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RESEARCH ARTICLE

Data-Aided Channel-Gain Calibration Algorithm for Recovering Channel Amplitude Reciprocity in SFLC-OFDM Systems

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ABSTRACT This paper proposes a practical gain calibration algorithm that estimates and compensates for channel amplitude asymmetry in a space-frequency line code (SFLC) orthogonal frequency division multiplexing (OFDM) system. During uplink (UL) communications from the base station (BS) to user equipment (UE), the SFLC BS can acquire channel state information (CSI) for downlink (DL) communications from the BS to the UE by utilizing UL-DL channel reciprocity. However, the channel amplitudes between UL and DL are generally observed to be asymmetric in practical channels. This asymmetry results in an unexpected phase rotation of the decoded SFLC symbols at the UE, as the CSI used for SFLC encoding at the BS does not align with the actual DL channels. To address this issue, we have designed a data-aided channel-gain calibration algorithm (CCA). Based on the mathematical derivation of the phase rotation of received SFLC data symbols at the UE, we developed the CCA to adjust the received gain and restore channel amplitude symmetry. To practically validate the designed CCA, we implemented an off-the-shelf SFLC-OFDM communication system on a software modem testbed using Universal Software Radio Peripherals (USRPs). Experiments conducted with real-time SFLC-OFDM signals demonstrate that the proposed CCA significantly enhances SFLC decoding performance.

INDEX TERMS Channel reciprocity, channel state information (CSI), channel-gain calibration, space-frequency line code (SFLC) and software modem.

I. INTRODUCTION

Mobile communications have advanced in the spatial domain using multiple input and multiple output (MIMO) technologies [1], [2]. To enhance the degrees of freedom in the spatial domain, recent MIMO technologies have evolved to multiplex the data of multiple users using massive antenna arrays [3], [4]. Most massive MIMO technologies with sophisticated precoding require channel state information at the transmitter (CSIT). The challenge of CSIT realization is conventionally addressed by utilizing feedback of the

estimated channel at the receiver [5], which inherently involves quantization errors, increasing the error vector magnitude (EVM) of transmitted signals. Additionally, pilot overhead is a critical issue in implementing a massive MIMO system as the number of pilot transmissions increases proportionally with the number of transmit antennas [6], [7]. This extensive use of pilot signals reduces available resources for data transmission.

To make massive MIMO practical in real-world applications, space-time line code (STLC) can be considered an effective solution to address the challenge of pilot overhead [8]. Like space-time block code (STBC), STLC fully leverages time and antenna diversity using acquired

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channel state information (CSI). On the other hand, STLC encodes symbols based on CSIT and does not utilize CSI during the decoding process at the receivers. This characteristic reduces the pilot overhead in systems with a massive number of transmit antennas, as the number of pilots required for estimating CSI at the transmitter does not depend on the number of transmit antennas.

Further studies have proposed advanced STLC schemes to enhance system throughput and spectral efficiency by increasing the number of data streams [9], [10], or the number of users [11], [12], [13]. The STLC has also been applied in vehicular communication scenarios [14], [15], physical layer security [16], [17], [18], and two-way relay communications [19], [20], [21]. Additionally, space–frequency line code (SFLC) has been proposed to integrate the STLC with orthogonal frequency division multiplexing (OFDM) [22].

Recent research in massive MIMO has utilized channel reciprocity for CSIT acquisition in time division duplexing (TDD) systems [23], [24], [25]. Since TDD systems use the same frequency band for bidirectional communications, the downlink (DL) and uplink (UL) wireless channels are expected to exhibit symmetric characteristics. The TDD system consequently allows for the estimation of a DL channel using a sounding signal received through a UL channel [26]. This CSIT acquisition is beneficial for massive MIMO technologies, as it does not require CSI feedback, which often leads to estimation error issues. Furthermore, CSIT acquisition utilizing channel-symmetric characteristics incurs less pilot overhead. The number of pilots in sounding signals for this CSIT acquisition is proportional to the number of UL transmit antennas, while remaining independent of the number of massive DL transmit antennas.

Several conventional studies demonstrated in Table. 1 have also highlighted the significance of achieving channel reciprocity in real-world systems. Basic calibration approach requires additional hardware circuits or amplifiers [27], [28]. The relative calibration methods which do not require extra hardware has been proposed for single cell environments [29], [30], and further enhanced to achieve channel reciprocity in presence of frequency offset [31]. On the other hand, the relative calibration in massive MIMO systems challenges the feedback overhead and the OTA self-calibration based on the Argos prototype has been proposed for less overhead [32]. This method has the drawback of its reliance on a reference antenna causing inherent high-sensitivity to antenna placement. In addition, a unified framework that comprehensively explains existing calibration methods is proposed [33] and LS-based calibration is leveraged to enhance the calibration accuracy [34], [35].

Conventional studies additionally shed lights on MIMO channel estimation by jointly addressing antenna calibration. These studies consider the integration of antenna calibration techniques into various antenna structures within MIMO systems. Basic calibration methods have been proposed with beamforming precoding algorithms [36], [37]. Some methods

calculate the coefficients for each base station antenna in various MIMO scenarios [38], [39], [40], [41]. Channel reciprocity is also analyzed when applying calibration methods to specific communication systems [42], [43], [44]. Advanced calibration methods address strong-coupled channels among base station antennas [45], [46] and guarantee robustness to multipath and time delay [47]. In addition, calibration methods are proposed to consider multipath channel environments [48], [49], [50]. However, these conventional methods also face the feedback overhead and computational load when applied to massive MIMO systems with multiple antennas at the BS.

Its practical application for channel estimation in OFDM systems remains challenging. The DL and UL channels experienced by the frequency-domain subcarriers are sensitive to the synchronization status of receivers and may not be reciprocal if the receivers are synchronized differently at the baseband level. A conventional study addresses asymmetric channel phase variations based on different states of time and frequency synchronization [51]. It proposes and implements an enhanced synchronization scheme, demonstrating the recovery of channel phase reciprocity in practical environments. However, it does not address channel amplitude asymmetry, which commonly arises from discrepancies in receiver gains within radio frequency (RF) chains [27]. This discrepancy leads to CSI errors during the MIMO precoding process and increases the EVM of decoded symbols.

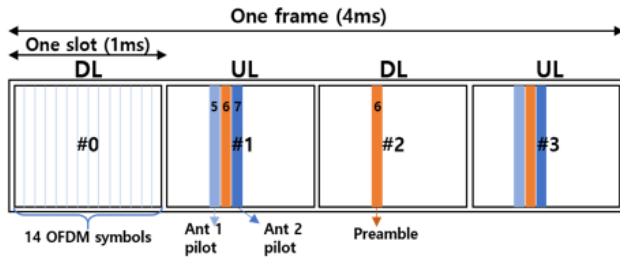
In this paper, we propose an algorithm to recover channel amplitude reciprocity and enhance the accuracy of SFLC precoding based on CSIT. The key contribution of this paper can be summarized as follows:

- An in-depth mathematical analysis of how channel amplitude reciprocity affects the performance of SFLC decoding is presented.
- Based on the analysis, a novel method for estimating amplitude imbalance is proposed, effectively achieving precise channel amplitude reciprocity.
- A novel data-aided channel-gain calibration algorithm (CCA) is proposed that reduces the complexity burden on the receiver by eliminating the need for separate reference signals and circuits.
- The proposed schemes have been implemented on a software modem testbed for practical assessment. They have been confirmed to resolve the amplitude imbalance issue in signal emission environments effectively.

The remainder of this paper is organized as follows: Section II introduces the system model considered in this study. In Section III, we analyze the mathematical offset that causes phase rotation of STLC symbols under channel amplitude asymmetry and propose a method for estimation and channel-gain calibration. Section IV describes the tested and experimental environments and presents the experimental results. Finally, Section V concludes the paper.

TABLE 1. Comparison with conventional related studies.

References	Descriptions	Limitations
[27], [28]	calibration with hardware	requiring additional circuits and amplifiers
[29], [30]	relative channel reciprocity calibration for single cell system	huge feedback overhead in massive MIMO systems
[32]	Argos calibration	sensitive to the relative placement of the reference antenna
[38]–[41]	estimating mismatch of calibration coefficients in MIMO scenarios	huge feedback overhead and computational complexity arise
[45]–[50]	calibration in coupling and multipath channel scenarios	huge computational complexity in massive MIMO systems
[51]	real-time synchronization for channel reciprocity in TDD	guaranteeing channel reciprocity only in a phase perspective

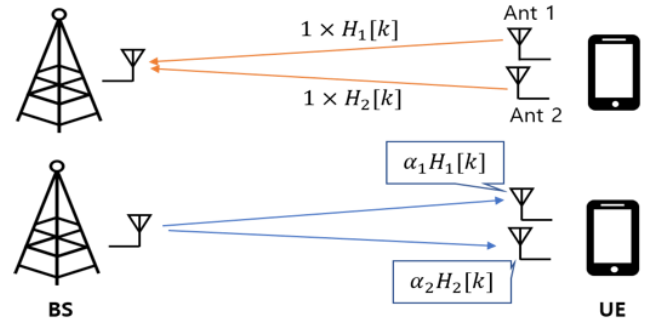
**FIGURE 1.** The 1×2 SFLC-OFDM system model.**FIGURE 2.** TDD frame structure of the SFLC-OFDM system.

II. SYSTEM MODEL WITH CHANNEL AMPLITUDE ASYMMETRY

As illustrated in Fig. 1, we consider a 1×2 SFLC-OFDM system with one antenna at the base station (BS) and two antennas at the user equipment (UE), which supports bidirectional signal transmission between the BS and UE.¹ The UE transmits pilot signals to acquire CSIT, while both the BS and UE transmit preamble signals for synchronization.

To realize CSIT, the system uses a TDD frame structure, as shown in Fig. 2. The main parts of the TDD frame structure are designed by modifying 5G New Radio (NR). Each frame includes four slots comprised of two UL and two DL slots, respectively, alternating in a one-to-one TDD pattern. Each slot has a duration of 1 ms and consists of 14 OFDM symbols. The BS transmits preamble symbols in the sixth OFDM symbol of every DL slot for DL synchronization at the UE. The UE transmits channel sounding blocks (CSBs) during the fifth, sixth, and seventh OFDM symbols of each UL slot to

¹For notational convenience, only one antenna is used at the BS. It is straightforward to extend the SFLC-OFDM and the proposed CCA to a BS with multiple transmit antennas [8].

**FIGURE 3.** Channel model under amplitude and phase asymmetry between UL and DL paths.

enable CSI estimation at the BS. Both the BS and UE employ an m-sequence of length 127 for the preamble symbols.

The CSB specifically consists of preamble symbols for UL synchronization and pilot symbols for channel estimation. The preamble symbols are allocated to the center subcarriers of the sixth OFDM symbol and transmitted only through the first antenna of the UE. The pilot symbols for the first and second antennas are allocated to the fifth and seventh OFDM symbols, respectively. The pilot symbols are identical to the symbol sequence used for 5G demodulated reference signals (DMRS).

The BS estimates the UL channel based on the pilot signals of the CSB and utilizes it as the DL channel to encode for the next DL slot. In this case, UL and DL channels can be asymmetry due to mismatches between the transmit and receive RF paths connected to each antenna. Thus, the UL and DL channels may exhibit disparities in their common amplitude and phase terms. Fig. 3 illustrates our channel model, highlighting the amplitude and phase asymmetry between the UL and DL antenna paths. Let $\alpha_i \in \mathbb{C}$ denote the ratio of channel gain between the UL and DL for the i th antenna, where $\mathbb{C}$ represents the set of complex numbers. If we denote $H_i[k]$ as the UL channel between the BS and the i th UE antenna for the k th subcarrier, the corresponding DL channel can be expressed as $\alpha_i H_i[k]$.

α_i has a key role of realizing channel amplitude asymmetry in this channel model. This is ultimately determined by the

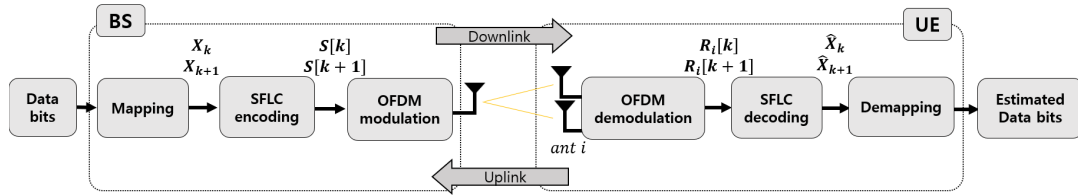


FIGURE 4. Baseband signal processing in the SFLC-OFDM system.

relative proportions of the gains of the Radio Frequency (RF) chains and antennas at transmitter and receiver. To this end, α_i incorporates the overall channel asymmetry between RF chains as well as antennas. The DL and UL channels are asymmetric for the i -th antenna if $\alpha_i \neq 1$. The channels are also asymmetric between UE antennas if $\alpha_1 \neq \alpha_2$.

A. SFLC ENCODING AND DECODING

The system theoretically achieves the maximum diversity order through STLC encoding and decoding when full CSI is available at the BS and no CSI is available at the UE [8].

Based on this fact, our system model employs SFLC, which transmits encoded symbols over two consecutive subcarriers. Fig. 4 illustrates the overall flow of baseband signal processing in this system. X_k and X_{k+1} denote the k th and $(k+1)$ th M-ary phase shift keying (M-PSK) symbols mapped from the input data bit sequence. The SFLC symbols are encoded based on $C_{(1,2)}^a$ of the STLC encoding functions [8]. The encoding function assumes that the coherent bandwidth is larger than subcarrier spacing and the channel gain remains constant for two consecutive subcarriers. Hence, SFLC encoding is performed with the assumption that $H_i[k] \simeq H_i[k+1]$ for $i \in \{1, 2\}$.

$S[k]$ and $S[k+1]$, representing the k th and $(k+1)$ th SFLC-encoded symbols, are mathematically expressed as follows:

$$\begin{bmatrix} S^*[k] \\ S[k+1] \end{bmatrix} = \begin{bmatrix} H_1[k] & H_2[k] \\ H_2^*[k] & -H_1^*[k] \end{bmatrix} \begin{bmatrix} X_k \\ X_{k+1} \end{bmatrix}, \quad (1)$$

where $(*)$ denotes complex conjugation. $S[k]$ is then mapped to the k th subcarrier and converted into a time-domain signal based on OFDM modulation [51].

The time-domain signal is transferred to the each antenna of the UE through the DL channels. The UE conducts OFDM demodulation for each antenna and obtains the received symbols for the k th and $(k+1)$ th subcarriers at antenna 1 and 2. $R_i[k]$, representing the received symbol for the k th subcarrier at i th antenna, is mathematically expressed as follows:

$$\begin{bmatrix} R_1[k] & R_1[k+1] \\ R_2[k] & R_2[k+1] \end{bmatrix} = \begin{bmatrix} H_1[k] \\ H_2[k] \end{bmatrix} \begin{bmatrix} S[k] & S[k+1] \end{bmatrix} + \begin{bmatrix} Z_1[k] & Z_1[k+1] \\ Z_2[k] & Z_2[k+1] \end{bmatrix}, \quad (2)$$

where $Z_i[k]$ is an additive white Gaussian noise (AWGN) with zero mean and variance σ^2 for the k th subcarrier at the i th UE antenna.

The UE recovers the k th and $(k+1)$ th M-PSK symbols from $R_i[k]$ and $R_i[k+1]$ using the SFLC combining provided by $C_{(1,2)}^a$ [8]. The SFLC combining acquires the estimated symbols, denoted as $\hat{X}_k$ and $\hat{X}_{k+1}$, using a fixed linear combination, as illustrated below:

$$\begin{bmatrix} \hat{X}_k \\ \hat{X}_{k+1} \end{bmatrix} = \begin{bmatrix} R_1^*[k] + R_2[k+1] \\ R_2^*[k] - R_1[k+1] \end{bmatrix}. \quad (3)$$

By substituting (1) and (2) into (3), $\hat{X}_k$ and $\hat{X}_{k+1}$ can be expressed as follows:

$$\begin{aligned} \hat{X}_k &= R_1^*[k] + R_2[k+1] \\ &= H_1^* S^*[k] + H_2 S[k+1] + Z_1^*[k] + Z_2[k+1] \\ &= H_1^* (H_1 X_k + H_2 X_{k+1}) + H_2 (H_2^* X_k - H_1^* X_{k+1}) \\ &\quad + Z_1^*[k] + Z_2[k+1] \\ &= (|H_1|^2 + |H_2|^2) X_k + Z_1^*[k] + Z_2[k+1], \end{aligned} \quad (4a)$$

$$\begin{aligned} \hat{X}_{k+1} &= R_2^*[k] - R_1[k+1] \\ &= H_2^* S^*[k] - H_1 S[k+1] + Z_2^*[k] - Z_1[k+1] \\ &= H_2^* (H_1 X_k + H_2 X_{k+1}) - H_1 (H_2^* X_k - H_1^* X_{k+1}) \\ &\quad + Z_2^*[k] - Z_1[k+1] \\ &= (|H_1|^2 + |H_2|^2) X_{k+1} + Z_2^*[k] - Z_1[k+1], \end{aligned} \quad (4b)$$

where the subcarrier index k is dropped in $H_1[k]$ and $H_2[k]$ for simplicity. $\hat{X}_k$ and $\hat{X}_{k+1}$ achieve the full channel diversity gain $|H_1|^2 + |H_2|^2$. The data bits can be decoded from $\hat{X}_k$ and $\hat{X}_{k+1}$ using maximum likelihood (ML) detection. In this case, the signal-to-noise ratio (SNR) is given by

$$\rho_k = \frac{|H_1[k]|^2 + |H_2[k]|^2}{2\sigma^2}. \quad (5)$$

B. SYNCHRONIZATION FOR RECOVERING CHANNEL PHASE ASYMMETRY

As presented in (4), the CSI used for generating $S[k]$ and $S[k+1]$ must match the actual channel experienced by the received symbols $R_i[k]$ and $R_i[k+1]$. Therefore, the accuracy of the CSIT estimation significantly impacts the SFLC decoding performance on the UE side. The BS estimates the CSIT for the k th subcarrier at the i th UE antenna, denoted by $\hat{H}_i[k]$, from the UL channel based on channel reciprocity, and uses it to encode the SFLC symbols. To maintain phase symmetry between DL and UL channels, the BS and UE fully compensate for time and frequency offsets during their synchronization processes [51].

Time and frequency offsets introduce phase rotations and inter-carrier interference, which lead to channel phase asymmetry.

During the synchronization procedures, the BS and UE compensate for symbol time offset (STO) and carrier frequency offset (CFO) using the received preamble signals [51]. Let us denote the transmitted and received preamble signals in the time domain as $p(n)$ and $r_p(n)$, respectively, with lengths N and N_r . The BS and UE acquire the STO $\hat{\delta}$ and the coarse CFO $\hat{\epsilon}$ as follows:

$$\hat{\delta} = \arg \max_{\delta \in \mathbb{N}} \Xi[\delta], \quad (6)$$

$$\hat{\epsilon} = \frac{\langle \Psi_0^*(\hat{\delta}) \Psi_1(\hat{\delta}) \rangle}{\pi}, \quad (7)$$

where $\langle x \rangle$ represents the phase of a complex x , $\mathbb{N} = \{1, 2, \dots, N_r\}$, and the metrics $\Xi[\delta]$ and $\Psi_i(\hat{\delta})$ are given by

$$\Xi[\delta] \triangleq \left| \sum_{n=0}^{N-1} \phi(\delta, n) \right|, \quad (8)$$

$$\Psi_i(\delta) \triangleq \sum_{n=\frac{Nm}{2}}^{\frac{N(m+1)}{2}-1} \phi(\delta, n), \text{ for } m \in \{0, 1\}, \quad (9)$$

and

$$\phi(\delta, n) \triangleq R_p(\delta + n)P^*(n). \quad (10)$$

The estimator in (6) provides the coarse STO as the value of δ that maximizes the correlation function $\Xi[\delta]$. The coarse CFO is estimated from the phase differential of $\Psi_0(\hat{\delta})$ and $\Psi_1(\hat{\delta})$ which represent the sums of the first and second halves of the received preamble signal.

To achieve precise phase symmetry between DL and UL channels, the system addresses the sampling clock offset (SCO), which generates phase rotation in both channels. The SCO is estimated from the received preamble signals and compensated for in the frequency domain at both the BS and UE. For the SCO estimation, the phase rotation ratio, denoted by $\hat{\Delta}\beta$, is obtained as follows:

$$\hat{\Delta}\beta = \angle (\chi_{m-1}^* \chi_m), \quad (11)$$

where χ_m is the average phase of the received preamble symbols in the m th DL or UL slot, given by

$$\chi_m \triangleq \sum_{s=0}^{N_p} R_p[m, s]P^*[m, s]. \quad (12)$$

Here, $P[m, s]$ and $R_p[m, s]$ represent the transmitted and received preamble symbols in the frequency domain, respectively, while N_p denotes the number of preamble symbols. To compensate for the SCO, the BS and UE adjust their center frequency f_c by a step size Δf_c , depending on the sign of $\hat{\Delta}\beta$. This 2-step compensation method ensures synchronization with high precision, and channel symmetry from a phase perspective is rigorously aligned.

III. RECOVERY OF CHANNEL AMPLITUDE ASYMMETRY

Due to the asymmetric channels shown in Fig. 3, the UL channel estimated at the BS can be different from the actual DL channel experienced by the received symbols at the UE. This mismatch generates unexpected phase rotation at the SFLC decoder, resulting in the phase errors of decoded symbols. In the following subsection, the SFLC-decoded symbols in asymmetric channel-gain environments will be expressed as a function of the phase error related to α_i . Then, we derive a new estimator for channel amplitude asymmetry and realize a data-aided CCA.

A. CHANNEL IMBALANCE ESTIMATION IN AN 1×2 SFLC SYSTEM

We analyze the channel imbalance in the 1×2 SFLC system based on the asymmetric channel model. Considering the channel-gain asymmetric ratio α_i , the received symbols in (2) can be modified as follows:

$$\begin{bmatrix} R_1[k] & R_1[k+1] \\ R_2[k] & R_2[k+1] \end{bmatrix} = \begin{bmatrix} \alpha_1 H_1[k] \\ \alpha_2 H_2[k] \end{bmatrix} \begin{bmatrix} S[k] & S[k+1] \end{bmatrix} + \begin{bmatrix} Z_1[k] & Z_1[k+1] \\ Z_2[k] & Z_2[k+1] \end{bmatrix}. \quad (13)$$

The SFLC decoding process produces $\hat{X}_k$ and $\hat{X}_{k+1}$ through the linear combination in (3) as follows:

$$\begin{aligned} \hat{X}_k &= \alpha_1^* H_1^* S^*[k] + \alpha_2 H_2 S[k+1] + Z_1^*[k] + Z_2[k+1] \\ &= (\alpha_1^* |H_1|^2 + \alpha_2 |H_2|^2) X_k + (\alpha_1^* - \alpha_2) H_1^* H_2 X_{k+1} \\ &\quad + Z_1^*[k] + Z_2[k+1], \end{aligned} \quad (14a)$$

$$\begin{aligned} \hat{X}_{k+1} &= \alpha_2^* H_2^* S^*[k] - \alpha_1 H_1 S[k+1] + Z_2^*[k] - Z_1[k+1] \\ &= (\alpha_1 |H_1|^2 + \alpha_2^* |H_2|^2) X_{k+1} - (\alpha_1 - \alpha_2^*) H_1 H_2^* X_k \\ &\quad + Z_2^*[k] - Z_1[k+1]. \end{aligned} \quad (14b)$$

Compared with (4), the results above include an unexpected phase rotation related to α_i . These additional terms become negligible when the DL and UL channels are symmetric, with $\alpha_i = 1$. The results also reveal that the phase rotations of $\hat{X}_k$ and $\hat{X}_{k+1}$ depend on the transmitted symbols X_k and X_{k+1} .

To compensate for the channel amplitude asymmetry, we focus on the common term in (14a) and (14b), with ϕ_α defined as follows:

$$\phi_\alpha = (\alpha_1^* - \alpha_2) H_1^* H_2. \quad (15)$$

α_i generally has a nonzero phase if DL and UL channels are asymmetric. According to the assumption the channel phase asymmetry is resolved by the precise synchronization described in Section II-B [51]. Hence, the phase of α_i approaches to zero, allowing α_i to be regarded as a real value representing the channel amplitude mismatch. From this fact, the common part in (15) can be simplified as follows:

$$\phi_\alpha = (\alpha_1 - \alpha_2) H_1^* H_2. \quad (16)$$

The unexpected phase rotation is more dominant in (14) as ϕ_α increases. When the ratio of channel amplitude asymmetry

is the same for both antennas, i.e. $\alpha_1 = \alpha_2$, the common part is removed, i.e. $\phi_\alpha = 0$. It is remarkable that ϕ_α can be zero even if channel amplitude is asymmetric for two UE antennas. In other words, the phase rotation caused by the imbalance between the DL and UL channels can be mitigated by adjusting one of the antenna gains to ensure the condition as follows:

$$\alpha_1 = \alpha_2. \quad (17)$$

Prior to the compensation, it is required to estimate $\phi_\alpha = \Delta\alpha\theta_H$, where $\Delta\alpha \triangleq \alpha_1 - \alpha_2$ and $\theta_H \triangleq H_1^*H_2$. By further defining that

$$A_\alpha = \alpha_1|H_1|^2 + \alpha_2|H_2|^2, \quad (18)$$

we can simplify (14) to

$$\hat{X}_k = A_\alpha X_k + \Delta\alpha\theta_H X_{k+1} + Z_1^*[k] + Z_2[k+1], \quad (19a)$$

$$\hat{X}_{k+1} = A_\alpha X_{k+1} - \Delta\alpha\theta_H^* X_k + Z_2^*[k] - Z_1[k+1]. \quad (19b)$$

Since the right-hand side of (19) can be expressed in a matrix-vector multiplication form, we can derive the ML estimator for two parameters A_α and $\phi_\alpha = \Delta\alpha\theta_H$ as follows:

$$\begin{bmatrix} \hat{A}_\alpha \\ \hat{\phi}_\alpha \end{bmatrix} = \begin{bmatrix} X_k & X_{k+1} \\ X_{k+1}^* & -X_k^* \end{bmatrix}^{-1} \begin{bmatrix} \hat{X}_k \\ \hat{X}_{k+1}^* \end{bmatrix}. \quad (20)$$

Here, the estimator of ϕ_α is rewritten as

$$\hat{\phi}_\alpha = \frac{X_{k+1}^* \hat{X}_k - X_k^* \hat{X}_{k+1}}{X_{k+1}^2 + X_k^2}. \quad (21)$$

Moreover, the presence of ϕ_α causes unexpected phase rotations in $\hat{X}_k$ and X_{k+1} . Thus, $\hat{\phi}_\alpha$ is referred to as the *phase rotation offset (PRO)* for the remainder of the paper.

We can extend the PRO estimator to data symbols by replacing X_k and X_{k+1} with the hard-decision symbols in (21). When denoting the M-PSK symbols as $X_k = e^{i(\frac{2\pi}{M}(m_k-1))}$, $m_k \in \{1, 2, \dots, M\}$, $\hat{\phi}_\alpha$ in (21) can be rewritten as

$$\hat{\phi}_\alpha = \frac{e^{-i(\frac{2\pi}{M}(\hat{m}_{k+1}-1))} \hat{X}_k - e^{-i(\frac{2\pi}{M}(\hat{m}_k-1))} \hat{X}_{k+1}}{e^{i(\frac{4\pi}{M}(\hat{m}_{k+1}-1))} + e^{i(\frac{4\pi}{M}(\hat{m}_k-1))}}, \quad (22)$$

where $\hat{m}_k$ represents the k th data bit sequence acquired by hard-decision. (22) allows a data-aided PRO estimation using only hard-decision symbols and no additional pilot symbols.

The denominator in (22) is decided by the phase difference of two hard-decision symbols. For QPSK ($M = 4$), the PRO estimator is further simplified to $\pm(\hat{X}_k + \hat{X}_{k+1})/2$ if the phase difference between the two symbols is π . In addition, the PRO estimator in (22) is applicable only when its denominator is nonzero. Hence, the proposed estimator needs to be selectively activated only when $|\hat{m}_k - \hat{m}_{k+1}| \neq \frac{M}{4}$.

B. CHANNEL IMBALANCE ESTIMATION IN A 2×2 SFLC SYSTEM

As the 1×2 SFLC system can be easily extended to 2×2 cases [8], the PRO estimator described in Section III-A can

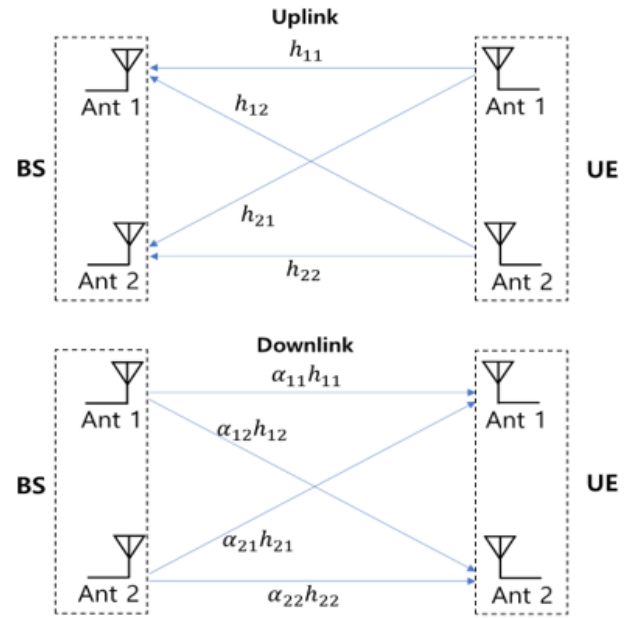


FIGURE 5. The 2×2 SFLC-OFDM channel model under channel asymmetry.

also be extensively applied to 2×2 SFLC system. We provide an analysis of PRO estimation with the assumption of a 2×2 channel asymmetry model demonstrated in Fig. 5. We denote the UL channel gain between the m -th transmit antenna at the BS and the n -th receive antenna at the UE as H_{mn} and the DL channel gain as $\alpha_{mn}H_{mn}$ where α_{mn} is the channel asymmetry factor for the channel. With this model, decoded symbols are mathematically expressed as follows:

$$\begin{aligned} \hat{X}_k &= R_1[k] + R_2^*[k+1] \\ &= \alpha_{11}H_{11}S_1[k] + \alpha_{21}H_{21}S_2[k] \\ &\quad + \alpha_{12}^*H_{12}^*S_1^*[k+1] + \alpha_{22}^*H_{22}^*S_2^*[k+1] \\ &= (\alpha_{11}|H_{11}|^2 + \alpha_{12}^*|H_{12}|^2 + \alpha_{21}|H_{21}|^2 + \alpha_{22}^*|H_{22}|^2)X_k \\ &\quad + (\alpha_{11} - \alpha_{12}^*)H_{11}H_{12}^* + (\alpha_{21} - \alpha_{22}^*)H_{21}H_{22}^*X_{k+1}^*, \end{aligned} \quad (23)$$

$$\begin{aligned} \hat{X}_{k+1} &= R_2^*[k] + R_1[k+1] \\ &= \alpha_{12}^*H_{12}^*S_1^*[k] + \alpha_{22}^*H_{22}^*S_2^*[k] \\ &\quad + \alpha_{11}H_{11}S_1[k+1] + \alpha_{21}H_{21}S_2[k+1] \\ &= (\alpha_{11}|H_{11}|^2 + \alpha_{12}^*|H_{12}|^2 + \alpha_{21}|H_{21}|^2 + \alpha_{22}^*|H_{22}|^2)X_{k+1} \\ &\quad - (\alpha_{11} - \alpha_{12}^*)H_{11}H_{12}^* + (\alpha_{21} - \alpha_{22}^*)H_{21}H_{22}^*X_k^*. \end{aligned} \quad (24)$$

(23) and (24) include the distortion terms similarly presented in (14a) and (14b) in the 1×2 case. For elimination of the distortion terms in X_k and X_{k+1} , both $(\alpha_{11} - \alpha_{12}^*) = 0$ and $(\alpha_{21} - \alpha_{22}^*) = 0$ should be satisfied after a proper calibration. If the precise synchronization described in Section II-B guarantees channel phase symmetry, calibration aims at satisfying channel amplitude symmetry as follows:

$$\alpha_{11} = \alpha_{12} \quad \text{and} \quad \alpha_{21} = \alpha_{22}. \quad (25)$$

Algorithm 1 Pseudo Code for the Proposed Data-Aided CCA

1. Obtain SFLC decoded symbols $\hat{X}_k$ and $\hat{X}_{k+1}$ according to (4).
2. Perform hard decision for $\hat{X}_k$ and $\hat{X}_{k+1}$ and obtain $\hat{m}_k$ and $\hat{m}_{k+1}$.
3. Obtain $\hat{\phi}_\alpha$ based on the PRO estimator in (22).
4. Perform one-tap IIR filtering as follows:

$$\hat{\phi}_{\alpha[n+1]} = a\hat{\phi}_{\alpha[n]} + (1-a)\hat{\phi}_\alpha,$$

where $a \in [0, 1)$ is an IIR filtering constant.

5. Adjust the receive gain for channel-gain calibration as follows:

if $\hat{\phi}_{\alpha n+1} > 0$ **then**

Increase the receive gain of the RF chain for antenna 2.

else if $\hat{\phi}_{\alpha n+1} < 0$ **then**

Decrease the receive gain of the RF chain for antenna 2.

end if

The PRO for the 2×2 case can also be estimated based on the ML method described in III-A. If we define $A_{\alpha mn}$ and $\phi_{\alpha mn}$ as follows:

$$A_{\alpha mn} = \alpha_{11}|H_{11}|^2 + \alpha_{12}^*|H_{12}|^2 + \alpha_{21}|H_{21}|^2 + \alpha_{22}^*|H_{22}|^2 \quad (26)$$

$$\phi_{\alpha mn} = (\alpha_{11} - \alpha_{12}^*)H_{11}H_{12}^* + (\alpha_{21} - \alpha_{22}^*)H_{21}H_{22}^*, \quad (27)$$

The phase rotation of $\hat{X}_k$ and $\hat{X}_{k+1}$ can be estimated through $\phi_{\alpha mn}$. The ML estimator for $\phi_{\alpha mn}$ can be formulated as follows:

$$\begin{bmatrix} \hat{A}_{\alpha mn} \\ \hat{\phi}_{\alpha mn} \end{bmatrix} = \begin{bmatrix} X_k & X_{k+1}^* \\ X_{k+1} & -X_k^* \end{bmatrix}^{-1} \begin{bmatrix} \hat{X}_k \\ \hat{X}_{k+1} \end{bmatrix}. \quad (28)$$

Hence, the estimator of $\phi_{\alpha mn}$ can be derived as follows:

$$\hat{\phi}_{\alpha mn} = \frac{X_{k+1}\hat{X}_k - X_k\hat{X}_{k+1}}{X_{k+1}^2 + X_k^2}, \quad (29)$$

The estimator in (29) reveals the PRO in the 2×2 SFLC system and is in a similar form with (21). This analysis result demonstrates that PRO in the 2×2 SFLC system can be estimated in the same way as the 1×2 SFLC system. Hence, the proposed estimator and calibration method are scalable to the SFLC systems with extended BS antennas.

C. REALIZATION OF DATA-AIDED CHANNEL-GAIN CALIBRATION

The proposed CCA estimates the PRO proportional to the channel amplitude asymmetry and compensate the receive gain to recover the channel amplitude reciprocity. The gain compensation aims at nullifying the PRO based on (17) and (25). The overall channel gain is decomposed into the transmit and receive gains. We denote the transmit gain at m -th antenna and the receive gain at the receive gain at the n -th antenna as T_m and R_n , respectively.

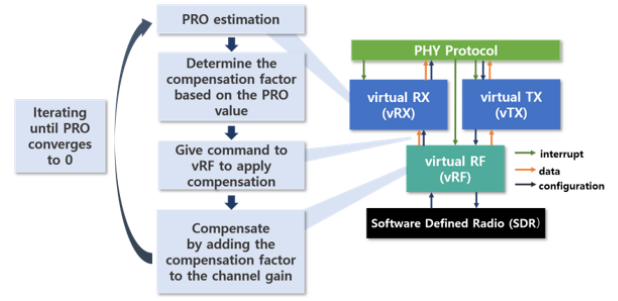


FIGURE 6. Proposed CCA implemented in a software modem.

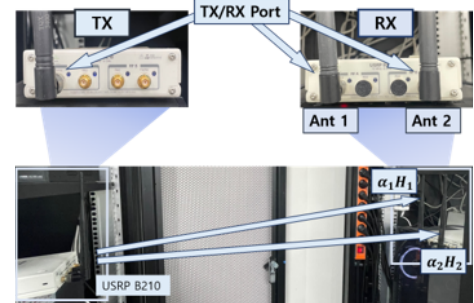


FIGURE 7. Real testbed environment using PC and USRP.

In the 1×2 SFLC system, (17) can be rewritten as follows:

$$T_1 R_1 = T_1 R_2, \quad (30)$$

and (25) can be rewritten as follows:

$$T_1 R_1 = T_1 R_2 \quad \text{and} \quad T_2 R_1 = T_2 R_2. \quad (31)$$

(30) and (31) reveal that receive gains should satisfy that $R_1 = R_2$ in both 1×2 and 2×2 SFLC systems. Based on this observation, the proposed CCA is designed to adjust the receive gains to satisfy $R_1 = R_2$. It is notable that the proposed CCA in the 2×2 SFLC system can compensate gains the same way as in the 1×2 SFLC system.

Fig. 1 illustrates the pseudo-code for the proposed data-aided CCA using the PRO estimator in (22). Using hard-decision symbols, $\hat{\phi}_\alpha$ is computed to estimate $\Delta\alpha$. To mitigate the instantaneous PRO estimation errors due to noise, we apply one-tap infinite impulse response (IIR) filtering to $\hat{\phi}_\alpha$ in (22). Finally, the receive gain in an RF chain is adjusted according to $\hat{\phi}_\alpha$. For example, $\hat{\phi}_\alpha$ is positive when $\alpha_1 > \alpha_2$. Consequently, the UE either increases the receive gain of the second antenna or decreases the receive gain of the first antenna. Conversely, when $\alpha_1 < \alpha_2$, the UE adjusts the receive gains in the opposite manner. Since the gain adjustment can be applied to either of the two UE antennas, we control the receive gain of the second antenna in Algorithm 1.

IV. IMPLEMENTATION AND EXPERIMENTAL RESULTS

For performance evaluation in practical environments, we implemented the proposed CCA on a software modem

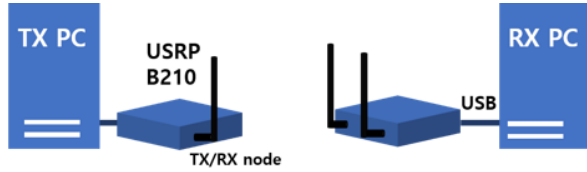


FIGURE 8. Schematic testbed with PC and USRP.

TABLE 2. Experiment parameters.

Communication parameters	Values
Center frequency	3.3 GHz
Bandwidth	10 MHz
Subcarrier spacing	15 kHz
Sampling rate	15.36 MHz
FFT size	1024
Modulation scheme	QPSK
Hardware parameters	Values
SDR	USRP B210
CPU	Core-i9

testbed system [51]. Fig. 6 illustrates the architecture of the software modem and the operational flow of the proposed CCA within it. The software modem is primarily composed of three threads: virtual RX (vRX), virtual TX (vTX), and virtual RF (vRF). The Software Defined Radio (SDR) handles actual RF signals according to the configuration provided by the vRF, while the vTX and vRX process signals at a baseband level. The vTX generates baseband signals based on SFLC symbols encoded from data bits and delivers it to the vRF. The baseband signal is passed to the SDR and emitted as an actual wireless signal through the antenna. As the SDR on the UE side receives a wireless signal, it converts it to a baseband signal and delivers it to the vRX via the vRF. The vRX then processes the baseband signal with SFLC decoding. A similar scenario occurs when sending UL sounding at the UE side and acquiring CSIT at the BS side.

To apply the proposed CCA to the software modem, we modified the vRX to handle estimated PRO after SFLC combining with the received SFLC symbols. Based on the estimated PRO, the vRX performs CCA to compensate for the channel amplitude asymmetry. To achieve this, we allow the vRX to command the receive gain adjustments to a certain RF chain, as shown in Fig. 1. As the vRF receives the command, it configures the SDR to adjust its receive gain by the compensation factor. The PRO estimation and receive gain adjustment are repeated until the PRO is estimated to be nearly 0, which reveals that channel amplitude is symmetric.

The software modem operates on a real-time testbed, which allows for experiments with on-air signal emission. Figs. 7 and 8 illustrate the testbed, where each component consists of a LINUX PC and a USRP B210. To configure a 1×2 SFLC-OFDM system, we attach one antenna

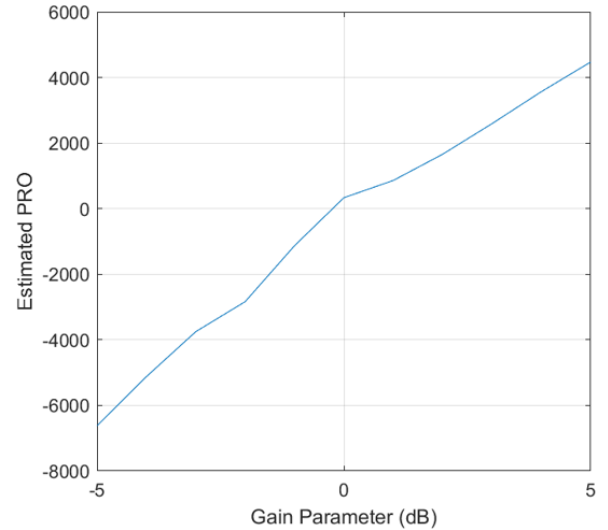


FIGURE 9. PRO values according to the receive gain.

to the BS-side USRP and two antennas to the UE-side USRP. The antennas are connected to the TX/RX ports of the USRPs, ensuring channel reciprocity in the wireless sections through simultaneous transmission and reception. The detailed parameters of the testbed are configured as shown in Table 2.

The testbed provides a channel environment that reflects the radiation of actual signals. In this environment, both the line-of-sight(LOS) path and non-LOS components naturally create a multipath environment. The asymmetry of the channel is controlled by adjusting the receive gain. We modify the receive gain of the 2nd antenna on the UE side to adjust the α_2 value. We modify the channel gain parameter of the 2nd antenna in Linux. It affects $(\alpha_1 - \alpha_2)$ and generates various asymmetric channel conditions. These setups enable the evaluation of CCA performance across various channel scenarios.

A. FUNCTIONALITY ASSESSMENT OF PRO ESTIMATION

We initially examine the functionality of the PRO estimator based on the asymmetric channel conditions. Fig. 9 shows the estimated PRO values according to the amount of the receive gain adjustment. The results show that the estimated PRO gets linearly higher as the receive gain increases and channel becomes more asymmetric. This trend validates that the PRO estimator is effective to measure channel asymmetry and to recognize how the receive gain needs to be compensated.

To evaluate the functionality of our PRO estimator from a stability perspective, we observe the instantaneous outputs of the PRO estimated in practical channel environments. The UE needs to estimate PRO using received signals that experience a certain amount of noisy, asymmetric, and randomly time-delayed channels. The estimated PRO values essentially fluctuate for each estimation trial due to the random characteristics of the received signals. This

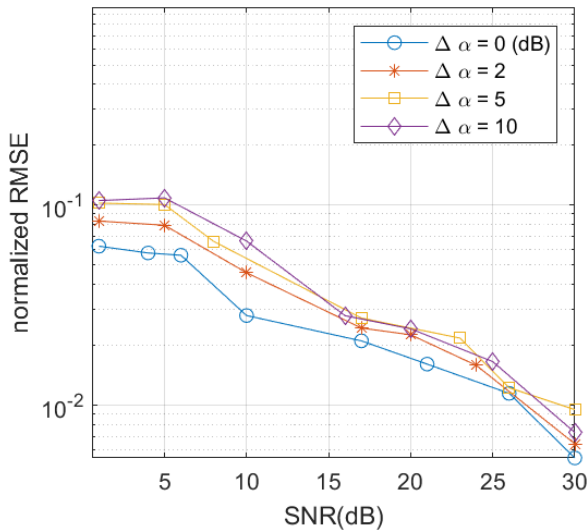


FIGURE 10. The N-RMSE of estimated PRO with various α_2 values.

fluctuation may impact the performance of data-aided CCA that utilizes the estimated PRO value as input. Therefore, this experiment focuses on observing the stability of the PRO estimator by measuring the normalized root mean square error (N-RMSE). The error term is derived from the difference between an instantaneously estimated PRO value and the average of estimated PRO values with SNR above 30dB.

Fig. 10 illustrates the measured results of N-RMSE for various channel amplitude asymmetry cases. The results demonstrate that the PRO estimator functions effectively across various signal strength cases. The N-RMSE falls below 0.02 when the SNR is 25 dB or higher, indicating that the estimated PRO values are stably acquired in high SNR conditions. The N-RMSE ranges between 0.02 and 0.04 and does not drastically increase when the signal strength is at a mid-level, with the SNR between 15 dB and 25 dB. From these results, we can conclude that PRO can be stably estimated if the signal strength of the received SFLC symbols is at a sufficient level. On the other hand, the N-RMSE increases noticeably if the SNR is lower than 15 dB, as noise introduces a certain level of error into the received SFLC symbols. To this end, we can see that the PRO estimator is affected by noise, similar to other data-aided algorithms.

The results also demonstrate the performance of the PRO estimator across different levels of channel amplitude asymmetry. The N-RMSE is minimal when the channel is symmetric and increases as $\Delta\alpha$ increases. This implies that the PRO estimator is more stable at lower levels of channel amplitude asymmetry. Moreover, the difference in N-RMSE between the cases where $\Delta\alpha$ is 0dB and 10dB decreases as the SNR increases and becomes negligible when the SNR is greater than 15 dB. This shows that the PRO estimator operates stably, regardless of the degree of channel amplitude asymmetry, in high SNR conditions.

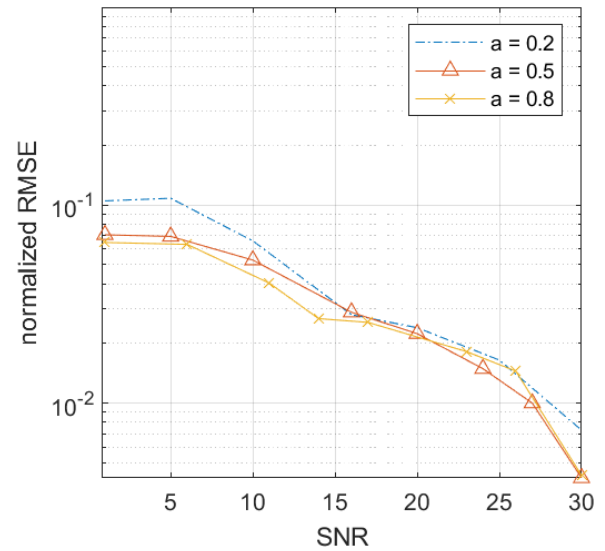


FIGURE 11. The N-RMSE of estimated PRO with various IIR filtering constants.

Fig. 11 depicts the measured results of N-RMSE for various IIR filtering constants when $\Delta\alpha$ is 10dB. The results demonstrate the contribution of the one-tap IIR filtering to the PRO estimator as the N-RMSE decreases with a higher filtering constant in low SNR conditions. The improvement from the one-tap IIR filtering is noticeable when the SNR ranges between 10 dB and 15 dB, indicating that the PRO estimator becomes more stable in these signal environments. The one-tap IIR filtering also shows a significant effect when the SNR is 15 dB, as the N-RMSE decreases to approximately 0.03 with an IIR filtering constant of 0.8. Hence, our design to utilize the one-tap IIR filtering helps to acquire stable outputs of the PRO estimator in weak signal environments when its constant is properly adjusted according to the signal conditions.

B. PERFORMANCE IMPROVEMENT OF SYMBOL DECODING

We further evaluate how the data-aided CCA benefits the overall SFLC-OFDM system from the perspective of decoding performance. The impact of channel gain feedback from the PRO estimator can be observed through the phase errors of the SFLC decoded symbols. These can be expressed using the normalized EVM, which represents the phase difference between the decoded and transmitted power-normalized symbols. After SFLC combining, the normalized EVM is calculated using the Euclidean distance between $\hat{X}_k$ and X_k [51].

Fig. 12 illustrates the normalized EVM of decoded symbols for various levels of channel amplitude asymmetry. The results confirm a reduced normalized EVM due to the CCA, ensuring that the decoded symbols after CCA contain fewer errors. We can also observe that the normalized EVM reduces more significantly when α_2 is larger. The normalized

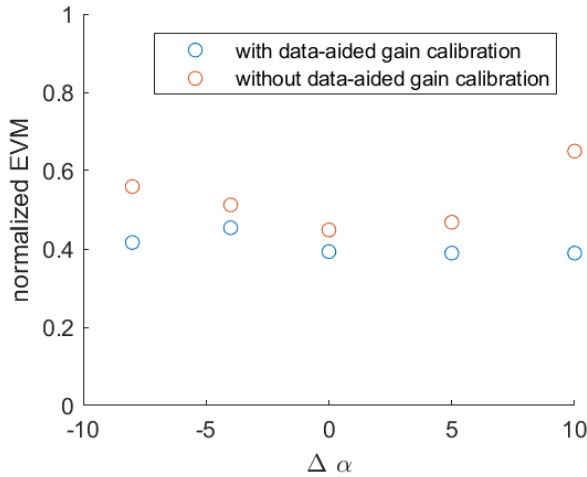


FIGURE 12. EVMs before and after applying the data-aided CCA.

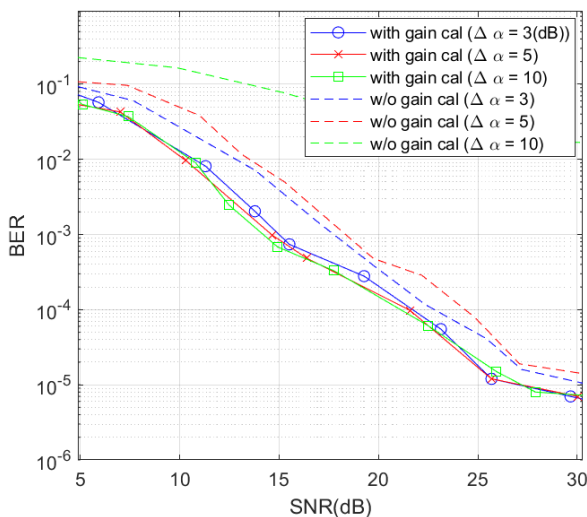


FIGURE 13. The BER of SFLC decoded symbols.

EVM with the CCA remains similar regardless of channel asymmetric level. We can conclude that the data-aided CCA effectively improves the quality of decoded symbols as channel amplitude asymmetry is significant. Additionally, the results show that the normalized EVM further reduces even in the configuration without channel amplitude asymmetry. This occurs because there is a residual amount of channel amplitude asymmetry even after we configure the channel to be symmetric during the test. Hence, the results demonstrate the delicate capability of the proposed CCA to recover channel amplitude asymmetry.

Fig. 13 presents the decoding performance in terms of bit error rate (BER) for various levels of signal and channel amplitude asymmetry. The dotted and solid lines represent the BER of SFLC decoded symbols with and without applying the data-aided CCA, respectively. Without the CCA, the BER does not drop below 1% in a high SNR regime if the channel amplitude asymmetry is significant. This is because

the channel amplitude asymmetry functions as additional noise and degrades the SFLC combining process. On the other hand, with the CCA, the BER consistently approaches 0 in a high SNR regime, regardless of the degree of channel amplitude asymmetry. This reveals that data-aided CCA effectively contributes to improving decoding performance in severely asymmetric channel environments. Additionally, the CCA achieves an SNR gain of up to 2 dB, even with a small level of channel amplitude asymmetry. This proves the feasibility of our proposed CCA in enhancing the decoding performance of the SFLC-OFDM system across a wide range of scenarios.

V. CONCLUSION

This paper highlights the performance degradation caused by channel amplitude asymmetry in a TDD-based SFLC-OFDM system that relies on CSIT. We mathematically demonstrate that channel amplitude asymmetry induces unexpected phase rotation in the decoded symbols, disrupting the decoding process. To address this issue, we designed a data-aided CCA to estimate and compensate for the phase rotation based on the decoded symbols. We provide a mathematical model to estimate the phase rotation through straightforward operations involving the decoded and hard-decision symbols. Utilizing this estimator, we developed a data-aided CCA to compensate for channel amplitude asymmetry. The proposed CCA requires no additional reference signals or hardware, offering operational complexity reduction and system scalability. We implemented the entire algorithm on a software modem operating within a real-time testbed for practical evaluation. The experimental results from the testbed demonstrate the stability and effectiveness of the proposed CCA, indicating its feasibility in addressing the issue of channel amplitude asymmetry in the SFLC-OFDM system.

The proposed CCA ensures the channel reciprocity in a TDD system and contributes to relaxing the pilot overhead in massive MIMO systems. The proposed CCA can realize channel reciprocity in Intelligent Reflecting Surface (IRS) systems with minimizing channel estimation overhead due to many IRS control elements. In the case of NOMA, the proposed CCA enables efficient precoding in Non-Orthogonal Multiple Access (NOMA) systems without requiring explicit channel feedback. These impacts highlight the potential for future research of extensive application of the proposed CCA to 5G and B5G technologies.

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