

Received 24 May 2025, accepted 7 June 2025, date of publication 12 June 2025, date of current version 23 June 2025.

Digital Object Identifier 10.1109/ACCESS.2025.3579312

RESEARCH ARTICLE

Channel Phase Nonreciprocity Compensation for Space-Frequency Line Coded OFDM Systems

HAN-GYEOL LEE¹, (Student Member, IEEE), JIHOON CHOI², (Senior Member, IEEE),
JUYEOP KIM³, (Member, IEEE), KOICHI ADACHI⁴, (Senior Member, IEEE),
YUE XIAO⁵, (Member, IEEE), AND JINGON JOUNG¹, (Senior Member, IEEE)

¹School of Electrical and Electronics Engineering, Chung-Ang University, Seoul 06974, South Korea

²School of Electronics and Information Engineering, Korea Aerospace University, Goyang, Gyeonggi-do 10540, South Korea

³Department of Electrical Engineering, Sookmyung Women's University, Seoul 04312, South Korea

⁴Department of Information and Computer Science, Keio University, Yokohama 223-8522, Japan

⁵National Key Laboratory of Wireless Communications, University of Electronic Science and Technology of China, Chengdu 611731, China

Corresponding authors: Jihoon Choi (jihoon@kau.ac.kr) and Jingon Jung (jjyoung@cau.ac.kr)

This work was supported in part by the National Research Foundation of Korea (NRF) grant funded by Korean Government [MSIT: Ministry of Science and Information and Communications Technology (ICT)] under Grant RS-2024-00405510, and in part by the Institute of Information and Communications Technology Planning and Evaluation (IITP) grant funded by Korean Government (MSIT) (Development of Next Generation Wireless Access Technology Based on Space Time Line Code, 30%, and Development of 5G Industrial Terminal Technology Supporting 28 GHz Band/Private 5G Band/NR-U Band, 20%) under Grant 2021-0-00874 and Grant 2022-0-00635.

ABSTRACT This paper investigates the impact of phase nonreciprocity in space-frequency line code-orthogonal frequency-division multiplexing (SFLC-OFDM) systems, leading to significant performance degradation due to non-ideal multi-channel radio-frequency transceivers. A novel demodulation reference signal (DMRS) structure with phase rotation is proposed to address this challenge, enabling effective phase nonreciprocity compensation. To compensate for the channel phase nonreciprocity, the user equipment, as the uplink (UL) transmitter, estimates the downlink (DL) phase offset (PO) and then embeds this offset information into the phase of UL DMRS sequences. The base station, as the UL receiver, extracts the embedded DL PO and estimates the UL PO from the received DMRS sequences. The DL and UL PO values are then used at the base station to compensate for phase discrepancies between the UL and DL channels during the UL channel estimation process. Extensive numerical simulations across various system parameters validate the effectiveness of the proposed method in mitigating phase nonreciprocity and ensuring reliable SFLC-OFDM performance.

INDEX TERMS Space-frequency line code (SFLC), channel reciprocity, orthogonal frequency-division multiplexing (OFDM), reference signal.

I. INTRODUCTION

A recently developed space-time line code (STLC) is a symmetric counterpart to space-time block code, achieving full spatial diversity [1]. An STLC transmitter (Tx) encodes information symbols using channel state information (CSI). In contrast, an STLC receiver (Rx) decodes the symbols through a linear combination of received signals, requiring only partial CSI related to the channel norm or no CSI in the case of M-ary phase shift keying. A key advantage of the

STLC scheme is the low-complexity decoding structure independent of the number of Tx antennas. This feature enables the STLC system to support scalable and flexible antenna configurations, contributing to its widespread adoption in various multiple-input and multiple-output (MIMO) systems, as summarized in Table 1.

In addition to extensive theoretical research, significant efforts have been made to implement the STLC system on hardware. In [20], a real-time testbed was developed by combining orthogonal frequency division multiplexing (OFDM) with STLC, resulting in the space-frequency line code (SFLC)-OFDM system. However, the testbed relied on

The associate editor coordinating the review of this manuscript and approving it for publication was Miguel López-Benítez^{1b}.

TABLE 1. Various MIMO systems employing STLC schemes.

MIMO Systems Employing STLC	References
Multiuser transmission	[2]–[6]
Multistream transmission	[7], [8]
Multicast systems	[9]
Relay systems	[10]–[12]
Intelligent reflecting surface-aided systems	[13], [14]
Secure communications	[15]–[17]
Non-orthogonal multiple access techniques	[18], [19]

CSI feedback through a wired connection, which is unavailable for practical deployments. As an alternative to wired CSI feedback, an SFLC-OFDM system based on time division duplexing (TDD) was designed that transfers the uplink (UL) demodulation reference signal (DMRS) [21]. Using channel reciprocity between the UL and downlink (DL) channels [22], the UL CSI estimated from the received DMRS is used for SFLC encoding in subsequent transmission of the DL signal. The TDD-based SFLC-OFDM system effectively substitutes the wired CSI feedback. However, the implementation in [21] focuses primarily on the SFLC-OFDM system equipped with a single Tx antenna.

Extending the TDD SFLC-OFDM system from a single Tx antenna to multiple Tx antenna configurations is challenging. One of the key issues is the antenna-specific phase offset (PO) caused by the initial phase of the phase-locked loop (PLL) [23], [24], [25], [26]. In TDD communication, a single radio frequency (RF) port alternates between Tx and Rx modes, with two independent PLLs operating separately for each mode. The PO difference introduced by these PLLs within the same antenna port leads to a phase discrepancy between the UL and DL channels, disrupting channel reciprocity. PO compensation is essential for reliable communications in TDD SFLC-OFDM systems with multiple Tx antennas to ensure channel reciprocity.

Because PLL-induced POs appear as common phase errors (CPEs) [26], various CPE correction techniques, including those developed in [27], [28], [29], [30], [31], [32], [33], can be used to estimate and compensate for these POs. However, when applied to TDD MIMO SFLC-OFDM systems, existing methods require DL DMRS transmissions across the full system bandwidth, such as joint CSI, CPE, and/or carrier frequency offset estimators [30], [31], [32], [33], resulting in prohibitive DL overhead. Other low-overhead techniques, such as Kalman filtering [27], a decision-directed estimator [28], and a codebook-based approach [29], are limited to SISO configuration, requiring substantial structural modifications to support MIMO deployments. These limitations motivate the development of a PLL-induced PO compensation method for SFLC-OFDM systems, maintaining a low DL overhead and MIMO capability.

This paper proposes a novel approach to resolve the phase nonreciprocity problem in TDD SFLC-OFDM systems equipped with multiple Tx antennas. Throughout the paper, the Tx and Rx of the SFLC-OFDM system are referred

TABLE 2. Comparison of proposed and existing methods in terms of DL overhead and antenna configuration.

Reference	DL Overhead	Antenna Configuration
[27]	Low	SISO
[28]	Low	SISO
[29]	Low	SISO
[30]	Moderate	SISO
[31]	High	MIMO
[32]	High	SISO
[33]	High	SISO
Proposed	Low	MIMO

to as the base station (BS) and user equipment (UE), respectively, accounting for the structural similarity to the fifth generation new radio (5G NR) standard. During DL communication, the BS transmits phase tracking reference signal (PTRS) sequences on a small set of subcarriers. The UE processes the received PTRS to estimate the DL POs. In the UL, the UE embeds the DL PO information into the phases of the UL DMRS via phase rotation (PR). By leveraging the phase-rotated DMRS, the BS estimates and compensates for the phase discrepancies between the DL and UL channels. The proposed method achieves accurate PLL-induced PO compensation with low DL overhead and MIMO compatibility. These capabilities are absent in existing CPE correction approaches. Table 2 compares the proposed and existing methods in terms of DL overhead and supported antenna configurations.

The main contributions of this paper are summarized as follows:

- To the best of the authors' knowledge, this is the first study to investigate phase nonreciprocity caused by PLL-induced POs in TDD MIMO SFLC-OFDM systems. Analytical expressions are derived to characterize the effects of PLL-induced POs, emphasizing the need for effective compensation techniques to ensure reliable SFLC-OFDM performance.
- The DMRS-PR method is proposed to convey the DL PO information estimated in the UE to the BS to avoid additional UL overhead and minimize an increase in DL overhead. Specifically, the PTRS is assigned to DL subcarriers for PO estimation in the UE, and the estimated PO information is embedded in a part of the UL DMRS.
- Simulation results demonstrate that the proposed DMRS-PR method effectively mitigates phase nonreciprocity in TDD-based SFLC-OFDM MIMO systems. In addition, comprehensive evaluations of various system parameters provide insight into the conditions and configurations required to ensure reliable communications using the proposed method.

The rest of this paper is organized as follows: Section II introduces the signal and system models for the TDD MIMO SFLC-OFDM system and analyzes the effects of phase nonreciprocity on SFLC decoding. Section III presents the detailed design of the proposed PR method to address the phase nonreciprocity issue. In Section IV, simulation results

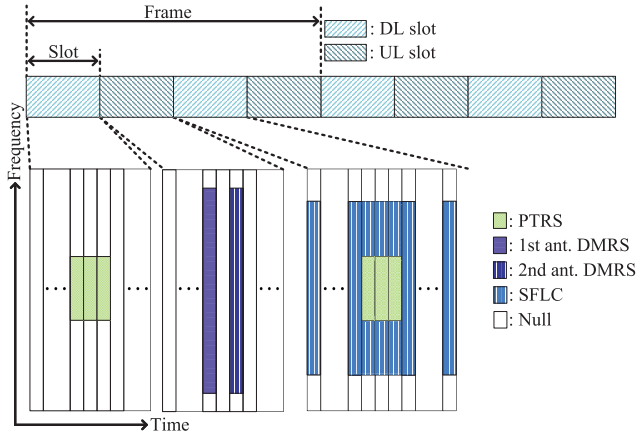


FIGURE 1. Frame structure for a TDD-based $M \times 2$ SFLC-OFDM system using the DMRS-PR.

are provided to evaluate the performance of the proposed PR method. Finally, Section V concludes the paper.

Notations: Scalars, vectors, matrices, and sets are denoted by lower-case, boldface lower-case, boldface capital and calligraphic capital letters, respectively (e.g., x , $\mathbf{x}$, $\mathbf{X}$, and $\mathcal{X}$); superscript T , H , and $*$ denote transposition, Hermitian transposition, and complex conjugate, respectively, for any scalar, vector, and matrix; $|x|$ and $\angle x$ represent the absolute value and phase of a complex number x ; $\lfloor x \rfloor$ returns the largest integer number less than or equal to a real number x ; and $\mathcal{CN}(0, \sigma^2)$ represents a zero-mean complex Gaussian distribution with variance σ^2 .

II. SIGNAL AND SYSTEM MODELS

This section describes the signal and system models for TDD-based $M \times 2$ SFLC-OFDM systems with M Tx antennas and two Rx antennas, considering the phase non-reciprocity between multiple RF ports. To maintain wireless channel reciprocity required for TDD SFLC, low-mobility scenarios are considered. In such environments, established synchronization techniques [21], [34] provide sufficient time and frequency alignment. Consequently, perfect time and frequency synchronization is assumed in the following signal and system models.

A. TDD-BASED $M \times 2$ SFLC-OFDM SYSTEM

Fig. 1 illustrates the frame structure for a TDD-based $M \times 2$ SFLC-OFDM system using the DMRS-PR method. It is worth mentioning that the proposed DMRS-PR method is applicable to general TDD MIMO-OFDM systems with PLL-induced POs, ensuring broad applicability beyond the specific frame structure shown here. A frame is composed of four time slots, each slot including 14 OFDM symbols. The DL and UL time slots are alternately allocated for the TDD operation. In the first DL slot, the BS transmits only PTRS sequences because CSI is not yet available at the BS. From the second DL slot onward, along with PTRS, the SFLC symbols encoded using the CSI estimated in the UL DMRS are transmitted. PTRS sequences are

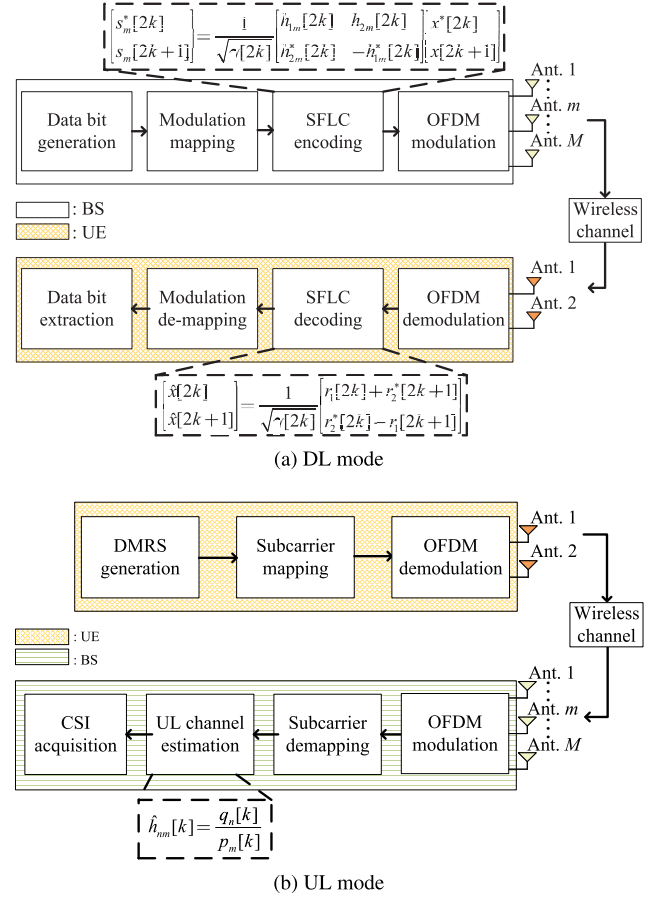


FIGURE 2. Block diagram for an $M \times 2$ SFLC-OFDM system.

assigned to M DL subcarriers and P contiguous OFDM symbols using MP resource elements (REs) per DL slot. Each BS antenna transmits a unique PTRS sequence, which is orthogonal to the other PTRS sequences transmitted through the other antennas. The first and second UE antennas transmit DMRS sequences in UL slots on the sixth and eighth OFDM symbols. The BS estimates the UL channels using the received DMRS sequences. REs excluding the DMRS can be allocated for UL data transmission. For example, UE can exploit the space-frequency block code for UL data transmission as the dual counterpart to DL SFLC [1].

Fig. 2 presents the block diagram for the DL and UL operations in the $M \times 2$ SFLC-OFDM system. From the second DL slot, the BS performs SFLC encoding by pairing two adjacent subcarriers for linear precoding using the CSI. The SFLC symbols transmitted by the m th BS antenna on each subcarrier pair are denoted as [20]

$$\begin{bmatrix} s_m^*[2k] \\ s_m^*[2k+1] \end{bmatrix} = \frac{1}{\sqrt{\gamma[2k]}} \begin{bmatrix} h_{1m}[2k] & h_{2m}[2k] \\ h_{2m}^*[2k] & -h_{1m}^*[2k] \end{bmatrix} \begin{bmatrix} x^*[2k] \\ x[2k+1] \end{bmatrix}, \quad (1)$$

where $x[k]$ represents a modulated information symbol allocated to the k th subcarrier, $h_{nm}[k]$ is the frequency-domain channel coefficient between the n th UE and m th BS antenna pair, and $\gamma[k]$ is the unit-power normalizer for the k th

subcarrier given by

$$\gamma[k] = \sum_{n=1}^2 \sum_{m=1}^M |h_{nm}[k]|^2. \quad (2)$$

The SFLC-encoded symbols and the PTRS sequences are mapped to the time-frequency REs in a DL slot. The BS performs OFDM modulation, inserts a cyclic prefix (CP), and transmits the time-domain signals through M antennas.

The Rician fading channel is assumed in which $h_{nm}[k]$ is characterized by the superposition of a dominant line-of-sight (LOS) component and non-LOS scattered multipath components as follows [35]:

$$h_{nm}[k] = \sqrt{\frac{K}{K+1}} e^{j\alpha_{nm}} + \sqrt{\frac{1}{K+1}} g_{nm}[k], \quad (3)$$

where K means the Rician K -factor, $e^{j\alpha_{nm}}$ is the LOS component with a constant phase, and $g_{nm}[k]$ is the non-LOS component represented as a zero-mean complex Gaussian random variable with unit variance. It is assumed that the channel gains are identical for adjacent subcarriers, i.e., $h_{nm}[2k] = h_{nm}[2k+1]$.

At the receiver, the UE removes the CP, conducts OFDM demodulation, and obtains the frequency-domain received symbols as below:

$$r_n[k] = \sum_{m=1}^M h_{nm}[k] s_m[k] + z_n[k], \quad n \in \{1, 2\}, \quad (4)$$

where $z_n[k]$ is an additive white Gaussian noise (AWGN) at the n th UE antenna, distributed with $\mathcal{CN}(0, \sigma_z^2)$. For the SFLC decoding, $r_1[k]$ and $r_2[k]$ are combined according to a fixed linear processing as follows [20]:

$$\begin{aligned} \begin{bmatrix} y[2k] \\ y[2k+1] \end{bmatrix} &= \begin{bmatrix} r_1[2k] + r_2^*[2k+1] \\ r_2^*[2k] - r_1[2k+1] \end{bmatrix} \\ &= \sqrt{\gamma[2k]} \begin{bmatrix} x[2k] \\ x[2k+1] \end{bmatrix} + \begin{bmatrix} z_1[2k] + z_2^*[2k+1] \\ z_2^*[2k] - z_1[2k+1] \end{bmatrix}. \end{aligned} \quad (5)$$

In (5), the UE detects $x[2k]$ and $x[2k+1]$ from $y[2k]$ and $y[2k+1]$.

In UL slots, the UE allocates DMRS sequences to the time-frequency REs for each antenna to estimate the UL channels corresponding to all pairs of Tx-Rx antennas. Suppose that the allocation of UL DMRS and data is completed. In that case, the UE generates a time-domain signal through OFDM modulation and CP insertion, which is then transmitted to the BS over the wireless channel.

After the removal of CP and OFDM demodulation, BS extracts the frequency-domain DMRS sequences via RE demapping. At the k th subcarrier, the received DMRS sequences associated with the n th UE and m th BS antenna pair are expressed as

$$q_{nm}[k] = h_{nm}[k] p_n[k] + z_m[k], \quad k \in \mathcal{D}_n, \quad (6)$$

where $p_n[k]$ is the DMRS sequence for the n th UE antenna, $z_m[k]$ is an AWGN at the m th BS antenna, distributed with

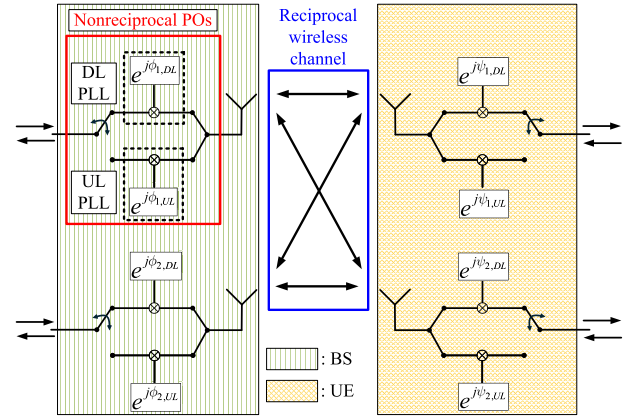


FIGURE 3. DL and UL channels with nonreciprocal POs in a 2×2 SFLC-OFDM system.

$\mathcal{CN}(0, \sigma_z^2)$, and $\mathcal{D}_n$ is the subcarrier index set denoting the DMRS allocation for the n th UE antenna, given by

$$\mathcal{D}_n = \{k_{n,DMRS} + Ld \mid d = 0, 1, \dots, D-1\}. \quad (7)$$

Here, $k_{n,DMRS}$ is the first DMRS subcarrier index for the n th UE antenna, L represents the spacing between two adjacent DMRS sequences, and D is the length of the DMRS sequence. For the pairing of subcarriers in the proposed DMRS-PR method, D is assumed to be an even number.

From (6), the BS estimates the frequency-domain channel coefficients using the received DMRS sequences as follows [21]:

$$\hat{h}_{nm}[k] = \frac{q_{nm}[k]}{p_n[k]}, \quad k \in \mathcal{D}_n. \quad (8)$$

Moreover, linear interpolation¹ over k is subsequently applied to estimate $h_{nm}[k]$ for $k \notin \mathcal{D}_n$. Once the CSI is estimated, as explained previously, the DL data can be transferred through SFLC encoding and decoding as shown in (1)–(5).

B. EFFECT OF PHASE NONRECIPROCITY IN TDD-BASED SFLC-OFDM

Fig. 3 illustrates the nonreciprocal POs in the DL and UL channels for a 2×2 antenna configuration. Independent operations of PLLs in the Tx and Rx modes induce phase discrepancies between the DL and UL channels [26]. Taking into account these discrepancies, we express the DL and UL channels between the n th UE and the m th BS antenna pair as

$$h_{nm,DL}[k] = h_{nm}[k] e^{j\omega_{nm,DL}}, \quad (9a)$$

$$h_{nm,UL}[k] = h_{nm}[k] e^{j\omega_{nm,UL}}, \quad (9b)$$

where $\omega_{nm,DL}$ and $\omega_{nm,UL}$ are the effective POs of DL and UL channels, denoted as

$$\omega_{nm,DL} = \psi_{n,DL} + \phi_{m,DL}, \quad (10a)$$

$$\omega_{nm,UL} = \psi_{n,UL} + \phi_{m,UL}. \quad (10b)$$

¹The proposed phase nonreciprocity correction method in Section III is independent of the channel estimator. It can be combined with advanced channel estimation techniques [36], [37] to enhance overall performance.

Here, $\psi_{n,DL}$ and $\psi_{n,UL}$ are the POs at the n th UE antenna in the DL and UL modes, and $\phi_{m,DL}$ and $\phi_{m,UL}$ are the POs at the m th BS antenna in the DL and UL modes. We substitute $h_{nm,DL}[k]$ and $h_{nm,UL}[k]$ in $h_{nm}[k]$ in (4) and (6), respectively, and neglect the noise terms for simplicity. Then, the UL channels with POs are perfectly estimated, and the SFLC-decoded symbols under nonreciprocal POs are given in (11), as shown at the bottom of the page, where the channel with distortion $A[k]$ is expressed as

$$A[k] = \frac{1}{\sqrt{\gamma[k]}} \sum_{m=1}^M \left(|h_{1m}[k]|^2 e^{j(\omega_{1m,DL} - \omega_{1m,UL})} + |h_{2m}[k]|^2 e^{-j(\omega_{2m,DL} - \omega_{2m,UL})} \right) \quad (12)$$

and the interference term $B[k]$ is written as

$$B[k] = \frac{1}{\sqrt{\gamma[k]}} \sum_{m=1}^M h_{1m}[k] h_{2m}^*[k] \times \left(e^{j(\omega_{1m,DL} - \omega_{2m,UL})} - e^{-j(\omega_{2m,DL} - \omega_{1m,UL})} \right). \quad (13)$$

Nonreciprocal POs evidently introduce distortion and interference, as shown in (11). Consequently, the PLL-induced POs significantly degrade symbol detection reliability. This degradation in symbol-level performance directly impacts system throughput because frequent decoding errors reduce the number of successfully transmitted bits. In addition, elevated error rates often lead to increased packet-level retransmissions, raising overall system latency. Therefore, uncompensated PLL-induced POs undermine the ability of TDD SFLC-OFDM systems to support high-throughput and low-latency communication.

III. PROPOSED DMRS-PR METHOD FOR PHASE RECIPROCITY RECOVERY

Since the PO-induced distortion in (12) and interference terms in (13) exhibit frequency selectivity, PO compensation

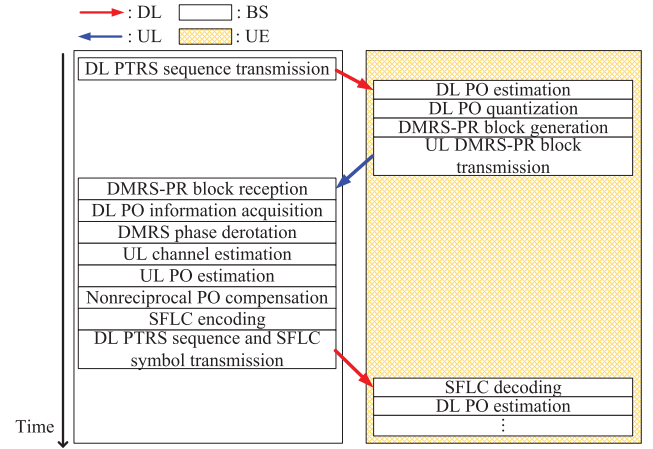


FIGURE 4. Flow diagram of the proposed DMRS-PR method for phase nonreciprocity compensation.

on the UE side requires DL DMRS transmissions that span the entire channel bandwidth. This paper proposes the DMRS-PR method that enables the BS to estimate and correct the PO discrepancies utilizing UL DMRS sequences as an alternative approach. In other words, PO compensation is conducted on the BS side. Fig. 4 presents the overall flow diagram of the proposed DMRS-PR method. The BS transmits the PTRS sequences in the DL slot, and the UE estimates the PO for each pair of Tx-Rx antennas. The quantized PO information is conveyed to the BS through UL DMRS sequences with PR. Subsequently, the BS jointly estimates the embedded POs and the UL channels. Finally, the BS performs SFLC encoding using the CSI with nonreciprocal PO compensation.

In a DL slot, the PTRS sequences are allocated to M subcarriers over P contiguous OFDM symbols for DL PO estimation at the UE. Without loss of generality, it is assumed that the subcarrier index of the PTRS sequences remains fixed across the OFDM symbols. The received PTRS symbol from

$$\begin{aligned} y[2k] &= r_1[2k] + r_2^*[2k + 1] \\ &= \frac{1}{\sqrt{\gamma[2k]}} \sum_{m=1}^M \left(|h_{1m}[2k]|^2 e^{j(\omega_{1m,DL} - \omega_{1m,UL})} + |h_{2m}[2k]|^2 e^{-j(\omega_{2m,DL} - \omega_{2m,UL})} \right) x[2k] \\ &\quad + \frac{1}{\sqrt{\gamma[2k]}} \sum_{m=1}^M h_{1m}[2k] h_{2m}^*[2k] \left(e^{j(\omega_{1m,DL} - \omega_{2m,UL})} - e^{-j(\omega_{2m,DL} - \omega_{1m,UL})} \right) x^*[2k + 1] \\ &= A[2k]x[2k] + B[2k]x^*[2k + 1], \end{aligned} \quad (11a)$$

$$\begin{aligned} y[2k + 1] &= r_2^*[2k] - r_1[2k + 1] \\ &= \frac{1}{\sqrt{\gamma[2k]}} \sum_{m=1}^M \left(|h_{1m}[2k]|^2 e^{j(\omega_{1m,DL} - \omega_{1m,UL})} + |h_{2m}[2k]|^2 e^{-j(\omega_{2m,DL} - \omega_{2m,UL})} \right) x[2k + 1] \\ &\quad - \frac{1}{\sqrt{\gamma[2k]}} \sum_{m=1}^M h_{1m}[2k] h_{2m}^*[2k] \left(e^{j(\omega_{1m,DL} - \omega_{2m,UL})} - e^{-j(\omega_{2m,DL} - \omega_{1m,UL})} \right) x^*[2k] \\ &= A[2k]x[2k + 1] - B[2k]x^*[2k]. \end{aligned} \quad (11b)$$

the m th BS antenna to the n th UE antenna is then expressed as

$$u_{nm}[k_m, PTRS] = h_{nm,DL}[k_m, PTRS]v_m[k_m, PTRS] + z_n[k_m, PTRS], \quad (14)$$

where $k_m, PTRS$ is the PTRS subcarrier index and $v_m[k]$ represents the PTRS sequence at the m th BS antenna. Using $u_{nm}[k_m, PTRS]$, the UE obtains an initial estimate related to the DL PO as follows:

$$\hat{\omega}_{nm,DL} = \angle \left(\frac{u_{nm}[k_m, PTRS]}{v_m[k_m, PTRS]} \right). \quad (15)$$

Here, $\hat{\omega}_{nm,DL}$ is given by an estimate of $\omega_{nm,DL} + \angle h_{nm}[k_m, PTRS]$. The channel phase component $\angle h_{nm}[k_m, PTRS]$ is removed by the BS, as described in detail later in this section. By averaging $\hat{\omega}_{nm,DL}$ across the P OFDM symbols containing PTRS sequences, the UE estimates the DL POs.

The UE applies the PR to the DMRS sequences using $\hat{\omega}_{nm,DL}$ to convey the DL PO information to the BS. Once the UE estimates the DL PO, $\hat{\omega}_{nm,DL}$ is quantized with Q bits using the 2^Q -PSK constellation and Gray coding, as illustrated in Fig. 5. The resulting binary sequences are concatenated and then modulated using binary PSK (BPSK), resulting in a PR vector $\theta \in \mathbb{Z}^{2MQ \times 1}$. Each element of θ represents a PR of π or 0. The UE alternately allocates a phase-rotated DMRS and phase-unrotated original DMRS sequences, creating a DMRS-PR block whose length is $4MQ$, as shown in Fig. 6. The DMRS-PR block can be transmitted repeatedly to enhance the accuracy of DL PO detection in the BS. When the number of repetitive transmissions is B , the DMRS sequence length satisfies $D \geq 4QMB$. Thus, the following condition holds

$$B \leq \left\lfloor \frac{D-1}{4QM} \right\rfloor. \quad (16)$$

The phase-rotated DMRS sequences are expressed as

$$\bar{p}_n[k] = \begin{cases} p_n[k]\bar{\theta}[l], & \text{if } k = k_{n,DMRS} + (2l+1)L, \\ p_n[k], & \text{if } k = k_{n,DMRS} + 2lL, \end{cases} \quad (17)$$

where $l \in \{0, \dots, 2QMB-1\}$ denotes the PR index and $\bar{\theta}[l]$ is the l th entry of the repetitive PR vector $\bar{\theta} \in \mathbb{Z}^{2QMB \times 1}$ obtained by stacking θ for B repetitions. After allocating the DMRS-PR block for repetitive transmission, the UE performs OFDM modulation and CP insertion to generate the time-domain signals containing the DMRS-PR blocks.

The BS extracts the DL PO information from the received DMRS-PR blocks prior to the UL channel estimation. The phase difference between two adjacent DMRS subcarriers depends on the PR at the UE and the phase component of the channels. If the LOS component is sufficiently strong, the channel phase difference is negligible between two adjacent DMRS subcarriers. Using this fact, the BS computes the average phase difference between two adjacent DMRS

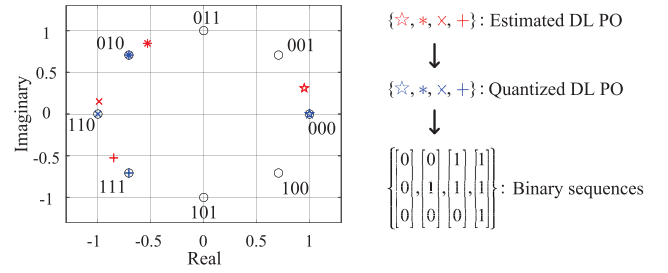


FIGURE 5. DL PO quantization in a 2×2 SFCL-OFDM system when $Q = 3$.

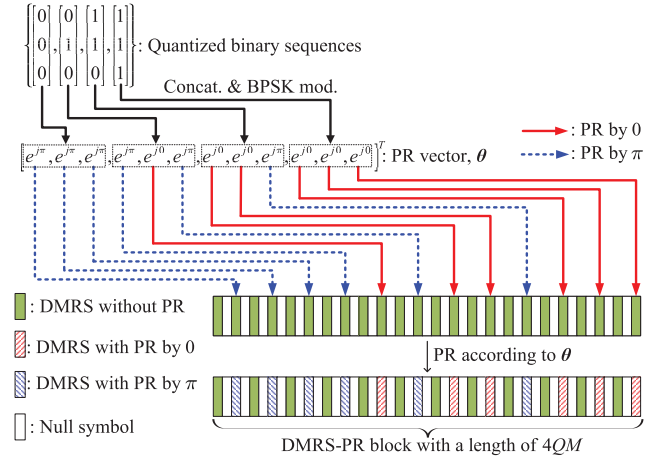


FIGURE 6. DMRS-PR block for a 2×2 SFCL-OFDM system when $Q = 3$.

subcarriers to obtain the DL PO information as follows:

$$\tilde{\theta}[l] = \frac{1}{2MB} \sum_{n=1}^2 \sum_{m=1}^M \frac{\bar{q}_{nm}[k_{n,DMRS} + (2l+1)L]}{\bar{q}_{nm}[k_{n,DMRS} + 2lL]}, \quad (18)$$

where $\bar{q}_{nm}[k]$ means the received DMRS-PR symbol associated with the n th UE and m th BS antenna pair at the k th subcarrier. The estimated PR $\hat{\theta}[l]$ is obtained by BPSK demodulation using $\tilde{\theta}[l]$, and the estimated PR vector $\hat{\theta} \in \mathbb{Z}^{2QM \times 1}$ is then formed by stacking $\hat{\theta}[l]$ in columnwise. Subsequently, $\hat{\theta}$ is reshaped into a matrix with dimensions $Q \times 2M$, resulting in the quantized DL PO matrix $\hat{\Theta} \in \mathbb{Z}^{Q \times 2M}$ where the nm th column of $\hat{\Theta}$ corresponds to the quantized binary sequence of $\hat{\omega}_{nm,DL}$. The BS detects $\hat{\omega}_{nm,DL}$ by modulating the columns of $\hat{\Theta}$ with the 2^Q -PSK symbols.

As mentioned in (15), $\hat{\omega}_{nm,DL}$ is an estimate of $\omega_{nm,DL} + \angle h_{nm}[k_m, PTRS]$. Thus, canceling $\angle h_{nm}[k_m, PTRS]$ from $\hat{\omega}_{nm,DL}$ is essential to compensate for only DL PO $\omega_{nm,DL}$. However, it is challenging to directly estimate the channel phase component $\angle h_{nm}[k_m, PTRS]$. To avoid estimating $\angle h_{nm}[k_m, PTRS]$, the BS estimates the difference between DL and UL POs, $\Omega_{nm} = \omega_{nm,DL} - \omega_{nm,UL}$, using the received DMRS sequences. Initially, the BS removes PR from the received DMRS sequences using $\hat{\theta}$. The BS then calculates the UL channel estimates, including the effect of $\omega_{nm,UL}$, denoted as $\hat{h}_{nm,UL}[k]$, based on the procedure provided in Section II. From $\hat{h}_{nm,UL}[k]$, an initial estimate of UL POs is obtained at the BS as below:

$$\hat{\omega}_{nm,UL} = \angle \hat{h}_{nm,UL}[k_m, PTRS]. \quad (19)$$

Here, $\hat{\omega}_{nm,UL}$ is an estimate of $\omega_{nm,UL} + \angle h_{nm}[k_{m,PTRS}]$. Since $\hat{\omega}_{nm,DL}$ is the estimate of $\omega_{nm,DL} + \angle h_{nm}[k_{m,PTRS}]$ in (15), the BS can estimate Ω_{nm} by subtracting $\hat{\omega}_{nm,UL}$ from $\hat{\omega}_{nm,DL}$, i.e.,

$$\hat{\Omega}_{nm} = \hat{\omega}_{nm,DL} - \hat{\omega}_{nm,UL}. \quad (20)$$

Using $\hat{\Omega}_{nm}$, the BS obtains the DL channel estimates with nonreciprocal PO compensation as follows:

$$\hat{h}_{nm,DL}[k] = \hat{h}_{nm,UL}[k]e^{j\hat{\Omega}_{nm}}. \quad (21)$$

When all estimations are perfect (i.e., $\hat{\Omega}_{nm} = \Omega_{nm}$ and $\hat{h}_{nm,UL}[k] = h_{nm,UL}[k]$), $\hat{h}_{nm,DL}[k]$ yields the effective DL channel coefficient $h_{nm,DL}[k]$, ensuring that the phase difference between the UL and DL channels is completely removed. In the subsequent DL slot, the BS performs SFLC encoding with $\hat{h}_{nm,DL}[k]$. The proposed DMRS-PR method is summarized in Algorithm 1.

Algorithm 1 Proposed DMRS-PR Method

- 1: **Procedure in UE:**
- 2: Obtain $\hat{\omega}_{nm,DL}$ as in (15).
- 3: Compute θ by quantizing $\{\hat{\omega}_{nm,DL}\}$.
- 4: Compute $\hat{\theta}$ by stacking θ for B repetitions.
- 5: Perform PR to DMRS sequences with $\hat{\theta}$ as in (17).
- 6: Transmit phase-rotated DMRS to the BS.
- 7: **Procedure in BS:**
- 8: Compute $\tilde{\theta}[l]$ as in (18).
- 9: Obtain $\hat{\theta}[l]$ via BPSK demodulation.
- 10: Form $\hat{\theta}$ by stacking $\hat{\theta}[l]$ columnwise.
- 11: Compute $\hat{\Theta}$ by reshaping $\hat{\theta}$ into $Q \times 2M$.
- 12: Detect $\hat{\omega}_{nm,DL}$ from the nm th column of $\hat{\Theta}$.
- 13: Remove the PR from the received DMRS using $\hat{\theta}$.
- 14: Compute $\hat{\omega}_{nm,UL}$ as in (19).
- 15: Obtain $\hat{h}_{nm,UL}[k]$ according to (8).
- 16: Compute $\hat{\Omega}_{nm}$ with (20).
- 17: Calculate $\hat{h}_{nm,DL}[k]$ with (21).

Remark 1: In frequency selective channels that violate the strong LOS assumption, the inter-subcarrier phase variation corrupts $\hat{\omega}_{nm,DL}$ in the DMRS-PR method, limiting its PO compensation capability. As an alternative, DL PO feedback can be used for PLL-induced PO compensation. In the feedback approach, the UE transmits quantized DL PO estimates to the BS over the UL channel. The BS with M antennas employs receive diversity techniques, such as maximum ratio combining [38], allowing accurate acquisition of DL PO.

Remark 2: The DL PO feedback incurs an additional $2QM$ -bit UL overhead, which practical systems can accommodate. Low-mobility settings, such as indoor environments, are well suited to SFLC because channel reciprocity can be maintained within the DL-UL interval. For typical sub-6 GHz indoor deployments, M is equal to or less than eight [39]. As confirmed in Fig. 7(a) of Section IV, $Q = 6$ suffices for accurate PLL-induced PO compensation, resulting in $2QM \leq 96$. 5G NR allows UE to multiplex feedback and data on

TABLE 3. Simulation parameters.

Parameters	Values
Number of subcarriers	1024
Subcarrier spacing	15 kHz
Number of OFDM symbols per slot	14
CP length	80 (1st symbol), 72 (others)
Carrier frequency	3.5 GHz
Information modulation	Quadrature PSK
DMRS length (D)	312
DMRS subcarrier spacing (L)	2
1st DMRS subcarrier index ($k_{n,DMRS}$)	-312 ($n = 1$), -311 ($n = 2$)
Number of BS antennas (M)	2
PTRS subcarrier index ($k_{m,PTRS}$)	2 ($m = 1$), 4 ($m = 2$)
Channel model	ITU-R Pedestrian A
Doppler spectrum	Jakes' model
Maximum Doppler shift	10 Hz
Phase of LOS components (α_{nm})	Uniform distribution over $(-\pi, \pi)$
DL and UL POs	Uniform distribution over $(-\pi, \pi)$

the physical UL shared channel (PUSCH), limiting overhead under 20% of PUSCH resources [40]. Because a physical resource block (PRB) comprises 168 resource elements for 15 kHz subcarrier spacing, the DL PO feedback requires one PRB when BPSK is used. In small indoor cells, UEs are typically assigned 8–16 PUSCH PRBs [41]. Therefore, a 96-bit feedback occupies only 4.5–9% of the PUSCH resources, which makes this overhead easily accommodated. Furthermore, PLL-induced PO remains constant over time [26]. The acquired DL PO information can be reused in subsequent communications, as verified in Fig. 10 of Section IV, reducing overall UL overhead.

IV. SIMULATION RESULTS

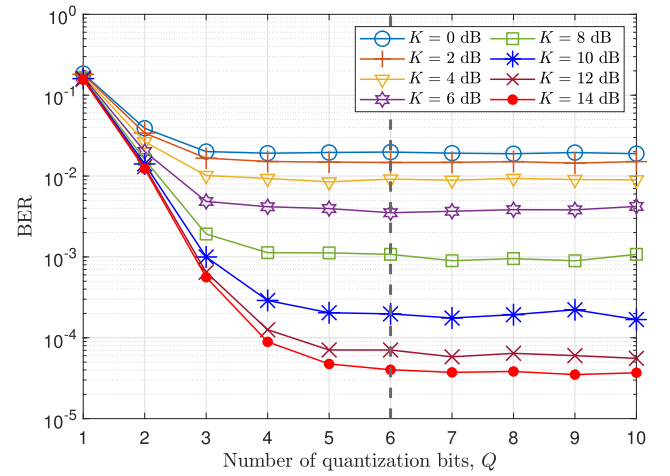
To verify the effectiveness of the proposed DMRS-PR method, we evaluated the performance of the bit error rate (BER) in the simulation environments summarized in Table 3. The number of subcarriers is fixed at 1024, corresponding to the subcarrier index range from -512 to 511 . The subcarrier spacing is 15 kHz, resulting in a channel bandwidth of 15.36 MHz. Each frame spans 4 ms and comprises four time slots with 1 ms each. Each time slot consists of 14 OFDM symbols. The length of the CP is 80 samples ($5.2 \mu s$) for the first OFDM symbol and 72 samples ($4.7 \mu s$) for the other OFDM symbols, following the configuration CP for a subcarrier spacing 15 kHz in the 5G NR standards. The carrier frequency is set to 3.5 GHz. Information symbols are modulated using the quadrature PSK. The DMRS length D and the DMRS subcarrier spacing L are set to 312 and 2, respectively. The first DMRS subcarrier indices $k_{n,DMRS}$ are -312 and -311 for $n = 1$ and $n = 2$, respectively. The number of BS antennas is two (i.e., $M = 2$). The PTRS subcarrier indices $k_{m,PTRS}$ are 2 and 4 for the first and second BS antennas, respectively. To ensure relevance to practical deployments, the parameter values are chosen based on the 5G NR specification [42].

The ITU-R pedestrian A channel model in [43] is adopted to realize a multipath fading channel. The Doppler spectrum follows Jakes's model with a ring of scatterers [44]. The maximum Doppler shift is 10 Hz, corresponding to the

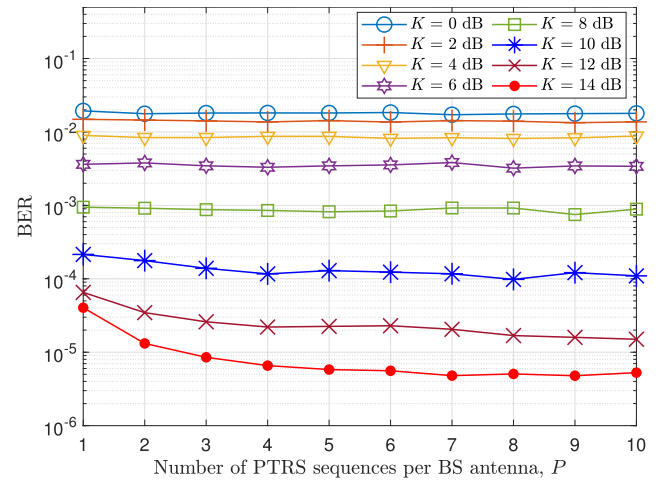
maximum relative velocity of 0.83 km/h between the BS and UE. The phase of LOS components, α_{nm} , is uniformly distributed over $(-\pi, \pi)$. Similarly, DL and UL POs are random variables with a uniform distribution on $(-\pi, \pi)$.

Fig. 7 presents the BER performance of the proposed DMRS-PR method on three critical system parameters: the number of quantization bits Q , the number of PTRS sequences per BS antenna P , and the number of DMRS-PR blocks B . These results illustrate how each parameter influences the overall BER performance of the proposed method. The simulation results indicate that Q has the most significant effect on BER performance, while the impact of P and B is limited. As Fig. 7(a) demonstrates, increase in Q improves BER performance, with performance saturation occurring at $Q = 6$. Increasing Q , the quantization error of DL PO is reduced, which in turn decreases the residual phase discrepancies after combining the SFLC signals. At $Q = 6$, the quantization error in the DL PO representation is negligible compared to the channel estimation error and AWGN. Consequently, no additional improvement in BER is observed for $Q \geq 6$. For P , only modest performance gains are observed when $K \geq 12$ dB, while for $K \leq 10$ dB, increasing P has a negligible effect, as shown in Fig. 7(b). Because the required value P is small, the proposed DMRS-PR method imposes only minimal DL overhead. In contrast, Fig. 7(c) shows that increasing B does not produce any improvement in BER regardless of K . The simulation results for P and B can be explained by considering the roles of P and B in the DMRS-PR method. Specifically, P is related to the accuracy of the DL PO estimation at the UE during DL communication, while B governs the accuracy of the DL PO acquisition at the BS in the UL communication. However, due to uncertainties in the estimation of the UL channel, the performance gains of increasing P and B remain limited, even when the estimation of DL PO in the UE and the acquisition in the BS are accurate.

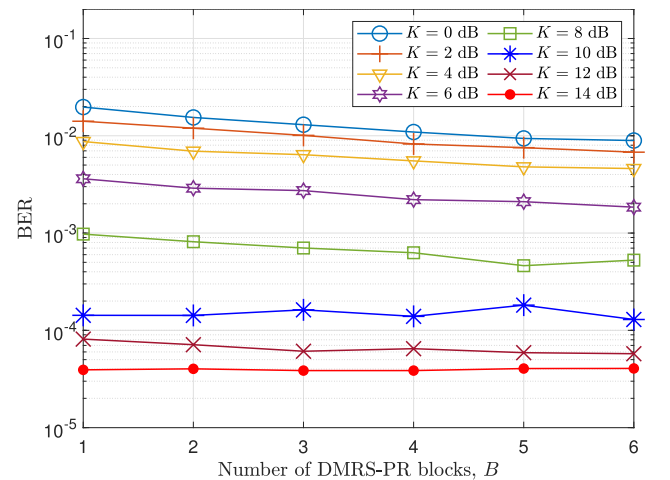
Fig. 8 compares the performance of the BER in K under four conditions: (i) no PLL-induced PO compensation, (ii) the DMRS-PR method with $Q = 6$, (iii) the DL PO feedback method with Q values from three to seven, and (iv) an ideal case without PLL-induced PO. In the DL PO feedback method, the UE and the BS use BPSK modulation and the maximum ratio combining for feedback transmission and detection, respectively. The BER performance for no PLL-induced PO is a benchmark where only the estimation of the UL channel is required. As shown, the SFLC-OFDM system cannot achieve acceptable BER without PLL-induced PO compensation. Specifically, BER remains nearly 0.5 in all K and SNR values, highlighting the essential role of PO compensation in TDD MIMO SFLC-OFDM systems. In contrast, both DMRS-PR and DL PO feedback methods effectively mitigate DL-UL phase discrepancies, yielding substantial BER performance improvements. Notably, these two methods show distinct behaviors as K increases. The DMRS-PR method shows a sharp improvement in BER at $K = 12$ dB. In comparison, the DL PO feedback



(a) BER across Q when $P = 1$ and $B = 1$



(b) BER across P when $Q = 6$ and $B = 1$



(c) BER across B when $Q = 6$ and $P = 1$

FIGURE 7. BER performance of DMRS-PR method with respect to Q , P , and B when SNR = 14 dB and QPSK is used.

approach maintains a nearly constant gap to the case without PLL-induced PO in all K when $Q \geq 5$, demonstrating its robustness under weak-LOS conditions.

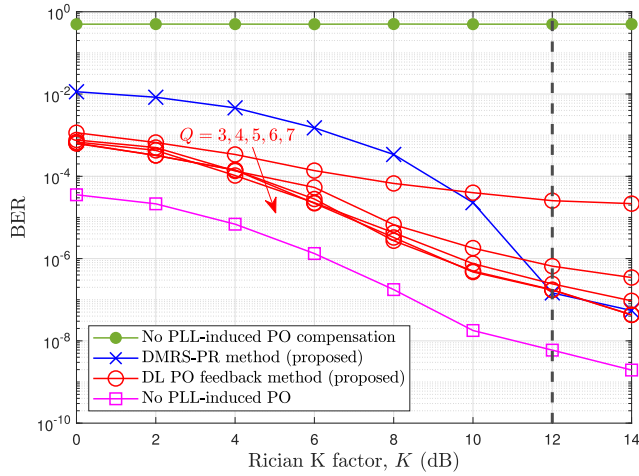


FIGURE 8. BER performance comparison across K when $\text{SNR} = 16$ dB, $P = 1$, and $B = 1$.

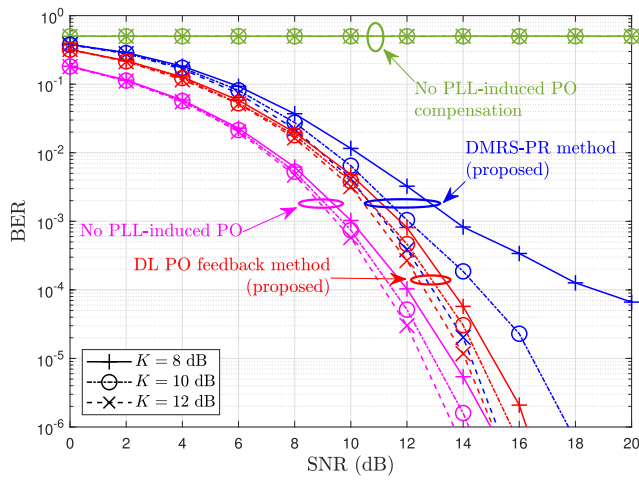
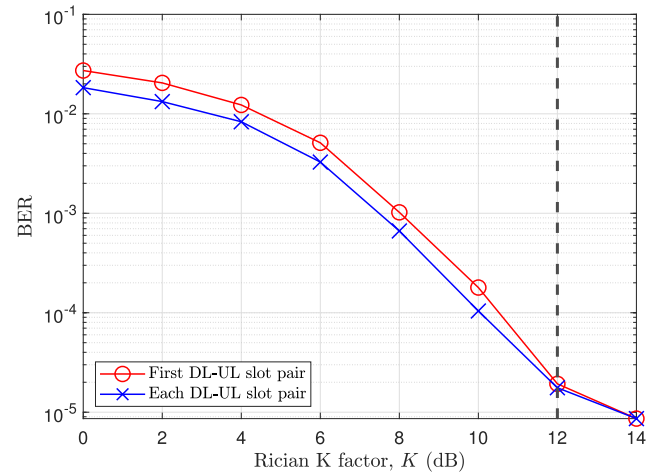


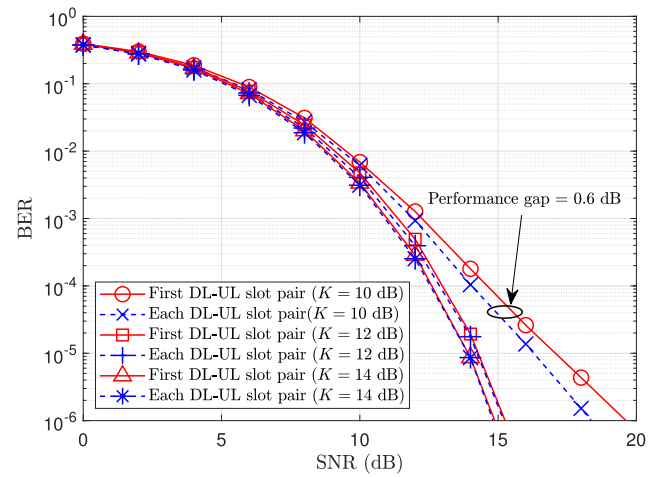
FIGURE 9. BER performance comparison across SNR when $Q = 6$, $P = 1$, and $B = 1$.

To further elaborate the BER performance evaluation, Fig. 9 presents the BER performance over SNR values with $K = 8, 10$, and 12 dB. For $K = 12$ dB, both the DMRS-PR and the DL PO feedback methods achieve the diversity order of the case without PLL-induced PO. However, only the DL PO feedback method reaches the same diversity order as in the case without PLL-induced PO when $K = 8$. These results confirm the robustness of the feedback approach in violation of the strong LOS assumption, while the DMRS-PR method requires $K \geq 12$ dB.

PLL-induced PO remains constant over time [26], allowing reuse of PO estimates obtained during initial DL and UL communications for subsequent transmissions of DL SFLC symbols. To validate this approach, Fig. 10(a) compares the BER performance in two scenarios: (i) reuse PO estimates from the first DL and UL slots for all subsequent transmissions and (ii) perform PO estimation for each pair of DL-UL slots. The simulation results are obtained with 200 transmitted frames per channel realization, $Q = 6$, $P = 1$, and $B = 1$. Fig. 10(a) plots BER across K with a fixed



(a) BER across K when $\text{SNR} = 14$ dB



(b) BER across SNR for $K = 10, 12$, and 14 dB

FIGURE 10. BER comparison between PO estimation using only the first DL-UL slot pair and PO estimation for each DL-UL slot pair when the number of frames is 200, $Q = 6$, $P = 1$, and $B = 1$.

SNR of 14 dB. The results show that both scenarios achieve nearly identical performance for $K \geq 12$ dB, indicating that a sufficiently strong LOS component enables the reuse of estimated POs with negligible performance degradation compared to reestimating POs in every DL-UL slot pair. For $K \leq 10$ dB, a modest performance gap emerges due to the weaker LOS components and the increased influence of multipath effects. However, this gap remains small, demonstrating that reusing the initial PO estimate remains viable even under low-LOS conditions. To further assess the consistency of this trend across different SNR values, Fig. 10(b) presents BER as a function of SNR for representative K values of 10, 12, and 14 dB. The results confirm that reusing the initial PO estimate achieves BER performance comparable to estimating POs in every DL-UL slot pair across practical SNR ranges. Therefore, applying PO estimates from the first DL and UL slots provides an effective strategy to mitigate DL-UL channel phase discrepancies while remarkably reducing system complexity and overhead.

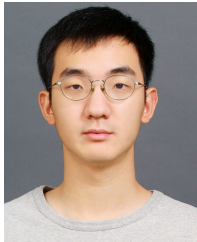
V. CONCLUSION

This work investigates the impact of phase nonreciprocity in SFLC-OFDM systems. It proposes the DMRS-PR method that embeds the DL PO information into UL DMRS sequences to compensate for phase nonreciprocity and ensure reliable system performance. The efficacy of the DMRS-PR scheme is validated through extensive numerical simulations. However, the applicability of the DMRS-PR method is based on the presence of a strong LOS component to address phase nonreciprocity, limiting its use in frequency-selective channels. The DL PO feedback scheme serves as an alternative, trading a modest UL overhead for robust BER performance.

REFERENCES

- [1] J. Joung, "Space-time line code," *IEEE Access*, vol. 6, pp. 1023–1041, 2018.
- [2] J. Joung, "Space-time line code for massive MIMO and multiuser systems with antenna allocation," *IEEE Access*, vol. 6, pp. 962–979, 2018.
- [3] J. Joung and E.-R. Jeong, "Multiuser space-time line code with optimal and suboptimal power allocation methods," *IEEE Access*, vol. 6, pp. 51766–51775, 2018.
- [4] J. Joung and J. Choi, "Multiuser space-time line code with transmit antenna selection," *IEEE Access*, vol. 8, pp. 71930–71939, 2020.
- [5] S. Han, J. Joung, and J. Choi, "Interference exploitation precoding based on multiple space-time line codes for MU-MIMO systems," *IEEE Wireless Commun. Lett.*, vol. 14, no. 2, pp. 410–414, Feb. 2025.
- [6] Y. Kim, S. Han, J. Joung, J. Kim, J. Zhao, and J. Choi, "Enhanced multiuser space-time line code for downlink multiple antenna transmission," *IEEE Trans. Commun.*, early access, 19, 2024, doi: 10.1109/TCOMM.2024.3502460.
- [7] J. Joung, J. Choi, and B. C. Jung, "Double space-time line codes," *IEEE Trans. Veh. Technol.*, vol. 69, no. 2, pp. 2316–2321, Feb. 2020.
- [8] J. Choi and J. Joung, "Generalized space-time line code with receive combining for MIMO systems," *IEEE Syst. J.*, vol. 16, no. 2, pp. 1897–1908, Jun. 2022.
- [9] J. Joung and J. Fan, "Linear precoding for space-time line code-based multicast systems," *IEEE Trans. Veh. Technol.*, vol. 71, no. 11, pp. 12386–12391, Nov. 2022.
- [10] J. Joung, "Energy efficient space-time line coded regenerative two-way relay under per-antenna power constraints," *IEEE Access*, vol. 6, pp. 47026–47035, 2018.
- [11] J. Joung and J. Choi, "Space-time line codes with power allocation for regenerative two-way relay systems," *IEEE Trans. Veh. Technol.*, vol. 68, no. 5, pp. 4884–4893, May 2019.
- [12] J. Joung, B. C. Jung, and J. Choi, "Space-time line coded regenerative two-way relay systems with power control," *IET Electron. Lett.*, vol. 55, no. 12, pp. 694–696, Jun. 2019.
- [13] J. Kim, J. Joung, and K. Lim, "Intelligent reflecting surface-aided space-time line coded systems," *IEEE Wireless Commun. Lett.*, vol. 11, no. 2, pp. 245–249, Feb. 2022.
- [14] J. Kim, J. Choi, J. Joung, and Y.-C. Liang, "Modified block coordinate descent method for intelligent reflecting surface-aided space-time line coded systems," *IEEE Wireless Commun. Lett.*, vol. 11, no. 9, pp. 1820–1824, Sep. 2022.
- [15] J. Choi, J. Joung, and B. C. Jung, "Space-time line code for enhancing physical layer security of multiuser MIMO uplink transmission," *IEEE Syst. J.*, vol. 15, no. 3, pp. 3336–3347, Sep. 2021.
- [16] J. Choi, J. Joung, and Y.-S. Cho, "Artificial-noise-aided space-time line code for enhancing physical layer security of multiuser MIMO downlink transmission," *IEEE Syst. J.*, vol. 16, no. 1, pp. 1289–1300, Mar. 2022.
- [17] J. Joung and C. Yuen, "Space-time line code hopping for physical-layer secure communications," *IEEE Trans. Inf. Forensics Security*, vol. 19, pp. 7720–7729, 2024.
- [18] J.-B. Seo, H. Jin, J. Joung, and B. C. Jung, "Uplink NOMA random access systems with space-time line code," *IEEE Trans. Veh. Technol.*, vol. 69, no. 4, pp. 4522–4526, Apr. 2020.
- [19] K.-H. Lee, J. S. Yeom, J. Joung, and B. C. Jung, "Performance analysis of uplink NOMA with constellation-rotated STLC for IoT networks," *IEEE Open J. Commun. Soc.*, vol. 3, pp. 705–717, 2022.
- [20] C. Wu, Y. Xiao, Y. L. Guan, J. Wang, X. Li, and P. Yang, "Space-time/frequency line coded OFDM: System design and practical implementation," *IEEE Access*, vol. 7, pp. 151915–151928, 2019.
- [21] S. Kim, H.-G. Lee, S. Lee, J. Kim, J. Joung, and J. Kim, "Single-user SFLC-OFDM system realization based on channel reciprocity recovery," *IEEE Access*, vol. 11, pp. 109082–109094, 2023.
- [22] G. S. Smith, "A direct derivation of a single-antenna reciprocity relation for the time domain," *IEEE Trans. Antennas Propag.*, vol. 52, no. 6, pp. 1568–1577, Jun. 2004.
- [23] M. Petermann, M. Stefer, F. Ludwig, D. Wubben, M. Schneider, S. Paul, and K.-D. Kammeyer, "Multi-user pre-processing in multi-antenna OFDM TDD systems with non-reciprocal transceivers," *IEEE Trans. Commun.*, vol. 61, no. 9, pp. 3781–3793, Sep. 2013.
- [24] J. Hershberger, T. Pratt, and R. Kossler, "A software-defined, dual-polarized radar system," in *Proc. IEEE Conf. Antenna Meas. Appl. (CAMA)*, Syracuse, NY, USA, Oct. 2016, pp. 1–4.
- [25] J. Hershberger, T. Pratt, and R. Kossler, "Implementations of coherent software-defined dual-polarized radars," *IEEE Trans. Microw. Theory Techn.*, vol. 65, no. 5, pp. 1673–1681, May 2017.
- [26] G. Xu and E. C. Kan, "Phase offset calibration in multi-channel radio-frequency transceivers," *IEEE J. Microw.*, vol. 4, no. 1, pp. 111–122, Jan. 2024.
- [27] D. Petrovic, W. Rave, and G. Fettweis, "Common phase error due to phase noise in OFDM-estimation and suppression," in *Proc. IEEE Int. Symp. Pers. Indoor Mobile Radio Commun. (PIMRC)*, Barcelona, Spain, Mar. 2005, pp. 1901–1905.
- [28] K. Nikitopoulos and A. Polydoros, "Phase-impairment effects and compensation algorithms for OFDM systems," *IEEE Trans. Commun.*, vol. 53, no. 4, pp. 698–707, Apr. 2005.
- [29] S. Negusse, P. Zetterberg, and P. Händel, "Phase-noise mitigation in OFDM by best match trajectories," *IEEE Trans. Commun.*, vol. 63, no. 5, pp. 1712–1725, May 2015.
- [30] D. Dexu Lin, R. A. Pacheco, T. Joon Lim, and D. Hatzinakos, "Joint estimation of channel response, frequency offset, and phase noise in OFDM," *IEEE Trans. Signal Process.*, vol. 54, no. 9, pp. 3542–3554, Sep. 2006.
- [31] Y. Kim, J. Lee, and S.-C. Kim, "Joint common phase error and channel estimation for OFDM-based WLANs in the presence of Wiener phase noise and residual frequency offset," in *Proc. IEEE Int. Conf. Commun. (ICC)*, Istanbul, Turkey, Jan. 2006, pp. 3040–3045.
- [32] P. Liu, S. Wu, and Y. Bar-Ness, "A phase noise mitigation scheme for MIMO WLANs with spatially correlated and imperfectly estimated channels," *IEEE Commun. Lett.*, vol. 10, no. 3, pp. 141–143, Mar. 2006.
- [33] Z. Wang, P. Babu, and D. P. Palomar, "Effective low-complexity optimization methods for joint phase noise and channel estimation in OFDM," *IEEE Trans. Signal Process.*, vol. 65, no. 12, pp. 3247–3260, Jun. 2017.
- [34] J. Joung, E.-R. Jeong, and H. Oh, "Blind block synchronization with carrier frequency offset estimation for space-time line coded systems," *ICT-Exp.*, vol. 9, pp. 235–240, Apr. 2022.
- [35] P. Tan and N. C. Beaulieu, "Effect of channel estimation error on bit error probability in OFDM systems over Rayleigh and Ricean fading channels," *IEEE Trans. Commun.*, vol. 56, no. 4, pp. 675–685, Apr. 2008.
- [36] K.-C. Hung and D. W. Lin, "Pilot-aided multicarrier channel estimation via MMSE linear phase-shifted polynomial interpolation," *IEEE Trans. Wireless Commun.*, vol. 9, no. 8, pp. 2539–2549, Aug. 2010.
- [37] G. Auer and E. Karipidis, "Pilot aided channel estimation for OFDM: A separated approach for smoothing and interpolation," in *Proc. IEEE Int. Conf. Commun.*, Seoul, South Korea, May 2005, pp. 2173–2178.
- [38] W. M. Gifford, M. Z. Win, and M. Chiani, "Diversity with practical channel estimation," *IEEE Trans. Wireless Commun.*, vol. 4, no. 4, pp. 1935–1947, Jul. 2005.
- [39] S. I. Siddiqui, S. Bashir, A. Khan, S. Ghafoor, and I. Aziz, "A dual-band high-gain beam steering antenna array for 5G sub-6 GHz base station," *Sci. Rep.*, vol. 14, no. 1, Nov. 2024, Art. no. 26517.
- [40] 3GPP, "NR: Physical layer procedures for control," 3rd Gener. Partnership Project, Sophia Antipolis, Alpes-Maritimes, Tech. Rep. TS 38.213, May 2022.
- [41] A. Kapovits, A. Gavras, J. Cosmas, M. Ghoraihi, X. Li, and Y. Zhang, "Delivery of 5G services indoors—The wireless wire challenge and solutions," 5G Public Private Partnership, Brussels, Belgium, Tech. Rep., Sep. 2021.

- [42] 3GPP, “NR; physical channels and modulation,” 3rd Gener. Partnership Project, Sophia Antipolis, Alpes-Maritimes, France, Tech. Rep. TS 38.211, May 2024.
- [43] *Guidelines for Evaluation of Radio Transmission Technologies for IMT-2000*, Standard ITU-R M.1225, Int. Telecommun. Union, Geneva, Switzerland, 1997. [Online]. Available: <https://www.itu.int/oth/R0A0E00000C/en>
- [44] T. L. Fulghum, K. J. Molnar, and A. Duel-Hallen, “The Jakes fading model for antenna arrays incorporating azimuth spread,” *IEEE Trans. Veh. Technol.*, vol. 51, no. 5, pp. 968–977, Sep. 2002.



HAN-GYEOL LEE (Student Member, IEEE) received the B.S. and M.S. degrees in electrical and electronics engineering from Chung-Ang University (CAU), Seoul, South Korea, in 2021 and 2023, respectively, where he is currently pursuing the Ph.D. degree with the Department of Electrical and Electronics Engineering. His research interests include space-time coding, next-generation multicarrier systems, WiFi sensing, and hardware-impaired-aware signal processing.

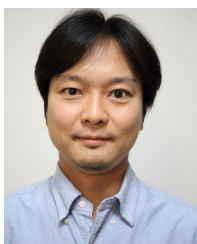


JIHOON CHOI (Senior Member, IEEE) received the B.S., M.S., and Ph.D. degrees from Korea Advanced Institute of Science and Technology (KAIST), Daejeon, South Korea, in 1997, 1999, and 2003, respectively.

From 2003 to 2004, he was with the Department of Electrical and Computer Engineering, The University of Texas at Austin, where he performed research on multiple antenna systems as a Postdoctoral Fellow. From 2004 to 2008, he was with Samsung Electronics, South Korea, where he worked on developments of commercial radio access stations for M-WiMAX and base stations for CDMA 1xEV-DO Rev.A/B. In 2008, he joined KAU, Goyang, South Korea, as a Faculty Member. He is currently a Professor with the School of Electronics and Information Engineering, KAU, where he is also the Chief Investigator of ISPL. His research interests include signal processing for wireless communications, radar signal processing, air-to-ground channel modeling, UAV trajectory optimization, machine learning, and modem design for future cellular networks, wireless LANs, the IoT devices, and digital broadcasting systems.



JUYEOP KIM (Member, IEEE) received the B.S. and Ph.D. degrees in electrical engineering from Korea Advanced Institute of Science and Technology (KAIST) in 2004 and 2010, respectively. He is currently an Assistant Professor with the Department of Electronics Engineering, Sookmyung Women's University, Seoul, South Korea. His current research interests include software modems and next-generation wireless communication.



KOICHI ADACHI (Senior Member, IEEE) received the B.E., M.E., and Ph.D. degrees in engineering from Keio University, Japan, in 2005, 2007, and 2009, respectively.

From 2007 to 2010, he was a Japan Society for the Promotion of Science (JSPS) Research Fellow. He was a Visiting Researcher with the City University of Hong Kong, in April 2009. He was a Visiting Research Fellow with the University of Kent, from June 2009 to August 2009. From May 2010 to May 2016, he was with the Institute for Infocomm Research, A*STAR, in Singapore. From May 2016 to March 2025, he was with The University of Electro-Communications, Japan. Since April 2025, he has been

an Associate Professor at Keio University. His research interests include cooperative communications and energy efficient communication technologies. He is a member of IEICE. He was recognized as an Exemplary Reviewer from IEEE WIRELESS COMMUNICATIONS LETTERS, in 2012, 2013, 2014, and 2015. He received the Excellent Editor Award from IEEE ComSoc MMTS, in 2013. He is the co-author of WPMC2020 Best Student Paper Award and the ICUFN 2022 Best Paper Award. He served as the General Co-Chair for the Tenth and 11th IEEE Vehicular Technology Society Asia-Pacific Wireless Communications Symposium (APWCS), the Track Co-Chair for the Transmission Technologies and Communication Theory of the 78th and 80th IEEE Vehicular Technology Conference, in 2013 and 2014, respectively, the Symposium Co-Chair for the Communication Theory Symposium of IEEE Globecom 2018 and the Wireless Communications Symposium of IEEE Globecom 2020, and the Tutorial Co-Chair for IEEE ICC 2019. He was an Associate Editor of *IET Transactions on Communications* from 2015 to 2017, IEEE WIRELESS COMMUNICATIONS LETTERS, since 2016, IEEE TRANSACTIONS ON VEHICULAR TECHNOLOGY, from 2016 to 2022, IEEE TRANSACTIONS ON GREEN COMMUNICATIONS AND NETWORKING, from 2016 to 2021, IEEE OPEN JOURNAL OF VEHICULAR TECHNOLOGY, since 2019, IEEE TRANSACTIONS ON WIRELESS COMMUNICATIONS, since 2023, and IEEE COMMUNICATIONS SURVEYS AND TUTORIALS, since 2023.



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 serves as a Senior Associate Editor for IEEE COMMUNICATIONS LETTERS and an Associate Editor of IEEE OPEN JOURNAL OF THE COMMUNICATIONS SOCIETY.



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, 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. He is currently a Full Professor with the School of Electrical and Electronics Engineering, CAU, where he is also the Principal Investigator of the Intelligent Wireless Systems (IWS) Laboratory and the Next-Generation Intelligent Sensing and Communication (NISC) Laboratory. His research interests include signal processing, numerical analysis, and algorithms.

Dr. Joung received the First Prize of the Intel-ITRC Student Paper Contest, in 2006. He served as a Guest Editor for IEEE ACCESS, in 2016, *Electronics*, and *Sensors*, in 2019. He served on the Editorial Board of APSIPA Transactions on Signal and Information Processing, from 2014 to 2019. He also served as an Associate Editor for IEEE TRANSACTIONS ON VEHICULAR TECHNOLOGY (Top Editor 2021), from 2018 to 2023. He is the inventor of a space-time line code (STLC), a fully symmetric scheme to a space-time block code. He was recognized as an Exemplary Reviewer of IEEE COMMUNICATIONS LETTERS, in 2012, and IEEE WIRELESS COMMUNICATIONS LETTERS, from 2012 to 2014 and in 2019.

...