Channel Estimation for Reconfigurable Intelligent Surface-Assisted Wireless Communications Considering Doppler Effect

Shu Sun and Hangsong Yan

Abstract—In wireless systems aided by reconfigurable intelligent surfaces (RISs), channel state information (CSI) plays a pivotal role in achieving the performance gain of RISs. In this letter, we propose a novel Doppler shift adjustment (DSA) method to improve CSI estimation accuracy in the presence of motions at the communication ends for RIS-assisted systems. Specifically, the estimated channel frequency response is transformed into the time domain to facilitate phase shift adjustment, which is then converted back to the frequency domain. The proposed DSA necessitates only one more symbol which incurs negligible extra overhead compared with the large number of symbols originally required for channel estimation. Further, low-complexity channel estimation and DSA schemes are designed for approximately single-path scenarios applicable to millimeter-wave and terahertz systems. Simulation results demonstrate the effectiveness of the proposed DSA and channel estimation methods.

Index Terms—Reconfigurable intelligent surface (RIS), channel estimation, Doppler shift, millimeter wave.

I. INTRODUCTION

RECONFIGURABLE intelligent surfaces (RISs), also known as holographic multiple-input-multiple-output (MIMO), intelligent beamforming metasurfaces, and intelligent reflecting surfaces (when used for reflection), among others, have recently stimulated an upsurge in research on their applications in wireless communications, due to their capability of smartly sensing and controlling the propagation environment [1]-[3]. An RIS comprises sub-wavelength metallic or dielectric scattering elements, which is capable of shaping electromagnetic waves via anomalous reflection, refraction, absorption, and/or polarization [2]-[4]. By appropriately and dynamically adjusting the amplitude and/or phase of each of the RIS elements based on the propagation environment, wireless signals can be coherently combined and steered to desired directions. Achieving the full potential of RISs requires the acquisition of accurate channel state information (CSI), which, however, usually incurs considerable overhead stemming from the large number of elements at an RIS.

Pioneering works have been conducted to tackle the channel estimation (CE) issue in RIS-assisted systems. For instance, the authors in [5] and [6] have employed RISs where all or a portion of the elements are active to facilitate CE. In contrast to active RIS elements, purely passive RIS elements are more energy-efficient and cost-effective. For RISs with all passive elements, CSI estimation algorithms capitalizing on the channel sparsity are presented in [7]-[9], while a more general CE framework is provided in [10]. Nevertheless, these works and analyzes are aimed for narrow-band wireless systems, while 5G and future 6G communications are likely to conduct broadband deployment. Wideband CE for RIS-aided systems is studied in [11], where the cascaded user-RIS-base-station channel is estimated by joint design of pilot sequences and the reflection pattern at the RIS, and a low-complexity scheme is then proposed to maximize the achievable rate by optimizing the RIS reflection coefficients. Furthermore, the authors in [12] leveraged compressive sensing to estimate the broadband channel for millimeter-wave (mmWave) systems.

All the investigations above assume a quasi-static channel that remains approximately constant during CSI estimation, which, in reality, may not hold owing to motions at the communication ends and/or in the environment. In this letter, we propose a method to compensate for the Doppler effect for wideband CE in RIS-assisted wireless systems. Specifically, we first adopt the basic CSI estimation framework proposed in [11] and design a novel mechanism to adjust Doppler-induced distortions in estimated signals via frequency- and time-domain transformations. Then low-complexity CE and Doppler shift adjustment (DSA) strategies are proposed tailored to mmWave and terahertz (THz) communication systems. To our best knowledge, this is the first work on DSA for CE in RIS-assisted systems.

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II. SYSTEM MODEL AND PROBLEM FORMULATION

A. System Model

We consider an uplink point-to-point communication system assisted by an RIS, which consists of an $N_G$-antenna next-generation node B (gNB), an $N_U$-antenna user equipment (UE), and a two-dimensional passive RIS comprising $N_R$ ($N_R \gg N_G, N_R \gg N_U$) reconfigurable elements, as illustrated in Fig. 1, where the RIS is employed to enhance the communication by creating a reflecting link between the gNB and the UE. To reduce the complexity of CE and hardware cost, and given the high correlation of close-by elements in the RIS due to sub-wavelength spacing, the entire RIS is divided into $M$ sub-surfaces, each of which contains $N_R/M$ (set to be an integer) adjacent elements.

In this work, frequency-selective fading channels are considered to represent many practical wireless channels. The total bandwidth allocated to the UE is divided into $N$ subcarriers, or equivalently, $N_B = N/12$ resource blocks (RBs) [13]. The channel frequency response (CFR) for the UE-gNB, UE-RIS, and RIS-gNB links are denoted by $\mathcal{H}_{\text{UG}} \in \mathbb{C}^{N \times N}$, $\mathcal{H}_{\text{UR}} \in \mathbb{C}^{M \times N \times N}$, and $\mathcal{H}_{\text{RG}} \in \mathbb{C}^{N_G \times M \times N}$, respectively. Without loss of generality, we assume that the maximum delay spread is of $L$ taps in the time domain for the equivalent baseband equivalent channels of both the UE-gNB direct link and the UE-RIS-gNB reflecting link. For the link between the $u$-th (for $u \in \{0, ..., N_U - 1\}$) antenna at the UE and the $g$-th (for $g \in \{0, ..., N_G - 1\}$) antenna at the gNB, each transmitted orthogonal frequency-division multiplexing (OFDM) symbol is denoted as $\mathbf{x} = [x_g, x_{g1}, ..., x_{g, N-1}]^T \in \mathbb{C}^{N \times 1}$ (where the subscripts for antenna indices are omitted for simplicity), which is first transformed into the time domain through an $N$-point inverse discrete Fourier transform (IDFT), followed by the appending of a cyclic prefix (CP) of length $L_{CP}$ that is assumed to be larger than $L$. At the gNB, after removing the CP and performing the $N$-point discrete Fourier transform (DFT), the baseband received signal in the frequency domain is given by (the symbol index is omitted herein)

$$
\mathbf{y} = \left( \sum_{m=0}^{M-1} \mathbf{h}_{\text{RG},m} \circ \phi_m \mathbf{h}_{\text{UR},m} \right) + \mathbf{h}_{\text{UG}} \circ \mathbf{x} + \mathbf{w}
$$

where $\circ$ denotes the Hadamard product, $\mathbf{y} \in \mathbb{C}^{N \times 1}$ is the received OFDM symbol, $\mathbf{h}_{\text{RG},m} \in \mathbb{C}^{N \times 1}$ is the CFR of the RIS-gNB link associated with the $m$-th sub-surface at the RIS, $\phi_m = \alpha_m e^{j\phi_m} \quad (\forall m \in \{0, ..., M-1\})$ is the reflection coefficient of the $m$-th sub-surface, $\mathbf{h}_{\text{UR},m} \in \mathbb{C}^{N \times 1}$ is the CFR of the UE-RIS link for the $m$-th sub-surface, $\mathbf{h}_{\text{UG}} \in \mathbb{C}^{N \times 1}$ represents the CFR of the UE-gNB direct link, and $\mathbf{w} \in \mathbb{C}^{N \times 1}$ stands for the received additive white Gaussian noise (AWGN) with $\mathbf{w} \sim \mathcal{CN}(\mathbf{0}, \sigma^2 \mathbf{I}_N)$. For the purpose of maximizing the reflection power of the RIS and simplify its hardware design, we set $\alpha_m = 1, \forall m \in \{0, ..., M - 1\}$.

Let $\mathbf{h}_{\text{UR},m} = \mathbf{h}_{\text{RG},m} \circ \mathbf{h}_{\text{UR},m} \in \mathbb{C}^{N \times 1}$ denote the effective cascaded CFR of the reflecting link for the $m$-th sub-surface (without the effect of RIS phase shift), and $\mathbf{X} = \text{diag}(\mathbf{x}) \in \mathbb{C}^{N \times N}$ the diagonal matrix of $\mathbf{x}$, then (1) can be recast as

$$
\mathbf{y} = \mathbf{X} \left( \sum_{m=0}^{M-1} \phi_m \mathbf{h}_{\text{UR},m} \right) + \mathbf{h}_{\text{UG}} + \mathbf{w} \quad (2)
$$

Note that it may not be accurate to first estimate $\mathbf{h}_{\text{UG}}$ separately, e.g., by turning off all the elements at the RIS, because the powered-down RIS can still act as an ordinary scattering object or (lossy) reflector, among other objects in the channel, especially when it is in the vicinity of the UE or the gNB, hence affecting the direct link. Therefore, it is desirable to estimate the holistic equivalent CFR incorporating both the direct link and (enhanced) reflecting link. By concatenating $\mathbf{h}_{\text{UR},m}$ into $\mathbf{H}_{\text{UR}} = [\mathbf{h}_{\text{UR},0}, ..., \mathbf{h}_{\text{UR},M-1}] \in \mathbb{C}^{N \times M}$ and set $\mathbf{\phi} = [\phi_0, \phi_1, ..., \phi_{M-1}]^T \in \mathbb{C}^{M \times 1}$, (2) can be expressed more succinctly as

$$
\mathbf{y} = \mathbf{X} (\mathbf{H}_{\text{UR}} \mathbf{\phi} + \mathbf{h}_{\text{UG}}) + \mathbf{w} = \mathbf{X} \mathbf{h} + \mathbf{w} \quad (3)
$$

where $\mathbf{h} = \mathbf{H}_{\text{UR}} \mathbf{\phi} + \mathbf{h}_{\text{UG}} \in \mathbb{C}^{N \times 1}$ embodies the superimposed CFR of the reflecting and direct links.

B. Wideband Channel Estimation

With the simple form in (3), $\mathbf{h}$ can be estimated via various conventional CE techniques. One popular and standard approach is to employ pilot signals, such as the sounding reference signal [13], to estimate the uplink channel to facilitate the subsequent resource allocation and scheduling. Denoting the estimation of $\mathbf{h}$ as $\hat{\mathbf{h}}$, $\mathbf{H}_{\text{UR}}$ and $\mathbf{h}_{\text{UG}}$ need to be resolved after $\hat{\mathbf{h}}$ is obtained. The authors in [11] proposed a strategy using a pre-designed reflection pattern at the RIS to aid the estimation of $\mathbf{H}_{\text{UR}}$ and $\mathbf{h}_{\text{UG}}$. Specifically, the reflection pattern matrix $\mathbf{\Theta}$ is designed to be an $(M + 1) \times (M + 1)$ DFT matrix where $[\mathbf{\Theta}]_{p,q} = e^{-j \frac{2 \pi p q}{M+1}}$ with $0 \leq p,q \leq M$. Thus, the estimation of $\mathbf{H}_{\text{UR}}$ and $\mathbf{h}_{\text{UG}}$ can be obtained as follows [11]

$$
\hat{\mathbf{h}}_{\text{UG}}, \hat{\mathbf{H}}_{\text{UR}} = \mathbf{H} \mathbf{\Theta}^{-1}
$$

where $\mathbf{\hat{H}} = [\mathbf{\hat{h}}^{(0)}, \mathbf{\hat{h}}^{(1)}, ..., \mathbf{\hat{h}}^{(M)}] \in \mathbb{C}^{N \times (M+1)}$ with $\mathbf{\hat{h}}^{(i)}$ representing the estimation of the superimposed CFR in the $i$-th symbol. The proposed CE scheme in [11] outperforms some prior methods in the literature. A crucial assumption in [11] is that the channel is quasi-static within the transmission frame such that $\mathbf{H}_{\text{UR}}$ and $\mathbf{h}_{\text{UG}}$ remain approximately constant. However, since the number of symbols required for CE is $M + 1$, the estimation duration is long for large values of $M$ and the assumption of quasi-static channel may not hold in practical scenarios, hence degrading CE performance and the resultant achievable rate.
C. Problem Formulation

Consider the situation where the UE is moving at a certain velocity \( v \), the induced phase change for a multipath component is [14] \( \Delta \psi = 2\pi v \Delta t \cos(\theta) / \lambda \), with \( \Delta t \) denoting the time passed with respect to a reference moment, \( \lambda \) the wavelength, and \( \theta \) the traveling angle of a multipath component with respect to the moving direction [14]. For a single multipath component, the Doppler effect gives rise to a phase change proportional to \( \Delta t \); while for a wideband signal containing more than one multipath component, it may encounter both phase and amplitude alterations due to the superposition of the phase-shifted multipath components. If the UE is moving, the true CFR \( h_{\text{UG}} \) and \( h_{\text{URG}} \) will vary over the \( M + 1 \) symbols during CE due to Doppler shift. Based on (3), the true CFR in the \( i \)-th symbol can be rewritten as

\[
h^{(i)} = h_{\text{URG}}^{(i)} \Phi^{(i)} + h_{\text{UG}}^{(i)}
\]

Without loss of generality, we take the 1\textsuperscript{st} symbol as the reference symbol, then the Doppler effect on the subsequent symbols can be modeled as

\[
h^{(i)} = (h_{\text{URG}}^{(i)} + \Xi^{(i-1)}) \Phi^{(i)} + (h_{\text{UG}}^{(i)} + \epsilon^{(i-1)})
\]

\[
= h_{\text{URG}}^{(i)} \Phi^{(i)} + \Xi^{(i-1)} \Phi^{(i)} + h_{\text{UG}}^{(i)} + \epsilon^{(i-1)}
\]

where \( \Xi^{(i-1)} \) and \( \epsilon^{(i-1)} \) represent the variation of \( h_{\text{URG}}^{(i)} \) and \( h_{\text{UG}}^{(i)} \) in the \( i \)-th symbol with respect to the 1\textsuperscript{st} symbol, respectively. \( h_{\text{URG}}^{(i)} = [h_{\text{URG}}^{(1)}, h_{\text{URG}}^{(2)}, ..., h_{\text{URG}}^{(M+1)}] \) and \( h_{\text{UG}}^{(i)} = \left[ \begin{array}{l} \tilde{h}_{\text{UG}}^{(i)} \\ \end{array} \right] \). Stacking \( h^{(i)} \) in (6) over \( M + 1 \) consecutive symbols yields

\[
H = h_{\text{URG}}^{(1)} \Theta + [\Xi^{(0)}, \Xi^{(1)}, ..., \Xi^{(M)}] \Theta^{-1} + V \Theta^{-1}
\]

Therefore,

\[
\hat{H} \Theta^{-1} = \hat{h}_{\text{URG}}^{(1)} \Theta + [\Xi^{(0)}, \Xi^{(1)}, ..., \Xi^{(M)}] \Theta^{-1} + V \Theta^{-1}
\]

where \( \Theta \) is the pre-designed reflection pattern used in (4), and \( V \in \mathbb{C}^{M \times (M+1)} \) is the noise term. It is evident from (8) that with Doppler effect, the equivalent CFR estimated by (4) is \( \hat{h}_{\text{URG}}^{(1)} + [\Xi^{(0)}, \Xi^{(1)}, ..., \Xi^{(M)}] \Theta^{-1} \) instead of \( h_{\text{URG}}^{(1)} \), resulting in a non-negligible estimation error. The problem is how to account for the Doppler effect in (8) to obtain more accurate estimation of \( \hat{h}_{\text{URG}}^{(i)} \) \((\forall i \in \{1, ..., M + 1\})\).

III. PROPOSED DOPPLER SHIFT ADJUSTMENT

In this section, we present a DSA algorithm to address the CE inaccuracy issue mentioned above. As shown in (8), the estimation error incurred by Doppler effect is tangled with the pre-designed reflection pattern \( \Theta \) which needs to be separated out for the evaluation of Doppler shift alone. To this end, it is necessary to add one more symbol that shares the reflection pattern with one of the \( M + 1 \) symbols adopted for CE, and the extra symbol should be adjacent to the pattern-sharing symbol to precisely capture the CFR variation over one symbol. For simplicity, the extra symbol is set as the 0\textsuperscript{th} symbol and the associated reflection pattern is \( \Phi^{(0)} = \Phi^{(1)} = [1, ..., 1]^T \in \mathbb{C}^{M \times 1} \), i.e., it shares the reflection pattern with the 1\textsuperscript{st} symbol.

Inspired by the fact that the motion of the UE leads to a phase shift for each multipath component in the channel impulse response (CIR), it is easier to adjust the Doppler effect in the time domain as opposed to the frequency domain. Thus, for the estimated superimposed CFR in the \( i \)-th symbol \( \hat{h}_{\text{URG}}^{(i)} \), the Doppler shift \( \Delta \beta_k \) in the \( k \)-th tap is adjusted as if it were the CIR

\[
\begin{align*}
\hat{g}_{\text{URG}}^{(i)}(p) & = \sum_{k=0}^{N-1} a_k^{(i)} \delta(p - k), p = 0, 1, ..., N - 1
\end{align*}
\]

(9)

where \( p \) is the tap index, \( a_k^{(i)} \) and \( \beta_k^{(i)} \) denote the amplitude and phase of the \( k \)-th tap in \( \hat{g}_{\text{URG}}^{(i)} \), respectively. The position and value of \( a_k^{(i)} \) are assumed to be consistent over the \( M + 2 \) symbols, which is reasonable since the Doppler effect only alters the phase of each tap. To lower the computational complexity and mitigate the impact of random noise, only the taps whose amplitudes are no smaller than a pre-defined threshold (e.g., 0.1) with respect to the maximum amplitude are selected in \( \hat{g}_{\text{URG}}^{(i)} \) for succeeding phase adjustment, and the set of such taps is denoted as \( \mathcal{P} \). Then (9) can be re-organized as

\[
\hat{g}_{\text{URG}}^{(i)}(p) = \sum_{k \in \mathcal{P}} a_k^{(i)} e^{j \beta_k^{(i)}(p - k)} + \sum_{k \notin \mathcal{P}} a_k^{(i)} e^{j \beta_k^{(i)}(p - k)}
\]

(10)

The phase of the \( k \)-th (\( \forall k \in \mathcal{P} \)) tap is compared between \( \hat{g}_{\text{URG}}^{(i)} \) and \( \hat{g}_{\text{URG}}^{(0)} \), i.e., \( \Delta \beta_k^{(i)} = \beta_k^{(i)} - \beta_k^{(0)} \), which is regarded as the phase difference of the \( k \)-th tap between adjacent symbols. Next, to estimate \( \hat{h}_{\text{URG}}^{(q)} \) \((\forall q \in \{1, ..., M + 1\})\), the phase of the \( k \)-th (\( \forall k \in \mathcal{P} \)) tap in each \( \hat{g}_{\text{URG}}^{(i)} \) is adjusted based on \( \Delta \beta_k^{(i)} \)

\[
\hat{g}_{\text{URG}}^{(i)}(p) = \sum_{k \in \mathcal{P}} a_k^{(i)} e^{j (\beta_k^{(i)} - \Delta \beta_k^{(i)}(p - q))} + \sum_{k \notin \mathcal{P}} a_k^{(i)} e^{j \beta_k^{(i)}(p - k)}
\]

(11)

This way, \( \hat{g}_{\text{URG}}^{(i)} \) is adjusted as if it were the CIR (including the reflection pattern at the RIS) associated with the i-th symbol without Doppler shift. \( \hat{g}_{\text{URG}}^{(i)} \) is then converted back to the frequency domain via the \( N \)-point DFT to yield \( \hat{h}_{\text{URG}}^{(i)} \). Finally, the adjustment of \( h_{\text{URG}}^{(q)} \) and \( h_{\text{URG}}^{(q)} \) is given by

\[
[\hat{h}_{\text{URG}}^{(q)}, \hat{h}_{\text{URG}}^{(q)}] = \hat{H}_{q, \text{adj}} \Theta^{-1}
\]

(12)

where \( \hat{H}_{q, \text{adj}} = [\hat{h}_{\text{URG}}^{(1)}, \hat{h}_{\text{URG}}^{(2)}], ..., \hat{h}_{\text{URG}}^{(M+1)}] \in \mathbb{C}^{N \times (M+1)} \).

Essentially, more accurate estimation of \( \hat{h}_{\text{URG}}^{(q)} \) and \( h_{\text{URG}}^{(q)} \) \((\forall q \in \{1, ..., M + 1\})\) is realized by reconstructing \( \hat{h}_{\text{URG}}^{(i)} \) via phase adjustment in the corresponding CIRs.

IV. CHANNEL ESTIMATION FOR SINGLE-PATH SCENARIO

In this section, we propose a novel and low-complexity CE scheme when the channel has (approximately) one path/tap in the UE-RIS and RIS-gNB links. This scenario is particularly applicable to mmWave and THz communication channels due to severe propagation loss caused by atmospheric attenuation, as well as reflection, scattering, and penetration loss at those high carrier frequencies [3]. For simplicity, the UE-gNB direct link is neglected for the subsequent analysis, yet the extension
to the scenario with the direct link is straightforward. Under this condition, the formulation in (3) for each sub-carrier can be simplified to
\[
y = x \sum_{m=0}^{M-1} \phi_m h_{UR,m} + w = x \sum_{m=0}^{M-1} [\tilde{h}_{RG} \odot h_{UR}]_m + w
\]
where we set \( \phi_m = 1 \) (\( \forall m \in \{0, ..., M - 1\} \)), \( h_{UR} \in \mathbb{C}^{M \times 1} \) and \( h_{RG} \in \mathbb{C}^{M \times 1} \) are the CFRs for the UE-RIS and RIS-gNB links, respectively, \( \tilde{h}_{RG} \) denotes the conjugate of \( h_{RG} \). The target herein is to estimate \( \tilde{h}_{RG} \odot h_{UR} \) for all sub-surfaces at the RIS (and for all sub-carriers as well). For each sub-carrier and each pair of the UE and gNB antennas, the mmWave propagation environment is well characterized by the Saleh-Valenzuela model depicted as [7] (with a slight abuse of notations)
\[
\tilde{h}_{UR} = \sqrt{M/P} \sum_{p=0}^{P-1} \alpha_p a_R(\theta_p, \varphi_p)
\]
\[
h_{RG} = \sqrt{M/Q} \sum_{q=0}^{Q-1} \alpha_q a_R(\theta_q, \varphi_q)
\]
where \( \alpha_p \) and \( \theta_p, \varphi_p \) denote the complex gain and azimuth (elevation) angle of arrival (AoA) of the \( p \)-th path of the UE-RIS channel, respectively. Likewise, \( \alpha_q \) and \( \theta_q, \varphi_q \) represent the complex gain and azimuth (elevation) angle of departure (AoD) of the \( q \)-th path of the RIS-gNB channel. Additionally, \( a_R \) denotes the array response vector at the RIS. For a uniform square array with \( \sqrt{M} \times \sqrt{M} \) elements, \( a_R \) is given by
\[
a_R(\theta, \varphi) = \frac{1}{\sqrt{M}} \begin{bmatrix} e^{j \frac{2\pi}{\lambda} d(u \sin \theta \sin \varphi + v \cos \varphi)} & \ldots & e^{j \frac{2\pi}{\lambda} d((\sqrt{M} - 1) \sin \theta \sin \varphi + (\sqrt{M} - 1) \cos \varphi)} \end{bmatrix}^T
\]
where \( d \) is the antenna spacing, and \( 0 \leq u, \tilde{u} < \sqrt{M} \) are the element indices in the 2D plane. When there is one dominant path (denoted as the 0-th path without loss of generality) whose amplitude is substantially larger than all the other paths, (14) and (15) can be approximated as
\[
\tilde{h}_{UR} \approx \sqrt{\frac{M}{P}} \alpha_0 a_R(\theta_0, \varphi_0), \quad h_{RG} \approx \sqrt{\frac{M}{Q}} \alpha_0 a_R(\theta_0, \varphi_0)
\]
such that
\[
\tilde{h}_{RG} \odot h_{UR} \approx A[1, ..., a^u b^v, ..., a^{\sqrt{M} - 1} b^{\sqrt{M} - 1}]^T
\]
where \( A = \frac{\alpha_0}{\sqrt{\lambda d}}, \quad a = e^{j \frac{2\pi}{\lambda} d(\sin \theta_0 \sin \varphi_0 - \sin \theta_0 \sin \varphi_0)}, \quad b = e^{j \frac{2\pi}{\lambda} d(\cos \theta_0 - \cos \varphi_0)} \) are constants relying solely on the complex gains, and azimuth and elevation AoA and AoD of the dominant path. (18) indicates that \( \tilde{h}_{RG} \odot h_{UR} \) can be calculated if \( A, a, \) and \( b \) are known. Theoretically only three pilot symbols are needed to acquire \( A, a, \) and \( b \), while DSA necessitates one more symbol. Therefore, only four pilot symbols are needed to obtain \( \tilde{h}_{RG} \odot h_{UR} \) regardless of the value of \( M \), which significantly reduces the CE time and hence the overall latency as compared with the situation in Section III.

The proposed CE approach is detailed below. When sending the first two pilot symbols, all the sub-surfaces at the RIS are turned off except the 0-th one (\( u = \tilde{u} = 0 \) in (18)), which gives
\[
\tilde{A}(0) = y(0)/x, \quad \tilde{A}(1) = y(1)/x. \quad \text{Since there is approximately only one path, the Doppler phase shift between adjacent symbols can be conveniently obtained by comparing the phases of \( \tilde{A}(1) \) and \( \tilde{A}(0) \), i.e.,}
\]
\[
\Delta \zeta = \angle \tilde{A}(1) - \angle \tilde{A}(0)
\]
Afterwards, for the next two pilot symbols, all the sub-surfaces at the RIS are turned off except the one corresponding to (\( u = 1, \tilde{u} = 0 \)) and (\( u = 0, \tilde{u} = 1 \)) in (18), respectively, yielding
\[
\tilde{A}(2)e^{j \Delta \zeta} = \tilde{A}(1)e^{j \Delta \zeta} = y(2)/x
\]
\[
\tilde{A}(3)e^{j \Delta \zeta} = \tilde{A}(1)e^{j \Delta \zeta} = y(3)/x
\]
(19) has been invoked. Accordingly,
\[
\tilde{A}(1) = \frac{e^{j(1-1)\Delta \zeta} y(1)}{y(1)}, \quad \tilde{b} = \frac{e^{-j \Delta \zeta} y(3)}{y(1)}
\]
Consequently, all the elements in \( \tilde{h}_{RG} \odot h_{UR} \) can be derived based on (19) and (22), hence solving the CE problem.

V. NUMERICAL RESULTS

In this section, we provide simulation results for performance evaluation of the proposed DSA and single-path CE methods. A uniform square array is adopted at the RIS consisting of 576 elements with \( \lambda/4 \) spacing. The center carrier frequency is 28 GHz, the number of RBs is 16 with 30 kHz sub-carrier spacing, the noise power is about -106 dBm, and 40 dB processing gain is assumed at the gNB which is practical for mmWave receivers to enhance the link budget [15]. The Zadoff-Chu sequence is utilized as the pilot sequence [13]. For both the direct and reflecting links, 6-tap channels are assumed with \( \eta \) being the ratio of the total power of non-dominant components to that of the dominant component. The UE-RIS and RIS-gNB distances are 5 m and 50 m, respectively, while the UE location is randomly selected in each simulation run. The path loss exponents for the UE-RIS, RIS-gNB, and UE-gNB links are set to 2.0, 2.1, and 3.5, respectively. The CE and beamforming method in [11] is adopted as the baseline in Fig. 2 to Fig. 4.

Fig. 2 illustrates the CE mean square error (MSE) normalized to the channel gain for the reflecting link with \( M = 16 \) and different UE velocities \( v \) and \( \eta \). It can be observed that DSA always produces more accurate estimation under the same condition. For the same \( \eta \), the MSE gap between with and without DSA cases increases with \( v \) as expected, since larger \( v \) incurs more severe Doppler shift. Moreover, the MSE increases with \( \eta \) and \( v \) for both with and without DSA cases, because more diffused multipath components and larger moving speed enlarges phase variation over symbols.

The associated achievable rate ratio with respect to the perfect CSI case is demonstrated in Fig. 3, which shows that the achievable rate ratio clearly deteriorates without DSA, especially for large \( v \). Specifically, the maximum achievable rate reduction reaches 16% without DSA, while it is only 6% with DSA. Note that although the MSE for \( \eta = 1 \) is higher than for \( \eta = 0.1 \) in Fig. 2, the corresponding rate ratios are the opposite, since the absolute achievable rate with perfect CSI decreases with \( \eta \), attributed to the strongest-CIR-alignment.
reflection optimization scheme in [11].

The effect of $M$ is examined in Fig. 4, which shows superior performance of the proposed DSA over that without DSA. The MSE and hence the achievable rate ratio degrade with $M$, as larger $M$ results in higher phase resolution thus making CE more vulnerable to phase variation caused by noise.

![Normalized MSE versus transmission power for $M = 16$.](image)

**Fig. 2.** Normalized MSE versus transmission power for $M = 16$.

![Achievable rate ratio versus transmission power for $M = 16$.](image)

**Fig. 3.** Achievable rate ratio versus transmission power for $M = 16$.

![Normalized MSE and achievable rate ratio versus transmission power for $v = 10$ m/s and $\eta = 0.1$.](image)

**Fig. 4.** Normalized MSE and achievable rate ratio versus transmission power for $v = 10$ m/s and $\eta = 0.1$.

![Normalized MSE and achievable rate ratio versus transmission power for $M$ for $\eta = 0.1$ and 20 dBm transmission power.](image)

**Fig. 5.** Normalized MSE and achievable rate ratio versus $M$ for $\eta = 0.1$ and 20 dBm transmission power.

Fig. 5 displays the performance of the proposed single-path CE and DSA methods. The proposed CE yields near-optimal achievable rate as manifested by the right plot, and the simple DSA works effectively by comparing the performance between moving velocities of 0 and 10 m/s. Note that the MSE and achievable rate degrade noticeably when $M$ reaches 64, which is again attributed to high sensitivity to phase changes.

VI. CONCLUSION

In this letter, we have proposed a practical DSA protocol to improve CE accuracy for RIS-assisted wideband wireless systems, and a low-complexity CE scheme and simplified DSA for single-path scenarios. Simulation results have corroborated the validity and effectiveness of our proposed approaches.

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