot symbols and/or training sequences are required to estimate the CSI at device B, which causes a significant decrease of downlink spectral efficiency. Note that the proposed STBLC requires CSI at device A only.

### *C. BENCHMARKING SYSTEMS*

In this section, we introduce two benchmarking systems to justify the proposed STBLC system. The first benchmarking system employs STBC for the uplink communications, and it uses a preprocessing weight $w$, which is obtained from the estimated CSI at device A. The received downlink signal at device B is expressed as follows:

$$\begin{bmatrix} r_{B,1,1} \\ r_{B,2,1} \end{bmatrix} = \begin{bmatrix} h_1 \\ h_2 \end{bmatrix} w a_1 + \begin{bmatrix} z_{B,1,1} \\ z_{B,2,1} \end{bmatrix}, \tag{64}$$

where the preprocessing weight $w$ follows $|w| = 1$. Since there is no CSI at device B, the two received signals are combined as

$$r_{B,1,1} + r_{B,2,1} = (h_1 + h_2) w a_1 + z_{B,1,1} + z_{B,2,1}, \tag{65}$$

which is a naive combination. The preprocessing weight $w$ is then designed such that the device B achieves the maximum effective channel gain in (65) as follows:

$$w = \frac{(\tilde{h}_1 + \tilde{h}_2)^*}{|\tilde{h}_1 + \tilde{h}_2|}, \tag{66}$$

which is addressed by a beamforming (BF) scheme.[1] Using $w$ in (66) to (65), the combined signal becomes

$$\begin{aligned} r_{B,1,1} + r_{B,2,1} &= \frac{(h_1 + h_2)(\tilde{h}_1 + \tilde{h}_2)^*}{|\tilde{h}_1 + \tilde{h}_2|} a_1 + z_{B,1,1} + z_{B,2,1} \\ &= |h_1 + h_2 + \epsilon_1 + \epsilon_2|\, a_1 \\ &\quad - (\epsilon_1 + \epsilon_2) \frac{(h_1^* + h_2^* + \epsilon_1^* + \epsilon_2^*)}{|h_1 + h_2 + \epsilon_1 + \epsilon_2|} a_1 \\ &\quad + z_{B,1,1} + z_{B,2,1}, \end{aligned} \tag{67}$$

where the effective channel gain of BF is $|h_1 + h_2 + \epsilon_1 + \epsilon_2|$.

[1]The term BF is a general term for preprocessing with multiple transmit antennas at device A, which will be introduced in (69).

The second benchmarking scheme is a no-diversity scheme with a single antenna at devices A and B, i.e., a single-input single-output (SISO) system. For the uplink communications through the SISO channel, denoted by $h$, device B sends $b$ without any processing and device A performs receive processing with $\tilde{h}^*$, such that the uplink effective channel gain is maximized. Similarly, for the downlink communications, device A sends $a$ with preprocessing $\tilde{h}^*/|\tilde{h}|$. Thus, the downlink effective channel gain becomes a real value such that device B can detect $a$ without any receiver processing, such as combining and equalization. The estimates of $b$ and $a$ in the uplink and downlink communications, respectively, are expressed as follows:

$$\begin{aligned}\tilde{b} &= \tilde{h}^* r_{A,1,1} = \tilde{h}^*(hb + z_{A,1,1}) \\ &= |h|^2 b + \epsilon^* hb + (h^* + \epsilon^*) z_{B,1,1}, \end{aligned} \tag{68a}$$

$$\begin{aligned}\tilde{a} &= r_{B,1,1} = h\frac{\tilde{h}^*}{|\tilde{h}|}a + z_{B,1,1} \\ &= |h+\epsilon|a - \frac{\epsilon h^* + |\epsilon|^2}{|h+\epsilon|}a + z_{A,1,1}. \end{aligned} \tag{68b}$$

For the $M$ multiple antennas at device A, to sustain the structure of device B, i.e., the naive combination in (65), the combined receive signals are written as follows:

$$r_{B,1,1} + r_{B,2,1} = (\boldsymbol{h}_1 + \boldsymbol{h}_2)\, \boldsymbol{w} a_1 + z_{B,1,1} + z_{B,2,1}, \tag{69}$$

where channel $\boldsymbol{h}_n$ is a $1 \times M$ row vector, the $m$th element of which is the channel between the $m$th transmit antenna of device A and the $n$th receive antenna of device B, and $\boldsymbol{w}$ is an $M \times 1$ BF vector. The preprocessing vector $\boldsymbol{w}$ is defined as

$$\boldsymbol{w} = \frac{(\tilde{\boldsymbol{h}}_1 + \tilde{\boldsymbol{h}}_2)^H}{\|\tilde{\boldsymbol{h}}_1 + \tilde{\boldsymbol{h}}_2\|}, \tag{70}$$

where $\tilde{\boldsymbol{h}}_n = \boldsymbol{h}_n + \boldsymbol{\epsilon}_n$ and $\boldsymbol{\epsilon}_n$ is a CSI estimation error vector. Using $\boldsymbol{w}$ in (70) to (69), the received signal becomes

$$\begin{aligned} r_{B,1,1} + r_{B,2,1} &= \frac{(\boldsymbol{h}_1 + \boldsymbol{h}_2)\left(\tilde{\boldsymbol{h}}_1 + \tilde{\boldsymbol{h}}_2\right)^H}{\left\|\tilde{\boldsymbol{h}}_1 + \tilde{\boldsymbol{h}}_2\right\|} a_1 + z_{B,1,1} + z_{B,2,1} \\ &= \|\boldsymbol{h}_1 + \boldsymbol{h}_2 + \boldsymbol{\epsilon}_1 + \boldsymbol{\epsilon}_2\|\, a_1 \\ &\quad - (\boldsymbol{\epsilon}_1 + \boldsymbol{\epsilon}_2) \frac{\left(\boldsymbol{h}_1^H + \boldsymbol{h}_2^H + \boldsymbol{\epsilon}_1^H + \boldsymbol{\epsilon}_2^H\right)}{\|\boldsymbol{h}_1 + \boldsymbol{h}_2 + \boldsymbol{\epsilon}_1 + \boldsymbol{\epsilon}_2\|} a_1 \\ &\quad + z_{B,1,1} + z_{B,2,1}, \end{aligned} \tag{71}$$

where the effective channel gain of BF is $\|\boldsymbol{h}_1 + \boldsymbol{h}_2 + \boldsymbol{\epsilon}_1 + \boldsymbol{\epsilon}_2\|$.

### D. PERFORMANCE COMPARISON

The three systems summarized in Table 9 are compared. Here, for reference, the SISO system is also included. An STBC-BF system is the benchmarking system introduced in Section V-C. The STBC-BF uses STBC for uplink communication from device B to device A, while BF is employed for downlink communication from device A to device B. An STBLC system is the proposed system in Sections V-A, in which STBC and STLC are used for uplink and downlink, respectively. BPSK and QPSK modulations are used in the BER simulation. Device A has $M$ antennas with CSI, while device B has two antennas.

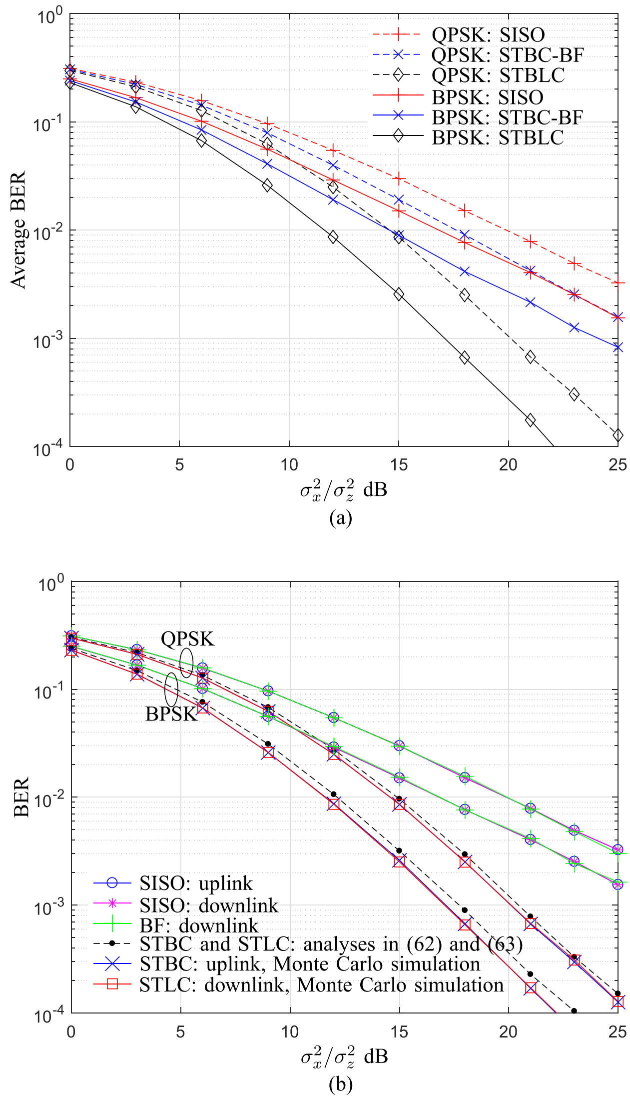


**FIGURE 15.** BER performance across $\sigma_x^2/\sigma_z^2$ when device A has one antenna and device B has two antennas, i.e., $M = 1$. (a) Average BER of uplink and downlink. (b) BER of each link.

**TABLE 9.** List of compared systems: device A has $M$ antennas with CSI and device B has two antennas with no CSI.

| Compared systems | Uplink ($\beta \to \alpha$) | Downlink ($\alpha \to \beta$) |
|---|---|---|
| SISO | no preprocessing | preprocessing |
| STBC-BF | $2 \times M$ STBC | $M \times 2$ BF |
| STBLC | $2 \times M$ STBC | $M \times 2$ STLC |

In Fig. 15(a), the average BER of uplink and downlink communications is shown to compare the system performance when $M = 1$. Since the proposed STBLC system achieves full-spatial-diversity gain (order of two) in both uplink and downlink communications, it provides the best performance. Clearly, the proposed method achieves higher-spatial diversity gain compared to SISO and STBC-BF

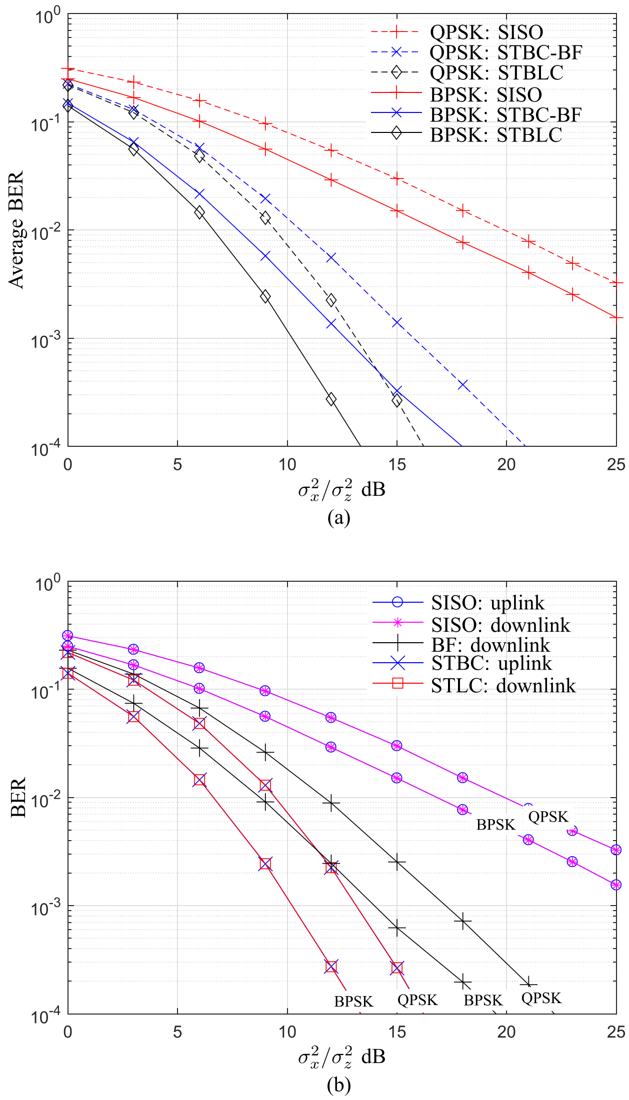


**FIGURE 16. Monte Carlo simulation results of BER performance across $\sigma_x^2/\sigma_z^2$ when device A has two antennas and device B has two antennas, i.e., $M = 2$. (a) Average BER of uplink and downlink. (b) BER of each link.**

systems. To clearly observe the cause of the performance gap, in Fig. 15(b), the BER performance of an each link is shown. Here, we note that BF does not achieve diversity gain as in a SISO system because device B does not have CSI. From the results, we also verify the analyses in (62) and (63).

In Fig. 16, the BER performance is evaluated when $M = 2$. From the results in Fig. 16(a), we see that the spatial-diversity orders of SISO, STBC-BF, and STBLC are one, two, and four, respectively. Therefore, STBLC system outperforms the others. In Fig. 16(b), the BER performance of each link is shown. Here, we see that BF with two antennas at device A increases diversity gain compared to the SISO system.

## VI. CONCLUSION

In this paper, full-spatial-diversity-achieving STLC was newly proposed. The proposed STLC uses channel information for the encoding at the transmitter and it enables the receiver to decode the STLC symbols without full channel information. Throughout the rigorous SNR analysis, BER simulation, and application example, the merits of STLC were verified. The proposed STLC can also be extended to exploit frequency diversity by using two adjacent carriers instead of two adjacent symbol periods. The proposed STLC is expected to be applied to many applications of various MIMO systems desiring full-spatial diversity gain.

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**JINGON JOUNG** (S'03–M'07–SM'15) 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 the Korea Advanced Institute of Science and Technology (KAIST), Daejeon, South Korea, in 2003 and 2007, respectively.

He was a Scientist with the Institute for Infocomm Research (I2R), Agency for Science, Technology and Research (A∗STAR), Singapore. He was a Post-Doctoral Research Scientist with KAIST, and a Post-Doctoral Fellow with UCLA, CA, USA. He is currently a Professor with the School of Electrical and Electronics Engineering, Chung-Ang University, Seoul, South Korea, and a Principal Investigator of the Wireless Systems Laboratory. His research activities are in the area of multiuser systems, multiple-input multiple-output communications, and cooperative systems. His current research area/interest includes energy-efficient ICT, IoT, and machine learning algorithms.

Dr. Joung was a recipient of the First Prize at the Intel-ITRC Student Paper Contest in 2006. He has been serving on the Editorial Board of the *APSIPA Transactions on Signal and Information Processing* since 2014. He also served as a Guest Editor of the IEEE Access for special section Recent Advanced in Full-Duplex Radios and Networks in 2016. He is recognized as an Exemplary Reviewer from the IEEE Communications Letters in 2012, and from the IEEE Wireless Communications Letters in 2012 and 2013.