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Fast Reconfiguration of Liquid Crystal-RISs: Modeling and Algorithm Design

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Abstract—liquid crystal (LC) technology is a promising hardware solution for realizing extremely large reconfigurable intelligent surfaces (RISs) due to its advantages in cost-effectiveness, scalability, energy efficiency, and continuous phase shift tunability. However, the slow response time of the liquid crystal (LC)-RIS phase shifters, especially in comparison to the silicon-based alternatives like radio frequency switches and positive-intrinsic-negative (PIN) diodes, limits the performance. This limitation becomes particularly relevant in time-division multiple-access (TDMA) applications where RIS must sequentially serve users in different locations, as the phase-shifting response time of LC-RIS phase shifters can constrain system performance. This paper addresses the slow phase-shifting limitation of LC by developing a physics-based model for the time response of an LC unit cell and proposing a novel phase-shift design framework to reduce the transition time. Specifically, exploiting the fact that LC-RIS at millimeter wave (mmWave) bands have a large number of elements, we optimize the LC phase shifts based on user locations, eliminating the need for full channel state information (CSI) and minimizing reconfiguration overhead. Moreover, instead of focusing on a single point, the RIS phase shifters are designed to optimize coverage over an area. This enhances communication reliability for mobile users and mitigates performance degradation due to user location estimation errors. The proposed RIS phase-shift design minimizes the transition time between configurations, a

critical requirement for TDMA schemes. Our analysis reveals that the impact of RIS reconfiguration time on system throughput becomes particularly significant when TDMA intervals are comparable to the reconfiguration time. In such scenarios, optimizing the phase-shift design helps mitigate performance degradation while ensuring specific quality of service requirements. Moreover, the proposed algorithm has been tested through experimental evaluations, which demonstrate that it also performs effectively in practice.

Index Terms—Liquid crystal, reconfigurable intelligent surfaces, tuning time, fast reconfiguration.

I. INTRODUCTION

RECONFIGURABLE intelligent surfaces (RISs) have recently emerged as a solution to realize programmable radio environments in the field of wireless communications [2], [3], [4]. reconfigurable intelligent surface (RIS)s comprise a large array of sub-wavelength elements capable of dynamically adjusting the phases of reflected electromagnetic waves. By intelligently controlling these phase shifts, RISs establish virtual links between base station (BS) and mobile user (MU), when the direct line of sight (LOS) BS-MU link is obstructed [5], [6]. These RISs require a large number of elements to compensate for the significant path loss caused by double-path reflection [5]. Such electrically large surfaces introduce several unique challenges, including the need for scalable low-cost hardware architectures, low-complexity low-overhead CSI acquisition and RIS configuration schemes, and accurate channel models that account for near-field (NF) propagation regime.

Several technologies exist for implementing RIS phase shifters, including varactor diodes [7], [8], PIN diodes [9], [10], [11], micro-electro-mechanical system (MEMS) [12], [13], and LC. Each technology presents a unique trade-off between performance, cost, and scalability. For instance, while PIN diodes and MEMS offer very fast response times, they typically provide only discrete phase shifts. Achieving finer phase resolution requires multiple diodes per element, which increases both cost and complexity, hindering scalability, especially at mmWave frequencies where the number of RIS elements is large. Varactor diodes, while offering continuous tuning, are generally limited to lower frequency ranges and are less effective for mmWave applications. Collectively, these limitations challenge the scalability of silicon-based technologies for future mmWave and sub-terahertz (THz)

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systems. In contrast, LC is notable for its cost-effectiveness, scalability, energy efficiency, and continuous phase tunability [14]. Despite these advantages, the adoption of LC-based RISs is constrained by their relatively slow phase response times compared to alternative technologies such as PIN diodes, MEMS, and varactor diodes [8] (See [15, Table I] for a comprehensive comparison of these technologies).

Given their unique advantages, this paper focuses on LC-RISs. We note that their relatively slow response time confines their primary application domains to quasi-static environments, where they can be leveraged to create stable and favorable propagation paths that overcome blockages or extend coverage.¹ Examples of quasi-static radio environments include fixed wireless access and for serving low-to-moderate mobility users in indoor hotspots. Even in these targeted applications, the reconfiguration latency remains a critical performance bottleneck, since the transition period represents an overhead during which users cannot be effectively served. In a TDMA system, for instance, reducing this overhead translates directly into higher effective data rates. Addressing this fundamental hardware limitation is therefore essential for realizing the full potential of LC-RISs, and constitutes the central focus of this paper.

The potential of LC-RISs to enhance wireless channels remains relatively underexplored within the wireless communications community. Existing studies have investigated RISs implemented with LC technology from several perspectives [16], [17], [18], [19], [20], [21], [22], [23]. For instance, [18], [19] designed LC-based RISs specifically for visible light communication systems. From an experimental viewpoint, the behavior of LC-RISs was analyzed in [20], whereas the physical mechanisms influencing the LC response time were explored in detail in [21] and [22]. Additionally, a comprehensive overview of LC-RIS properties was presented in [15].

In this paper, by investigating the physical dynamics of LC molecules, we develop a comprehensive model for their response time. Exploiting this model, we formulate an optimization problem to design RIS configurations that minimize the transition time between users. By providing a precise model and optimization framework for this fundamental hardware latency, our work provides a critical physical-layer foundation that enables more practical and efficient system integration and protocol design [24]. Our main contributions are summarized as follows:

- **Physics-based Modeling of LC-RIS Phase-shift Response Time:** We develop an analytical model for the response time of an LC-based phase shifter, rooted in the physical properties of its constituent LC molecules. Specifically, we study the partial differential equation (PDE) that governs the spatio-temporal evolution of LC

molecules upon applying an electric field. Using this, we develop a model for the phase shift of each LC-RIS unit cell as a function of t .

- **Fast-transitionable Phase-shift Design:** We formulate an optimization problem to design multiple RIS phase-shift configurations that enable fast transitions between them. As the CSI acquisition is a challenging task for RISs [24], we formulate the phase-shift design problem based on user locations. The accuracy of this design increases in mmWave since LOS component becomes dominant [25]. Additionally, to increase the communication reliability against the location estimation error, the RIS is designed to serve users over a designated area rather than a single point.
- **Proposed Algorithm:** Because this problem is non-convex, finding the global optimal solution is computationally challenging. Based on the dual Lagrangian formulation and exploiting the large number of the LC-RIS elements, we propose an approach to decouple the optimization across users and RIS elements, enabling parallel and iterative updates. Based on this result, we propose an algorithm that obtains a sub-optimal, but low-complexity, and scalable solution. We analyse the complexity of the proposed algorithm and show that it increases linearly with the number of LC-RIS elements.
- **Performance Evaluation:** We assess the performance of the proposed RIS design through a comprehensive set of numerical simulations. These results demonstrate that the proposed approach significantly reduces reconfiguration time (e.g., by 66% for the considered setup), thereby improving the effective system throughput compared to the transition-unaware benchmark schemes.
- **Experimental Verification:** Finally, we validate our algorithm through a real-world test in an indoor scenario [26]. We observe that the optimized phase shift design significantly reduces the required transition time compared to a linear phase shift design algorithm, which only focuses on maximizing the final received signal to noise ratio (SNR).

The remainder of this paper is organized as follows. In Section II, we present the system, channel, and steady state of LC models. Section III details the proposed optimization framework based on the dynamics LC models, followed by simulation results in Section IV. Then, we test our algorithm in an experimental implementation in Section V. Finally, Section VI concludes the paper.

Notation: Bold uppercase and lowercase letters are used to denote matrices and vectors, respectively. $(\cdot)^T$ and $(\cdot)^H$ denote the transpose and Hermitian, respectively. Moreover, $\mathbf{0}_n$ and $\mathbf{1}_n$ denote a column vectors of size n whose elements are all zeros and ones, respectively. $\mathcal{O}$ is the big-O notation. $[\mathbf{X}]_{m,n}$ and $[\mathbf{x}]_n$ denote the element in the m th row and n th column of matrix $\mathbf{X}$ and the n th entry of vector $\mathbf{x}$, respectively. $\mathcal{CN}(\boldsymbol{\mu}, \boldsymbol{\Sigma})$ denotes a complex Gaussian random vector with mean vector $\boldsymbol{\mu}$ and covariance matrix $\boldsymbol{\Sigma}$. $x!$ denotes factorial of number x . $\mathbb{E}\{\cdot\}$ and $\mathbb{V}\{\cdot\}$ represent expectation and variance,

¹Due to their slow response, LC-RISs are not suitable for compensating fast-fading variations at the microsecond timescale typically associated with resource block lengths. Consequently, a location-based illumination approach, as considered in this paper, provides a practical and effective strategy for exploiting the benefits of LC-RISs in quasi-static environments. It is also worth noting that even for faster RIS technologies (e.g., PIN-diode-based designs), tracking fast fading remains highly challenging because of the prohibitive overhead associated with CSI acquisition.

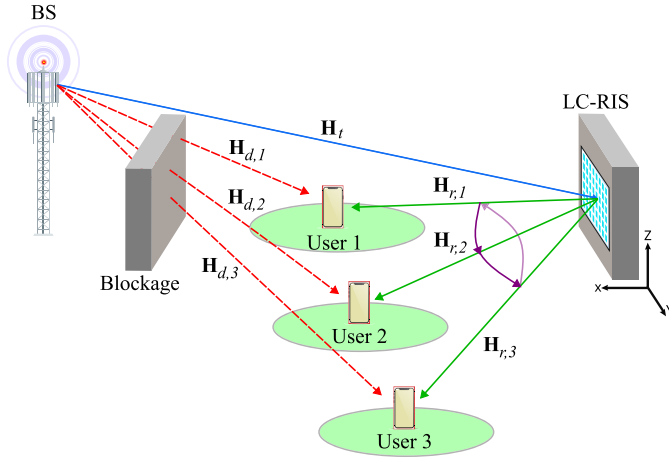


Fig. 1. An RIS assists to establish a virtual link between a transmitter and multiple receivers in a time-sequential manner, while the direct channel is blocked by an obstacle.

respectively. Finally, $\mathbb{Z}$, $\mathbb{R}$, and $\mathbb{C}$ represent the sets of integer, real, and complex numbers, respectively.

II. SYSTEM, CHANNEL, AND LC-RIS CELL MODEL

In this section, we first present the considered system model. Subsequently, we introduce the adopted channel model and steady-state model of LC phase shifters.

A. System Model

We consider a narrow-band downlink communication scenario comprising a BS with N_t antenna elements, an RIS with N LC-based unit cells, and K single-antenna MUs as illustrated in Fig. 1. The users are served in a TDMA scheme.² The received signal at user k in their allocated time slot is given by:

$$y_k = (\mathbf{h}_{d,k}^H + \mathbf{h}_{r,k}^H \mathbf{\Gamma}_k \mathbf{H}_t) \mathbf{x}_k + n_k, \quad k = 1, \dots, K, \quad (1)$$

where $\mathbf{x}_k \in \mathbb{C}^{N_t}$, $\forall k$ is the transmit signal vector, $y_k \in \mathbb{C}$ is the received signal vector at the k th MU, and $n_k \in \mathbb{C}$ represents the additive white Gaussian noise (AWGN) at the k th MU, i.e., $n_k \sim \mathcal{CN}(0, \sigma_n^2)$, where σ_n^2 is the noise power. Assuming linear beamforming, the transmit vector $\mathbf{x}_k \in \mathbb{C}^{N_t}$ can be written as $\mathbf{x}_k = \mathbf{q}_k s$, where $\mathbf{q}_k \in \mathbb{C}^{N_t}$ is the beamforming vector for user k and $s \in \mathbb{C}$ is the data symbol. Assuming $\mathbb{E}\{|s|^2\} = 1$, the beamformer satisfies the transmit power constraint $\sum_{k=1}^K \|\mathbf{q}_k\|^2 \leq P_t$, where P_t is the total power constraint. Moreover, $\mathbf{h}_{d,k} \in \mathbb{C}^{N_t}$, $\mathbf{H}_t \in \mathbb{C}^{N \times N_t}$, and $\mathbf{h}_{r,k} \in \mathbb{C}^N$ denote the BS-UE, BS-RIS, and RIS-MU channel matrices, respectively. Furthermore, $\mathbf{\Gamma}_k \in \mathbb{C}^{N \times N}$ is a diagonal matrix with main diagonal entries $[\mathbf{\Gamma}_k]_n = [\Omega_k]_n e^{j[\omega_k]_n}$ denoting the reflection coefficient applied by the n th RIS unit cell comprising phase shift $[\omega_k]_n$ and reflection amplitude

²We adopt a simple setting consisting of a single cell and a TDMA scheme for serving single-antenna users in order to focus primarily on the LC-RIS transition time between different phase-shift configurations. Nevertheless, the model proposed in Section III-A for LC-RIS time response is valid for more sophisticated scenarios and the algorithm proposed in Section III-B can be extended to other multiple-access schemes and multiple-antennas users [27].

$[\Omega_k]_n$ when the RIS serves k th user in its allocated time slot. We assume throughout the paper, $[\Omega_k]_n = 1$, $\forall n$ holds for LC-RIS cells regardless of k [28] and we can control the phase shifts of each cell, $[\omega_k]_n$.

B. Channel Model

Given the assumption of an extremely large RIS, the distances between the RIS and both the BS and the MU may fall within the NF regime of the RIS [29], [30]. Therefore, an NF channel model is adopted. Furthermore, RISs are typically installed at elevated heights, ensuring LOS links between the RIS and both the BS and MUs. High-frequency further emphasizes the dominance of these LOS links over non-LOS (nLOS) links. As a result, the channels are modeled using Rician fading with a high K -factor, reflecting the significant contribution of LOS components relative to nLOS components. For the sake of simplicity, we present the model of general $\mathbf{H} \in \mathbb{C}^{N_{rx} \times N_{tx}}$ where N_{tx} and N_{rx} are the number of transmit and receive antennas, respectively.

A Rician multiple-input multiple-output (MIMO) channel model can be written as

$$\mathbf{H} = \sqrt{\frac{K_f}{K_f + 1}} \mathbf{H}^{\text{LOS}} + \sqrt{\frac{1}{K_f + 1}} \mathbf{H}^{\text{nLOS}}, \quad (2)$$

where K_f denotes the K -factor and determines the relative power of the LOS component to the nLOS components of the channel. In NF regime, $\mathbf{H}^{\text{LOS}}$ and $\mathbf{H}^{\text{nLOS}}$ are functions of the locations of antennas in order to capture the underlying spherical wave propagation. This leads to [29]

$$cc[\mathbf{H}^{\text{LOS}}]_{m,n} = c_0 e^{j\kappa \|\mathbf{u}_{rx,m} - \mathbf{u}_{tx,n}\|}, \quad (3)$$

$$\mathbf{H}_s^{\text{nLOS}} = c_s \mathbf{a}_{rx}(\mathbf{u}_s) \mathbf{a}_{tx}^T(\mathbf{u}_s), \quad (4)$$

$$[\mathbf{a}_{tx}(\mathbf{u}_s)]_n = e^{j\kappa \|\mathbf{u}_{tx,n} - \mathbf{u}_s\|} \text{ and } [\mathbf{a}_{rx}(\mathbf{u}_s)]_m = e^{j\kappa \|\mathbf{u}_{rx,m} - \mathbf{u}_s\|}, \quad (5)$$

where $\mathbf{H}^{\text{LOS}}$ denotes the LOS NF channel matrix and is only a function of antenna locations in transmitter (Tx) and receiver (Rx), c_0 represents the channel amplitude of the LOS path, and $\mathbf{u}_{tx,n}$ and $\mathbf{u}_{rx,m}$ are the locations of the n th Tx antenna and the m th Rx antenna, respectively. Moreover, $\kappa = 2\pi/\lambda$ is the wave number with λ being the wavelength, $\mathbf{a}_{tx}(\cdot) \in \mathbb{C}^{N_{tx}}$ and $\mathbf{a}_{rx}(\cdot) \in \mathbb{C}^{N_{rx}}$ denote the Tx and Rx NF array response, respectively, $\mathbf{u}_s$ is the location of the s th scatterer, and c_s denotes the end-to-end amplitude of the s th non-LOS path [31].

For the LOS link between the BS and the MUs, we incorporate an additional high penetration loss due to a blockage, as suggested in [32]. Consequently, in the optimization section, we assume $\mathbf{h}_{d,k} \approx \mathbf{0}$ for all k to simplify the analytical analysis while we account for its impact in the simulation results in Section IV.

C. Steady-State Model of LC Phase Shifter

The anisotropy of LC molecules allows the control of the phase shift of radio frequency (RF) signals by adjusting the orientation of the LC molecules. To achieve this effect, a thin LC layer, typically on the order of micrometers, is

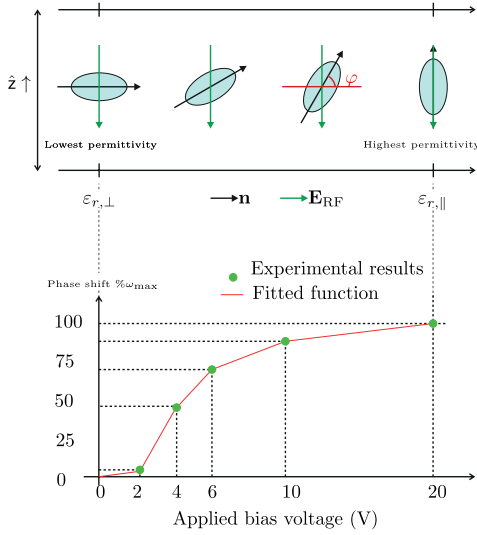


Fig. 2. Phase shift vs. applied voltage for LC with $4.6 \mu\text{m}$ LC layer thickness. The experimental data is taken from [26], being fitted by a piecewise linear function.

placed between two electrodes. As illustrated in Fig. 2, in the absence of an applied voltage (a low kHz alternating current (AC) voltage for biasing), i.e., $E = 0$, the LC molecules remain in a relaxed state. As a result, the radio-frequency electric field $\mathbf{E}_{\text{RF}}$ is perpendicular to the unit vector $\mathbf{n}$ ($\|\mathbf{n}\| = 1$), representing the average molecular alignment in a nematic LC.³ This leads to a minimum permittivity, denoted by $\varepsilon_{\perp} = \varepsilon_{r,\perp}\varepsilon_0$, where $\varepsilon_{r,\perp}$ and ε_0 are minimum relative and vacuum permittivities, respectively. Conversely, under maximum applied voltage ($E = E_{\text{max}}$), most of the LC molecules align with the induced external electric field. This alignment causes $\mathbf{E}_{\text{RF}}$ to become parallel to $\mathbf{n}$, resulting in achieving maximum permittivity, denoted by $\varepsilon_{\parallel} = \varepsilon_{r,\parallel}\varepsilon_0$, where $\varepsilon_{r,\parallel}$ is maximum relative permittivity. The maximum achievable phase shift, $\Delta\omega_{\text{max}}$, is proportional to the maximum differential wave number in birefringent⁴ materials, e.g., LCs, denoted as $\Delta\kappa_{\text{max}}$, and can be expressed as:

$$\Delta\omega_{\text{max}} = \ell_{\text{lc}}\Delta\kappa_{\text{max}}, \quad (6)$$

where ℓ_{lc} is the length of the phase shifter. Here, $\Delta\kappa_{\text{max}}$ is defined as $\kappa_{\parallel} - \kappa_{\perp}$, where the wave number κ_g is given by $2\pi f\sqrt{\varepsilon_g\mu_g}$, $g \in \{\perp, \parallel\}$. Here, f and μ_g represent frequency and permeability, respectively. For LC materials, $\mu_{\perp} = \mu_{\parallel} = \mu_0$ typically holds [33], where μ_0 is vacuum permeability. Assuming the phase shift corresponding to $\kappa_{\perp}$ as the reference

³The nematic phase is a state of LCs where the rod-like molecules exhibit long-range orientational order (i.e., they tend to point in the same general direction) but lack long-range positional order. Crucially for RIS applications, this collective orientation can be manipulated by applying an external electric field ($\mathbf{E}_{\text{RF}}$).

⁴Birefringence is the property of a material where its permittivity depends on the polarization of the EM wave relative to the material's structure. For LCs, this means the permittivity experienced by a wave is different when it is polarized parallel (high permittivity) versus perpendicular (low permittivity) to the long axis of the LC molecules, which is the key mechanism for tuning the phase shift.

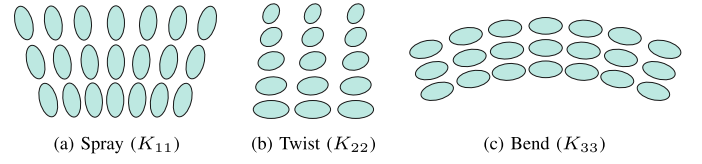


Fig. 3. Demonstrations of three basic deformations of LC.

(i.e., $\omega = 0$), the effective phase shift that each cell can apply in addition to the reference phase shift is given by:

$$\omega = \ell_{\text{lc}}(\kappa - \kappa_{\perp}), \quad (7)$$

where $\kappa_{\perp} \leq \kappa \leq \kappa_{\parallel}$. The temporal evolution of κ and ω is analyzed in the next section, where their dynamic behavior is discussed in detail.

III. TRANSITION-AWARE LC-RIS PHASE-SHIFT DESIGN

In this section, we begin by deriving the relationship between the LC phase shifter and time using the underlying fundamental physics principles. Next, using this model, we formulate an optimization that minimizes the transition time among phase-shift configurations and develop an efficient low-complexity solution to this problem.

A. Transient Dynamic Model of LC Phase Shifter

The operational principle of an LC molecule is based on its electromagnetic anisotropy, wherein the electromagnetic characteristics of the LC molecule depend on its orientation relative to the RF electric field $\mathbf{E}_{\text{RF}}$ [15]. Due to its ellipsoidal molecular shape, an LC molecule exhibits higher permittivity, i.e., a greater phase shift, when the electric field $\mathbf{E}_{\text{RF}}$ is aligned with the molecule's major axis $\mathbf{n}$, compared to when it is aligned with the molecule's minor axis, as depicted in Fig. 2. The time required to achieve a certain angle (vector $\mathbf{n}$ with respect to (w.r.t.) $\mathbf{E}_{\text{RF}}$) depends on the applied electrically induced force at the electrodes as well as the mechanical characteristics of the LC molecules. In the following, we first recap the fundamental properties of the LC molecules which is the basis for derivation of the proposed LC response time model in Section III-A.4.

1) *Three-Dimensional System*: Consider LC molecules with a unit vector $\mathbf{n} = [n_x, n_y, n_z]^T$. The time evolution of this director field $\mathbf{n}$ can be described by the principle of dissipative dynamics. This principle states that the rate of change of the system is proportional to the total forces (or torques) arising from the gradient of the free energy, balanced by viscous dissipation [34]:

$$\gamma_1 \frac{\partial \mathbf{n}}{\partial t} = - \frac{\delta F}{\delta \mathbf{n}} + \lambda_{\text{lc}} \mathbf{n}. \quad (8)$$

Here, $\frac{\partial \mathbf{n}}{\partial t}$ is the time derivative of the director field, and $-\frac{\delta F}{\delta \mathbf{n}}$ is the functional derivative of the free energy w.r.t. $\mathbf{n}$, which gives the torque acting on $\mathbf{n}$. γ_1 and λ_{lc} are rotational viscosity coefficient quantifying resistance to director motion, and Lagrange multiplier enforcing the unit-vector constraint $\|\mathbf{n}\| = 1$, respectively. Assuming no fluid flow and ignoring backflow and inertial effects [22], the total free energy F of

the nematic LC is the integral over the volume V representing the volume of LC molecules:

$$F = \int_V f_{\text{elastic}} + f_{\text{electric}} dv, \quad (9)$$

where f_{elastic} and f_{electric} are the Frank elastic energy density and the electric energy density, respectively. The Frank elastic energy density captures the cost of deforming the director field $\mathbf{n}$ as follows [35],

$$f_{\text{elastic}} = \frac{1}{2} K_{11} \|\nabla \cdot \mathbf{n}\|^2 + \frac{1}{2} K_{22} \|\mathbf{n} \cdot \nabla \times \mathbf{n}\|^2 + \frac{1}{2} K_{33} \|\mathbf{n} \times \nabla \times \mathbf{n}\|^2, \quad (10)$$

where K_{11} , K_{22} , and K_{33} are the elastic constants associated with splay, twist, and bend deformation, respectively, see Fig. 3. When an electric field $\mathbf{E} = [E_x, E_y, E_z]^T$ is applied, it interacts with the LC molecule, aligning it in the direction of the field. The electric energy density is:

$$f_{\text{electric}} = -\frac{1}{2} \mathbf{E}^T \cdot \boldsymbol{\varepsilon} \cdot \mathbf{E}, \quad (11)$$

where $\boldsymbol{\varepsilon}$ is the dielectric permittivity tensor, and for a uniaxial LC, $\boldsymbol{\varepsilon}$ can be expressed as:

$$\boldsymbol{\varepsilon} = \varepsilon_{\perp} \mathbf{I} + \underbrace{(\varepsilon_{\parallel} - \varepsilon_{\perp})}_{\Delta\varepsilon} \mathbf{n} \mathbf{n}^T. \quad (12)$$

Here, $\mathbf{I}$ denotes an identity matrix. Substituting (10) and (11) in (8) has in general a complicated form. However, for the LC-RIS consisting of LC molecules encapsulated between two large planes can be simplified to an one-dimensional PDE. In the following, we focus on a single spatial dimension ($\bar{z}$) by assuming uniformity in the other dimensions, thereby simplifying the problem while retaining the essential characteristics.

2) *One-Dimensional System*: Consider an electric field applied along the z -axis, with LC molecules rotating w.r.t. this direction. Substituting the one dimension of Frank elastic energy density in (10) and the electric energy density in (11) into (8) yields the following equation [22], [36], [37]:

$$\begin{aligned} \gamma_1 \frac{\partial \varphi}{\partial t} &= (K_{11} \cos^2 \varphi + K_{33} \sin^2 \varphi) \frac{\partial^2 \varphi}{\partial z^2} \\ &+ (K_{33} - K_{11}) \sin \varphi \times \cos \varphi \left(\frac{\partial \varphi}{\partial z} \right)^2 \\ &+ \varepsilon_0 \Delta\varepsilon E^2 \sin \varphi \cos \varphi, \end{aligned} \quad (13)$$

where E is the applied electrical field in z direction and φ is the angle of between major axis of the LC molecule and $x - y$ plane, see Fig. 2. Even in this simplified form, (13) is not straightforward to be solved analytically, which we need to draw insights for system design. To address this issue, we adopt two assumptions, namely $K_{11} = K_{33} = \bar{K}$ [22] and

$$\sin \varphi \cos \varphi \approx \Phi(\varphi) \triangleq \varphi \left(1 - \frac{\varphi^2}{2} \right), \quad 0 \leq \varphi \leq \frac{\pi}{2}, \quad (14)$$

reducing it to:

$$\gamma_1 \frac{\partial \varphi}{\partial t} = \bar{K} \frac{\partial^2 \varphi}{\partial z^2} + \varepsilon_0 \Delta\varepsilon E^2 \Phi(\varphi). \quad (15)$$

The PDE in (15) governs the spatio-temporal evolution of the LC molecule alignment under an applied electric field. We will

use this equation in Section III-A.4 as the basic to develop a model for the phase shift response time of LC-RISs, which is then used to design a fast reconfigurable phase shift in TDMA application. But before then, in order to fully describe φ as a function of t and z , appropriate boundary conditions are required, which will be discussed in the following section.

3) *Boundary and Initial Conditions and Solution*: Typically, LC-RISs consist of parallel planes with an LC layer positioned between them. To solve (15), we require boundary and initial conditions. Consider two planes located at $z = 0$ and $z = d$, where d is the LC-RIS phase shifter thickness. Due to anchoring forces, the LC molecules near these planes satisfy $\varphi \approx 0$. Taking into account the initial state, the angle $\varphi(z, t)$ can be described as the solution to the PDE in (15), subject to its boundary and initial conditions as follows:

$$\text{C1: } \varphi(0, t) = 0, \quad (16a)$$

$$\text{C2: } \varphi(d, t) = 0, \quad (16b)$$

$$\text{C3: } \varphi(z, 0) = A_1 \sin\left(\frac{\pi z}{d}\right), \quad (16c)$$

where C3 has been confirmed through experimental observations, as detailed in [22] and [38] and $0 < A_1 < 1$ is a constant. The solutions of this PDE in two extreme cases $E = 0$ and $E = E_{\text{max}}$ are provided in Lemma 1.

Lemma 1: The solutions to (15), subject to the constraints C1, C2, and C3 at two extreme cases $E = 0$ and $E = E_{\text{max}}$, can be expressed as

$$\varphi(z, t) = \varphi(z, \infty) + (\varphi(z, 0) - \varphi(z, \infty)) e^{-\frac{t}{\tau_{\text{mol}}}}, \quad (17)$$

with an error bound of $\mathcal{O}(e)$ with assuming $e \triangleq E_{\text{max}}^{-2} \frac{\pi^2 \bar{K}}{d^2 \varepsilon_0 \Delta\varepsilon} \ll 1$. Here, the LC molecule exponential time constant $\tau_{\text{mol}} \in \{\tau_{\text{mol}}^-, \tau_{\text{mol}}^+\}$ depends on whether φ is increasing or decreasing, where $\tau_{\text{mol}}^- \gg \tau_{\text{mol}}^+$ holds.

Proof: The proof is provided in Appendix A. $\square$

The assumption $e \ll 1$ is valid for typical physical parameters [39]. We have shown that the average angle of the LC molecules φ evolves as an exponential function of time at two extreme cases. Based on this result, the next section explores in detail the time response characteristics of the LC phase shifter.

4) *LC Phase Shifter Response Time*: In Proposition 1, we establish that while φ follows an exponential function w.r.t. time, ω can be expressed as a summation of multiple exponential terms over time.

Proposition 1: Based on the result of the Lemma 1, the LC phase shifter ω evolves as a summation of exponential functions over time, given by

$$\omega(t) = \ell_{1c} \sum_{p=0}^{\infty} D_p e^{-\frac{pt}{\tau_{\text{mol}}}}, \quad (18)$$

where D_p are constant coefficients w.r.t. time.

Proof: The proof is provided in Appendix B. $\square$

There are two important observations about the summation in (18), which we use to develop a simple approximation:

- The term $e^{-\frac{pt}{\tau_{\text{mol}}}}$ decays more rapidly for larger p .

- The coefficients D_p tend to decrease as p increases due to the behavior of (59).

Motivated by these insights, we approximate (18) with a single exponential function with an exponent that is not necessarily identical to τ_{mol} and can be found to improve the approximation accuracy by fitting it to measurement data. The resulting model for the LC-RIS time response is parameterized as follows:

$$[\omega(t)]_n = [\omega_d]_n + ([\omega_0]_n - [\omega_d]_n)e^{\frac{-t}{[\tau_c]_n}}, \quad (19)$$

where $[\omega_d]_n$ and $[\omega_0]_n$ are the desired and initial phase shift applied in n th LC-RIS, respectively. In addition, $[\tau_c]_n \in \{\tau_c^+, \tau_c^-\}$ is the LC director reorientation time constant, which is a function of LC parameters γ , $\bar{K}$, and E . It is equal to τ_c^+ if $[\omega_d]_n \geq [\omega_0]_n$ and τ_c^- if $[\omega_d]_n < [\omega_0]_n$ [22]. To enhance the accuracy of our proposed model, we incorporate experimentally obtained values for the exponent factor. The experimental values for τ_c^+ and τ_c^- have been reported in [26], see Fig. 4. This figure suggests that the proposed model with tuned parameters can indeed follow the experimental data.

5) *Over-and Undershooting Technique*: In order to accelerate the time response of *each individual* LC-RIS phase shifter, we can exploit so-called techniques named over- and undershooting [40]. To do that, when the phase of the LC must increase, i.e., $[\omega_d]_n > [\omega_0]_n$ the V_{max} is applied to the corresponding cell. In contrast, when it needs to be decreased, i.e., $[\omega_d]_n < [\omega_0]_n$, zero voltage which is related to relaxed state is applied. In both cases, we change the voltage control to the $[v_d]_n$ when the phase reaches the desired value. The required time for n th cell to reach its desired phase shift by exploiting this technique can be reduced as follows

$$[t_r]_n = \tau_c^+ \ln \left(\frac{\omega_{\text{max}} - [\omega_0]_n}{\omega_{\text{max}} - [\omega_d]_n} \right) = \tau_c^+ \ln \left(1 + \frac{[\omega_d]_n - [\omega_0]_n}{\omega_{\text{max}} - [\omega_d]_n} \right), \quad (20)$$

where $[t_r]_n$ determines the required time for n cell to reach its desired phase. Similarly, to decrease the phase shift, i.e., $[\omega_d]_n < [\omega_0]_n$, we first set the voltage to zero using the undershooting technique so the required time is obtained as

$$[t_r]_n = \tau_c^- \ln \left(\frac{[\omega_0]_n - \omega_{\text{min}}}{[\omega_d]_n - \omega_{\text{min}}} \right) = \tau_c^- \ln \left(1 + \frac{[\omega_0]_n - [\omega_d]_n}{[\omega_d]_n - \omega_{\text{min}}} \right), \quad (21)$$

where we usually assume $\omega_{\text{min}} = 0$.

B. Problem Formulation and the Proposed Solution

In the following, we employ the derived model in Section III-A.5 for the LC-RIS time response in order to develop phase shifters that enable fast reconfigurations. We consider a TDMA setup where, in each time slot, the RIS is configured to serve an area that includes the user, rather than focusing on a single point as in [1]. Serving an area instead of a specific point improves communication reliability because the RIS can maintain the required quality of service (QoS) despite user movement within the area or in the presence of location estimation errors [41, Fig. 3]. Achieving this objective necessitates deriving multiple phase-shift configurations that not only maximize the minimum SNR across the target area but also ensure that the time required to switch between

configurations remains minimal. We define SNR of user k as follows

$$\text{SNR}_k(\mathbf{u}_k) = \frac{|\mathbf{h}_k^H(\mathbf{u}_k)\mathbf{q}_k|^2}{\sigma_n^2}, \quad (22)$$

where $\mathbf{h}_k^H(\mathbf{u}_k) = \mathbf{h}_{r,k}^H(\mathbf{u}_k)\mathbf{\Gamma}_k\mathbf{H}_t$ represents the end-to-end channel at the $\mathbf{u}_k \in \mathcal{U}_k$ location, $\mathcal{U}_k$ denotes the set of possible locations of user k where $k = 1, \dots, K$, and $\mathbf{q}_k$ denotes the beamforming vector in BS for user k . Since user location data may be imprecise, we define the QoS metric as:

$$\text{SNR}_k^{\min} \triangleq \min_{\mathbf{u}_k \in \mathcal{U}_k} \frac{|\mathbf{h}_k^H(\mathbf{u}_k)\mathbf{q}_k|^2}{\sigma_n^2}. \quad (23)$$

The transition time for the n th cell from $[\omega_0]_n$ to $[\omega_d]_n$ is determined by the differential between these two phase shifts. Our objective is to achieve phase shifts that satisfy the required SNR, γ_k^{thr} , for any user location within the target area, while minimizing the time needed for reconfiguration. Assuming a given order for the serving the users,⁵ we formulate the following optimization problem

$$\text{P1: } \min_{\omega_k, \mathbf{q}_k, \forall k} \sum_{k=1}^K \max_n [t_k]_n \quad (24a)$$

$$\text{s.t. } \frac{|\mathbf{h}_k^H(\mathbf{u}_k)\mathbf{q}_k|^2}{\sigma_n^2} \geq \gamma_k^{\text{thr}}, \forall \mathbf{u}_k \in \mathcal{U}_k, \forall k, \quad (24b)$$

$$0 \leq [\omega_k]_n < \omega_{\text{max}}, \forall n, k, \quad (24c)$$

$$\sum_{k=1}^K \|\mathbf{q}_k\|_2^2 \leq P_t, \forall k, \quad (24d)$$

where each $[t_k]_n, \forall k$ is defined in (20) and (21) w.r.t. current user k and previous user $k-1$. $\omega_k \in \mathbb{C}^N$ and $\mathbf{q}_k \in \mathbb{C}^{N_t}$ are phase shift and beamforming variables, respectively. Here, (24b) is the minimum SNR_k constraint for k th user to be served by RIS. (24c) is the realizable phase shift range where ω_{max} shows the maximum achievable phase shift by each LC element in the RIS. (24d) forces the transmit power in the BS. The problem P1 is a non-convex problem due to the non-convexity of the objective function and constraint (24b). Two variable vectors ω_k and $\mathbf{q}_k$ are also coupled in (24b). Hence, obtaining the global solution for this problem is challenging in term of complexity. Due to these challenges, we adopt alternating optimization (AO), where the problem is separated into two sub-problems where in each one, we fix one of the variable vectors and optimize the problem w.r.t. the other variable vector.

Beamformer Design:

For a fixed RIS phase shift $\omega_k, \forall k$, the problem P1 reduces to a feasibility search for the beamformer $\mathbf{q}_k$. One approach to obtain a feasible solution is to choose to maximize SNR in $\mathbf{q}_k$ for the given ω_k . By doing so, we create the largest possible SNR margin above the required threshold γ_k^{thr} . This SNR margin is beneficial in the next step of the AO, where $\mathbf{q}_k$ is fixed and ω_k is optimized. A larger margin means that the SNR constraint remains satisfied for a wider range of ω_k

⁵In a multi-user scenario where $K > 2$, the order in which users are served affects the overall reconfiguration process, i.e., there are $(K-1)!$ possible permutations. In section III-B.2, we investigate the impact of user order through simulations.

values, thus hopefully expanding the feasible set for the RIS phase-shift optimization (there is no guarantee). This provides the algorithm with more flexibility to find a ω_k that minimizes the primary objective, which is the transition time. Hence, $\forall k = \{1, \dots, K\}$, the following problem is solved:

$$\text{P2: } \max_{\mathbf{q}_k} \text{SNR}_k^{\min} \quad (25a)$$

$$\sum_{k=1}^K \|\mathbf{q}_k\|_2^2 \leq P_t, \quad (25b)$$

where $\text{SNR}_k^{\min}$ was defined in (23). Under the assumptions of blockage, i.e., $\mathbf{h}_{d,k} \approx \mathbf{0}$, $\forall k$, and dominant LOS BS-RIS link, i.e., $K_f \rightarrow +\infty$, we can rewrite channel as follows:

$$\mathbf{h}_k(\mathbf{u}_k) = \mathbf{h}_{r,k}^H \mathbf{\Gamma}_k \mathbf{a}_{\text{RIS}}(\mathbf{u}_{\text{BS}}) \mathbf{a}_{\text{BS}}^H(\mathbf{u}_{\text{RIS}}), \quad (26)$$

where $\mathbf{a}_{\text{BS}}(\cdot)$ and $\mathbf{a}_{\text{RIS}}(\cdot)$ are steering vectors in BS toward the RIS and RIS toward the BS, respectively. They were defined in (5) by substituting $\mathbf{u}_s$ with $\mathbf{u}_{\text{RIS}}$ and $\mathbf{u}_{\text{BS}}$ which are the locations of the RIS and BS. This is now a standard SNR maximization problem in multi-input single-output (MISO) systems, which has the following matched-filter precoder solution [42]:

$$\mathbf{q}_k \stackrel{(a)}{\approx} \frac{\sqrt{P_t}}{\|\mathbf{a}_{\text{BS}}(\mathbf{u}_{\text{RIS}})\|} \mathbf{a}_{\text{BS}}(\mathbf{u}_{\text{RIS}}). \quad (27)$$

This problem is feasible when $\min_{\mathbf{u}_k \in \mathcal{U}_k} P_t \frac{\|\mathbf{h}_k(\mathbf{u}_k)\|^2}{\sigma_n^2} \geq \gamma_k^{\text{thr}}$, $\forall k$. Because the chosen beamformer does not depend on the RIS phase shifter, we denote this specific beamforming vector as $\mathbf{q}_{\text{LOS}}$ and from now, we focus on optimizing the RIS phase shifts.

RIS Design:

For the fixed BS beamformer given in (27) the phase configuration problem is given by:

$$\text{P3: } \min_{\omega_k, \forall k} \sum_{k=1}^K \max_n [\mathbf{t}_k]_n \quad (28a)$$

$$\text{s.t. } \text{SNR}_k^{\min} \geq \gamma_k^{\text{thr}}, \forall k \quad (28b)$$

$$0 \leq [\omega_k]_n < \omega_{\max}, \forall n, k. \quad (28c)$$

First, we redefine the SNR_k in (22) w.r.t. RIS phase shifts:

$$\text{SNR}_k = \underbrace{m_k + \mathbf{v}_k^H \mathbf{m}_k + \mathbf{m}_k^H \mathbf{v}_k}_{\approx 0 \text{ because the BS-MU link is blocked}} + \mathbf{v}_k^H \mathbf{M}_k \mathbf{v}_k, \quad (29)$$

where $\mathbf{v}_k = [\mathbf{e}^{j[\omega_k]_1}, \dots, \mathbf{e}^{j[\omega_k]_N}]^T$ is phase shift vector in RIS for user k , $\mathbf{m}_k^T = \frac{(\mathbf{h}_{d,k}^H \mathbf{q}_k) \mathbf{q}_k^H \mathbf{H}_t^H \text{diag}(\mathbf{h}_{r,k}^H)}{\sigma_n^2}$, and $\mathbf{M}_k = \frac{\text{diag}(\mathbf{h}_{r,k}) \mathbf{H}_t \mathbf{q}_k \mathbf{q}_k^H \mathbf{H}_t^H \text{diag}(\mathbf{h}_{r,k}^H)}{\sigma_n^2}$. For each point within the area allocated to user k , the SNR expressed as $\text{SNR}_k(\mathbf{u}_k) = \mathbf{v}_k^H \mathbf{M}_k(\mathbf{u}_k) \mathbf{v}_k$, $\forall \mathbf{u}_k \in \mathcal{U}_k$, where each element corresponds to a specific $\mathbf{u}_k \in \mathcal{U}_k$. Next, we introduce an auxiliary variable $t_k^{\max}$ for each user, which satisfies the constraint $t_k^{\max} \geq [\mathbf{t}_k]_n$, $\forall n \in \{1, \dots, N\}$. Using this formulation, the problem P3 can be reformulated as:

$$\text{P4: } \min_{t_k^{\max}, \omega_k, \forall k} \sum_{k=1}^K t_k^{\max} \quad (30a)$$

$$\text{s.t. } \mathbf{v}_k^H \mathbf{M}_k(\mathbf{u}_k) \mathbf{v}_k \geq \gamma_k^{\text{thr}}, \forall \mathbf{u}_k \in \mathcal{U}_k, \forall k, \quad (30b)$$

$$[\mathbf{t}_k]_n \leq t_k^{\max}, \forall n, k, \quad (30c)$$

$$0 \leq [\omega_k]_n < \omega_{\max}, \forall n, k. \quad (30d)$$

In P4, constraints (30c) and (30d) are convex and linear w.r.t. variable $[\omega_k]_n$, $\forall k$, respectively. However, (30b) is non-convex w.r.t. variable $[\omega_k]_n$. In the following, we develop a sub-optimal solution to problem P4 based on the Lagrangian formulation and exploiting the fact that N is large.

Let us define $\mathbf{s}_k \triangleq [\mathbf{v}_k^H \mathbf{M}_k(\mathbf{u}_k^{(1)}) \mathbf{v}_k, \dots, \mathbf{v}_k^H \mathbf{M}_k(\mathbf{u}_k^{(|\mathcal{U}_k|)}) \mathbf{v}_k]^T$ where $\mathbf{u}_k^{(j)}$ denotes j th location point of the possible locations for user k in $\mathcal{U}_k$. We define the Lagrange function $L(\mathbf{W}, \mathbf{\Lambda}, \mathbf{\Xi})$ as [43]

$$\sum_{k=1}^K \left(t_k^{\max} + \lambda_k^T (\gamma_k^{\text{thr}} \mathbf{1}_{|\mathcal{U}_k|} - \mathbf{s}_k) + \xi_k^T (\mathbf{t}_k - t_k^{\max} \mathbf{1}_N) \right), \quad (31)$$

where $\lambda_k \in \mathbb{R}_+^{|\mathcal{U}_k|}$ and $\xi_k \in \mathbb{R}_+^N$, $\forall k$ denote the associated Lagrangian multipliers. Let us assume the cardinality of the allocated area for all users is equal, i.e., $|\mathcal{U}_1| = \dots = |\mathcal{U}_K| = |\mathcal{U}|$. Moreover, $\mathbf{\Lambda} = [\lambda_1, \dots, \lambda_K] \in \mathbb{R}_+^{|\mathcal{U}| \times K}$ and $\mathbf{\Xi} = [\xi_1, \dots, \xi_K] \in \mathbb{R}_+^{N \times K}$ are matrices that collect the Lagrangian multipliers. $\mathbf{W} = [\omega_1, \dots, \omega_K] \in \mathcal{W}^{N \times K}$ where $\mathcal{W} = \{\omega \in \mathbb{R} | 0 \leq \omega \leq \omega_{\max}\}$. Based on these definitions, the dual Lagrangian problem can be written

$$\text{P5: } \max_{\mathbf{\Lambda}, \mathbf{\Xi}} \inf_{\mathbf{W}, t_k^{\max}, \forall k} L(\mathbf{W}, \mathbf{\Lambda}, \mathbf{\Xi}) \quad (32a)$$

$$\text{s.t. } \lambda_k \geq 0, \xi_k \geq 0, \forall k, \quad (32b)$$

$$\mathbf{W} \in \mathcal{W}^{N \times K}. \quad (32c)$$

This problem is still non-convex due to the non-convexity of each element in $\mathbf{s}_k$, $\forall k$ vector in the cost function. As can be seen from (31), the phase-shift configuration of all users is coupled together in $\mathbf{t}_k$, $\forall k$. To be more exact, the phase shift set of each user affects the required configuring time for the next and previous phase-shift configuration. To cope with this issue, we continue the analysis of this problem based on a few assumptions, as will be detailed below. We will show in Section IV that despite these simplifying assumptions, the proposed phase-shift design is able to significantly reduce the LC-RIS reconfiguration time.

1) *User Decoupling*: Our first assumption is about the decoupling user phase shifts. With this assumption, we solve P5 iteratively where at each iteration, we focus on one of the ω_k , $k = \{1, \dots, K\}$ where all other $\omega_{k'}$, $k' \neq k$ are fixed. Let us define for $k = 1, \dots, K$:

$$l\Delta\omega_k \triangleq \begin{cases} \omega_k - \omega_{k-1}, & \text{if } k \neq 1 \\ \omega_1 - \omega_K, & \text{if } k = 1. \end{cases} \quad (33)$$

Then, we can transform the problem P5 into a decoupled problem on each user k . Let us define L_k for k th user as

$$L_k \triangleq t_k^{\max} + t_{k+1}^{\max} + \lambda_k^T (\gamma_k^{\text{thr}} \mathbf{1}_{|\mathcal{U}|} - \mathbf{s}_k) + \xi_k^T (\mathbf{t}_k - t_k^{\max} \mathbf{1}_N) + \xi_{k+1}^T (\mathbf{t}_{k+1} - t_{k+1}^{\max} \mathbf{1}_N), \quad (34)$$

where $\mathbf{t}_k$ and $\mathbf{t}_{k+1}$ are functions of ω_k , ω_{k-1} and ω_{k+1} . Here, we assumed ω_{k-1} and ω_{k+1} are fixed and without change at each iteration. From now on, we optimize $L_k(\omega_k, \lambda_k, \xi_k, \xi_{k+1})$ sequentially for users $k = 1, \dots, K$.

Due to the non-convexity of the $\mathbf{s}_k$, we still cannot find a tractable solution for this problem. Hence, we apply the second assumption in the following.

2) *Decoupling on Each RIS Element*: The phase shifts of an LC-RIS are inherently coupled within the vector $\mathbf{s}_k$, as represented in (34). This coupling significantly increases computational complexity, particularly as the number of RIS elements grows. To mitigate this issue, we employ the parallel coordinate descent (PCD) method [44]. A key challenge in using this approach here is that $\mathbf{s}_k$ represents a vector over multiple spatial points, making direct optimization intractable. To address this, we adopt a strategy inspired by [41], where each RIS element is associated with a single representative point in the targeted area. The method consists of the following steps:

Step 1: Mapping RIS elements to pre-defined points: We adopt a strategy similar to those proposed in [31] and [41]. In particular, we assign each RIS element to a single point within the set $\mathcal{U}_k$. This mapping simplifies the problem by ensuring a one-to-one correspondence between RIS elements and spatial points. With this mapping, we approximate the L_k as

$$L_k \approx t_k^{\max} + t_{k+1}^{\max} + \sum_{n=1}^N \left([\lambda_k]_n (\gamma_k^{\text{thr}} - \text{SNR}(\mathbf{u}_k^{(n)})) \right) + \xi_k^T (\mathbf{t}_k - t_k^{\max} \mathbf{1}_N) + \xi_{k+1}^T (\mathbf{t}_{k+1} - t_{k+1}^{\max} \mathbf{1}_N), \quad (35)$$

where $\mathbf{u}_k^{(n)}$ represents the point in $\mathcal{U}_k$ associated with the n -th RIS element. Unlike (34), where optimization is performed over the entire set for each element, each LC-RIS element is optimized for a specific assigned point in (35). This phase shift mapping is effective when the number of RIS elements (N) is sufficiently large relative to the targeted coverage area. According to the $\mathbf{u}_k^{(n)}$ point, there is a best phase shift for n th element to maximize the SNR in the targeted point. We denote that optimized phase shift for the n -th element and k th user as $\phi_k(\mathbf{u}_k^{(n)})$.

Step 2: PCD optimization method: After assigning each RIS element to a specific point, we optimize the phase shifts using the PCD method. Given that the original function $\text{SNR}(\mathbf{u}_k^{(n)})$ depends on all elements of $[\omega_k]_1, \dots, [\omega_k]_N$, we define a coordinate-wise function for each element in the following Lemma to reduce more the complexity of the optimization.

Lemma 2: For sufficiently large N , blocked BS-MU link, dominant LOS link between both BS-RIS and RIS-MU, and fixing all variables $[\omega_k]_n, \forall n$ except one of them ($[\omega_k]_p$), the L_k can be written as

$$L_k = C_k + [L_k]_p, \quad (36)$$

where C_k is a constant for k th user and

$$[L_k]_p = [\xi_{k+1}]_p [\mathbf{t}_{k+1}]_p + [\xi_k]_p [\mathbf{t}_k]_p - [\lambda_k]_p \cos([\omega_k]_p - \phi_k(\mathbf{u}_k^{(p)})). \quad (37)$$

Proof: The proof is provided in Appendix C. $\square$

Using Lemma 2, we can derive the optimal phase shifts for each element independently, reducing the problem's dimensionality. To further reduce computational complexity, we

optimize all elements in parallel using the result of Lemma 2. Now, we have $[L_k]_n$ in (37), $\forall n$ where

$$[\mathbf{t}_k]_n = \begin{cases} \tau_c^- \ln \left(\frac{[\omega_{k-1}]_n - \omega_{\min}}{[\omega_k]_n - \omega_{\min}} \right), & \text{if } [\omega_{k-1}]_n > [\omega_k]_n, \\ \tau_c^+ \ln \left(\frac{\omega_{\max} - [\omega_{k-1}]_n}{\omega_{\max} - [\omega_k]_n} \right), & \text{if } [\omega_{k-1}]_n < [\omega_k]_n, \end{cases} \quad (38)$$

and

$$[\mathbf{t}_{k+1}]_n = \begin{cases} \tau_c^- \ln \left(\frac{[\omega_k]_n - \omega_{\min}}{[\omega_{k+1}]_n - \omega_{\min}} \right), & \text{if } [\omega_k]_n > [\omega_{k+1}]_n, \\ \tau_c^+ \ln \left(\frac{\omega_{\max} - [\omega_k]_n}{\omega_{\max} - [\omega_{k+1}]_n} \right), & \text{if } [\omega_k]_n < [\omega_{k+1}]_n. \end{cases} \quad (39)$$

We can define $[\mathbf{t}_k]_n$ in a more concise way as follows:

$$[\mathbf{t}_k]_n = \frac{\text{sign}([\omega_k]_n - [\omega_{k-1}]_n) + 1}{2} \tau_c^+ \ln \left(\frac{\omega_{\max} - [\omega_{k-1}]_n}{\omega_{\max} - [\omega_k]_n} \right) - \frac{\text{sign}([\omega_k]_n - [\omega_{k-1}]_n) - 1}{2} \tau_c^- \ln \left(\frac{[\omega_{k-1}]_n - \omega_{\min}}{[\omega_k]_n - \omega_{\min}} \right). \quad (40)$$

The algorithm steps and complexity analysis are described in the following part.

Algorithm 1 Proposed Algorithm for Problem (P4)

- 1: **Initialize:** Derive $[\phi_k(\mathbf{u}_k^{(n)})]_n, \forall n, k$ from [45, Eqs. (5),(6)]. Set $\mathbf{W}^{(0)} = [\phi_1, \dots, \phi_K], \mathbf{q}_k = \mathbf{q}_{\text{LOS}}, \lambda_k^{(0)} > 0$ and $t_k^{\max}, \forall k = 1, \dots, K$. Set $0 < \alpha < 1, I_{\max}, \Delta t_k$.
 - 2: **for** $i = 1 : I_{\max}$ **do**
 - 3: Set $[\xi_k]_n = 1$ if $[\mathbf{t}_k]_n \geq t_k^{\max}$, and 0 otherwise, $\forall k, n$.
 - 4: **for** $k = 1 : K$ **do**
 - 5: Find $[\omega_k]_n^{(i)} = [\omega_k^g]_n \in (0, \omega_{\max})$ as a minimizer of $L_n, \forall n$ in (37) with line search.
 - 6: Calculate $\text{SNR}_k^{\min(i)} = \min_{\mathbf{u}_k \in \mathcal{U}_k} \text{SNR}_k^{(i)}(\mathbf{u}_k)$.
 - 7: **if** $\text{SNR}_k^{\min(i)} < \gamma_k^{\text{thr}}$ **then**
 - 8: Update $\omega_k^{(i)} = \omega_k^{(i-1)}$, and $\lambda_k^{(i)} = \frac{\lambda_k^{(i-1)}}{\alpha}$.
 - 9: **else**
 - 10: Update $\omega_k^{(i)} = \omega_k^g$, and $\lambda_k^{(i)} = \alpha \lambda_k^{(i-1)}$.
 - 11: **end if**
 - 12: **end for**
 - 13: Update $t_k^{\max} = t_k^{\max} - \Delta t_k$.
 - 14: **end for**
-

Algorithm and complexity analysis: The proposed algorithm to obtain a suboptimal solution to P1 is summarized in Algorithm 1. We start with an initial number for $t_k^{\max}$ and $\lambda_k, \forall k$. At each round ($i = 1 : I_{\max}$), we aim in decreasing $t_k^{\max}$, where $I_{\max}$ is the maximum number of iterations and is determined in practice according to the available budget for computation time and complexity. In each iteration ($k = 1 : K$), the focus is on one user, adjusting its phase shifters within the range of $(0, \omega_{\max})$ to minimize the transition time from ω_{k-1} to ω_k and also from ω_k to ω_{k+1} . With the help of Lemma 2, we can optimize the phase shift of each element in parallel. To do so, first, if $[\mathbf{t}_v]_n < t_v^{\max}$ where $v = \{k, k+1\}$, we set the value of $[\xi_g]_n = 0$. Otherwise, we set $[\xi_g]_n = 1$ leading to a higher penalty for that LC-RIS phase

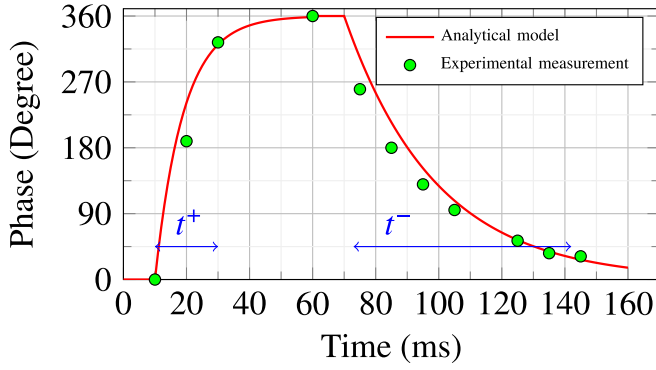


Fig. 4. Experimental result of an LC phase shifter response time (green points) [26, Fig. 4a], and analytical model (red line). The time from 10% (90%) to reach 90% (10%) of the desired phase is 15 ms (72 ms) for positive (negative) phase shifts, corresponding to exponent $\tau_c^+ = 9$ ms ($\tau_c^- = 29$ ms).

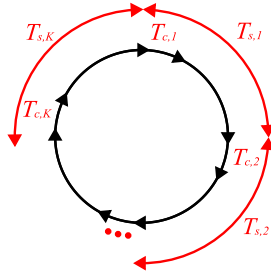


Fig. 5. Configuring time (T_c) and switching time (T_s) are shown for each user when they are served by an RIS sequentially.

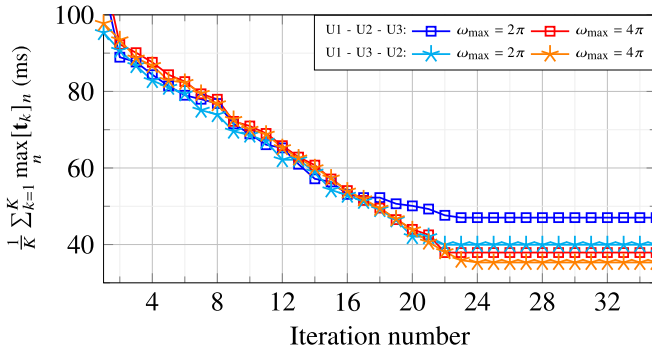


Fig. 6. $\frac{1}{K} \sum_{k=1}^K \max_n[t_k]_n$ vs. number of the iteration.

shifter element. Second, to find the minimizer of each $[L_k]_n$ to minimize (37) per element, we perform a one-dimensional line search. Specifically, we employ a grid search over the feasible phase-shift range $(0, \omega_{\max})$. This range is discretized into L uniformly spaced points, and the cost function is evaluated at each point. The phase value that yields the minimum cost is then chosen as the updated solution. We name this minimizer as $[\omega_k^g]_n$. Once a new phase shift is determined for the k -th user, the SNR constraint is verified. If the constraint remains satisfied, λ_k decreases to explore additional potential solutions, otherwise, if the constraint is violated, λ_k is increased to prioritize meeting the SNR constraint.

The dominant computational complexity of the algorithm arises from determining $[\omega_k^g]_n$ in each iteration. Since our

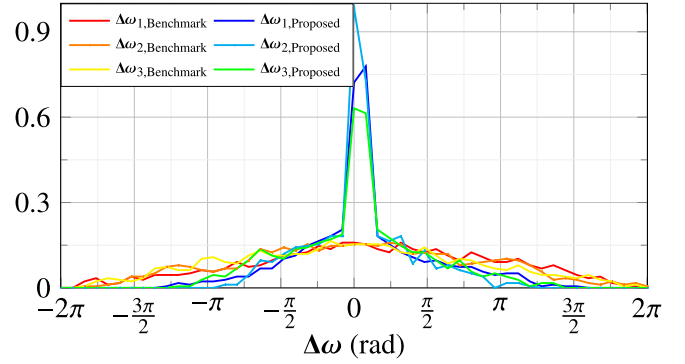


Fig. 7. Histogram of $\Delta\omega$ in the proposed and benchmark designs for each configuration.

grid search requires evaluating the cost function at L points for each of the N RIS elements, the complexity of one full update for user k is $\mathcal{O}(NL)$. Therefore, the overall dominant complexity of the algorithm is $\mathcal{O}(I_{\max} NKL)$, which is linear with the number of RIS elements N and the search resolution L .

IV. PERFORMANCE EVALUATION

A. Simulation Setup

We employ the simulation configuration for coverage extension illustrated in Fig. 1. We assume there are three users located in different and random locations with uniform distribution inside the areas $[10 \pm 1, 2 \pm 1, -5]$ m, $[10 \pm 1, -5 \pm 1, -5]$ m, and $[10 \pm 1, 5 \pm 1, -5]$ m, respectively. The BS comprises a $4 \times 4 = 16$ uniform planar array (UPA) in $x-z$ axes positioned at $[40, 20, 5]$ m. The RIS comprises a UPA located at $[0, 0, 0]$ m, consisting of $N_y \times N_z = 150 \times 5$ elements along the y - and z -axes, respectively (except in Fig. 8, where the number of RIS elements is specified in each subfigure). The element spacing for UPAs at both the BS and RIS corresponds to half of the wavelength. The MUs have a single antenna. The noise variance is computed as $\sigma_n^2 = WN_0N_f$ with $N_0 = -174$ dBm/Hz, $W = 20$ MHz, and $N_f = 6$ dB. We assume 28 GHz carrier frequency, the path-loss $\beta = -61$ dB at $d_0 = 1$ m, and $\gamma_{\text{thr}} = 10$ dB. Given these parameters, the users' locations fall into NF regime of the LC-RIS (i.e., the distances between the RIS and the center of the coverage areas are 11.4 m, 12.2 m, and 12.2 m for users 1-3, respectively, which are below the Fraunhofer distance $d_{\text{FF}} = \frac{2D^2}{\lambda} = 120$ m, where D is the largest physical dimension of the RIS [29], [31]). Moreover, we adopt path-loss exponent $\eta = (2, 2, 2)$ and $K_f = (-100, 10, 10)$ dB for the BS-UE, BS-RIS, and RIS-MU channels, respectively, and additional blockage $h_{\text{blk}} = -40$ dB for the direct link BS-UE. The main focus of this analysis is the SNR comparison. As a benchmark, we consider the analytical RIS design based on anomalous reflection in [5] which is unaware of the transition behavior of the LC-RIS. This applies to all figures except Fig. 8, where we also include semidefinite relaxation (SDR) [2], [46] as an additional benchmark. While SDR achieves a better steady-state SNR than analytical methods [5], its complexity scales with $\mathcal{O}(N^{3.5})$, making it unscalable for

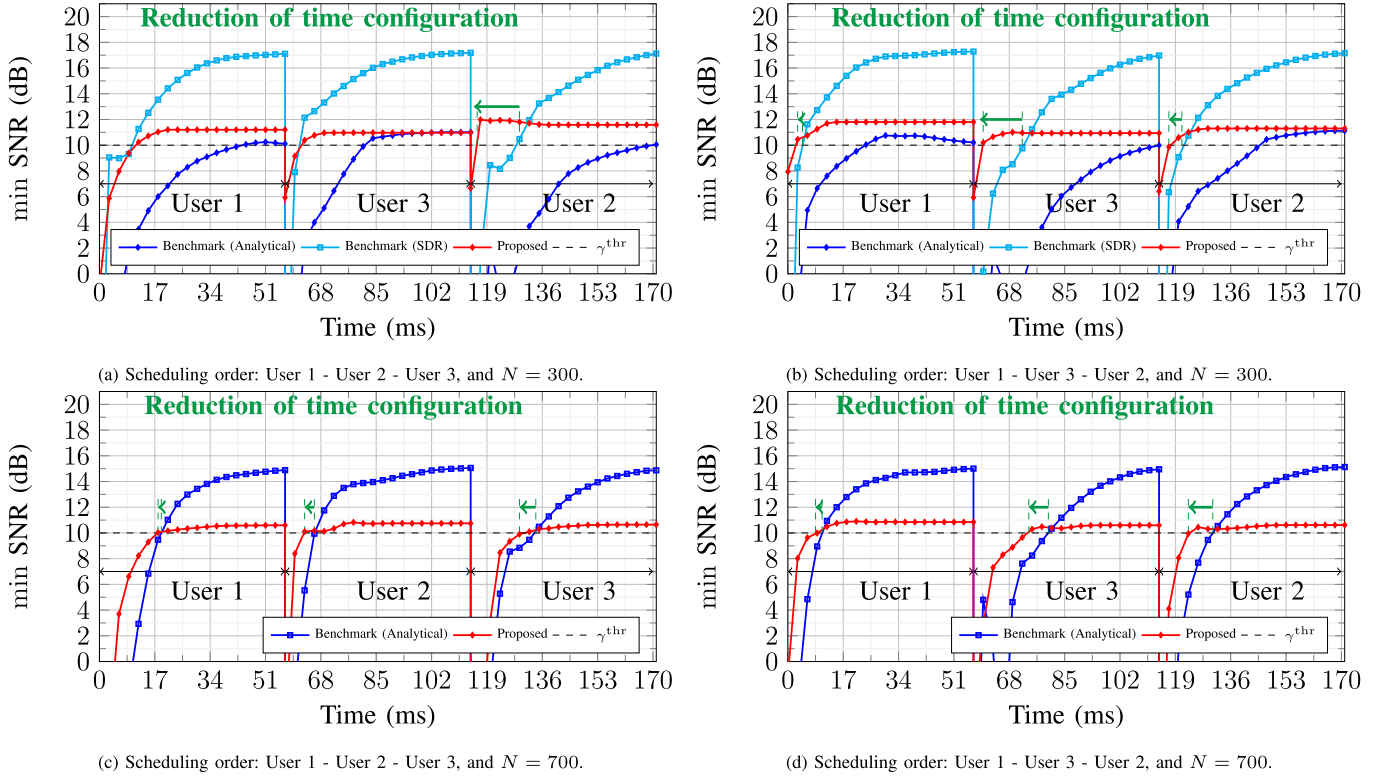


Fig. 8. SNR comparison between benchmarks and proposed algorithm with 57 ms serving the users in two different cases of users' sequence.

large arrays. Therefore, SDR is used only in Fig. 8, and for the remainder of the figures, 'Benchmark' refers to the analytical method [5]. We adopted $\tau_c^+ = 9$ ms and $\tau_c^- = 29$ ms based on the data provided in [26]. The parameters used in this simulation are $\alpha = 0.95$, $I_{\max} = 35$, $P_t = 38$ dBm, $t_k^{\max} = 100$ ms, and $\Delta t_k = \frac{t_k^{\max}}{f_{\max}}$ ms $\forall k$.

Remark 1: The MATLAB codes used to generate the simulation results in this section are publicly available online at <https://github.com/MohamadRezaDelbari/LC-RIS-TDMA-Journal>

B. Convergence Behavior

We first discuss the convergence properties of Algorithm 1. This algorithm is designed such that the auxiliary variables $t_k^{\max}$, $\forall k$ are decreasing in each outer loop. While this does not strictly guarantee a monotonic decrease of the objective function, $\max_n [\mathbf{t}_k]_n$, $\forall k$, in every inner step, our simulation demonstrates that the algorithm consistently converges to a stable solution, typically within 20-30 iterations for the parameters adopted in this work.

This convergence is illustrated in Fig. 6, where we show the averaged cost function of P1 over K versus the number of iterations. As shown, these transition-time values decrease progressively with each iteration. This behavior occurs because the initial configuration ensures that $\text{SNR}_k(\mathbf{u}_k)$, $\forall \mathbf{u}_k \in \mathcal{U}_k$ is greater than the threshold. Furthermore, each iteration is designed to guarantee that $\text{SNR}_k(\mathbf{u}_k)$, $\forall \mathbf{u}_k \in \mathcal{U}_k$, in subsequent iterations, does not drop below this threshold. Consequently, the final $\min_{\mathbf{u}_k \in \mathcal{U}_k} \text{SNR}_k(\mathbf{u}_k)$ value approaches the

threshold, owing to the reduction in $\max [\mathbf{t}_k]_n$. Additionally, the maximum differential phases of LC-RISs can be increased beyond 2π by increasing the phase shifter length (ℓ_{lc}), see (7). Typically, due to periodicity of complex exponential function, $\omega_{\max} > 2\pi$ does not improve the SNR; however, as can be seen from Fig. 6, $\omega_{\max} > 2\pi$ can significantly improve the LC-RIS transition time due to the influence of $\omega_{\max}$ as can be observed in (20) and (21). When the length of the LC is increased to generate a 4π differential phase shift, the transition time decreases further, highlighting the efficiency improvement.

C. Histogram of Differential Phase Shift Values

In Fig. 7, we plot the histogram of the differential phase shifts before and after optimization for $K = 3$ users. In particular, the horizontal axis represents the value of the differential phase shifts in radians, ranging from -2π to 2π . This reflects how much the phase values differ between successive configurations. The vertical axis represents the frequency (or normalized probability density) of these differential values, i.e., how often each range of differential phase values occurs. For the benchmark method, the phase-shift distribution is approximately uniform. Hence, the differential phase shifts span the full range and resemble a triangular-like distribution. This is expected from subtracting two uniform distributions (the convolution of two uniform distributions is a triangular-like distribution). This uniformity is observed because for a large NF RIS serving a designated area, the optimal phase required to satisfy the SNR constraint varies significantly

across the surface, causing the phase values to be spread across the entire $(0, 2\pi)$ range.

However, for the proposed method, the distribution becomes sharply peaked around zero, indicating that most differential shifts are small. This change reflects the impact of optimizing the RIS phase profiles to reduce configuration switching time, as smaller phase differences between successive configurations result in faster reconfiguration. Hence, this figure visually confirms the connection between smaller differential phase shifts and faster configuration times.

D. Transition Time Reduction and Impact of Scheduling Order

Fig. 8 shows the SNR (dB) when the RIS configuration is switched every 57 ms to serve the users. The selected time interval is intended to showcase the proposed algorithm's real-time performance. DAs is consistently illustrated in the Figs. 8a, 8b, 8c, and 8d, our algorithm reaches the required SNR threshold (i.e., 10 dB) significantly faster than the benchmarks (analytical and SDR), which simply maximizes the minimum received SNR in the targeted area without accounting for the transition time. In particular, our algorithm prioritizes minimizing transition time and is therefore designed to find a phase-shift configuration that satisfies the minimum SNR constraint in the targeted area as quickly as possible. Specifically, the algorithm does this by identifying a phase-shift configuration that satisfies the SNR requirement while making adjustments to the existing RIS phase-shifts. DThe benchmarks, conversely, only searches for the configuration that maximizes the final SNR, regardless of the time taken. For time-critical TDMA applications, reaching the operational threshold quickly is an important performance metric that confirms the practical advantage of our approach. DSDR outperforms the analytical method by achieving a higher steady-state SNR, but at the cost of higher complexity. Consequently, it is feasible to implement it with $N = 300$. Moreover, the basis for the design of the low-complexity analytical design in [41] is that the RIS is extremely large. This is why the analytical benchmark performs poorly for small N but well for large N .

In addition, we analyze the impact of ordering users in accelerating the required time for phase configuration in RIS. In general, when K users must be served in TDMA application, there are $(K - 1)!$ different possibilities for ordering due to circularity. In the simplest case, let us assume $K = 3$ and there are two different way to support users. DAs can be seen in Figs. 8a, 8b, 8c, and 8d, the minimum SNR in the targeted area is plotted against the time in millisecond (ms). In all figures, there are improvements in time but in comparison of two set of configurations, the second configuration (User 1 - User 3 - User 2) can decrease more the required time compared to first configuration (User 1 - User 2 - User 3) showing the importance of ordering in LC-RISs.

E. Effective Data Rate

In Fig. 9, we plot the effective data rate given by

$$R = \frac{\max(T_s - T_c, 0)}{T_s} \log_2(1 + \text{SNR}_{\text{thr}}), \quad (41)$$

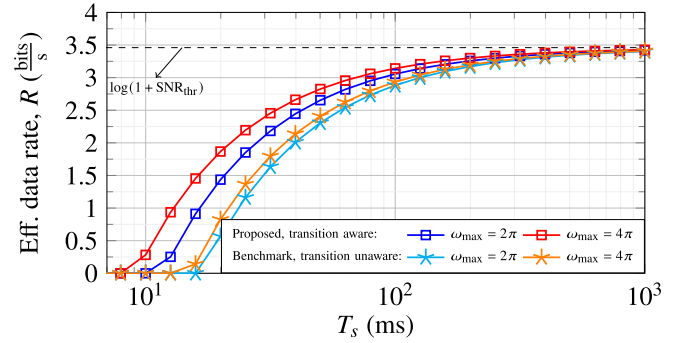


Fig. 9. The average effective data rate (R) versus switching time between three users, T_s , for two different algorithms.

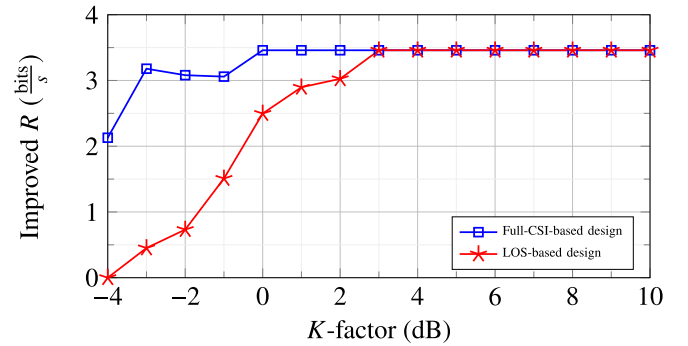


Fig. 10. Improvement in effective data rate (R) of the proposed transition-aware design (Algorithm 1) relative to a transition-unaware benchmark versus the Rician K -factor, for LoS-only knowledge and full CSI (LoS and NLoS).

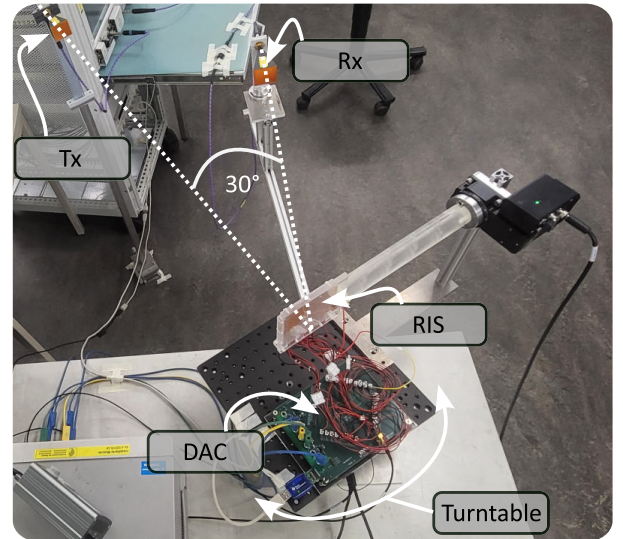
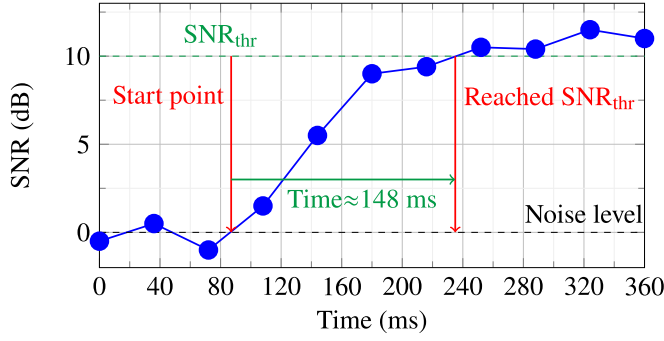
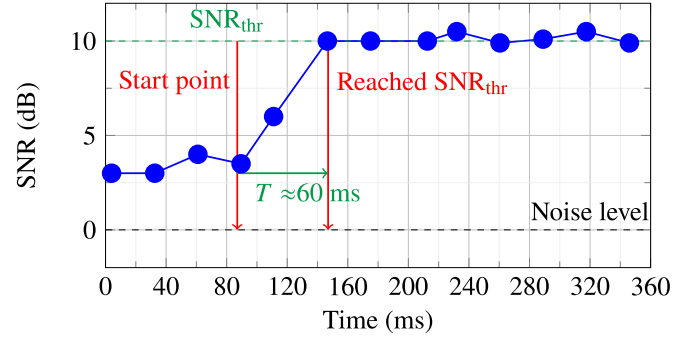


Fig. 11. Experimental setup for measuring the SNR in different positions. For more details of the setup, you can see [26, Fig. 5].

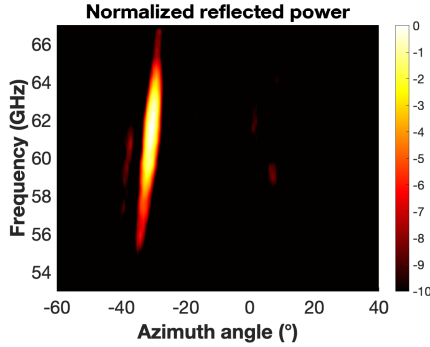
where T_c is the time RIS needs to reconfigure and reach SNR_{thr} and T_s is the time interval that RIS switches between serving the users (determined by the application scenario, e.g., delay requirement, users' mobility). Since there are multiple users, we plot the results based on the average T_c across all users. When T_s is smaller than this average value, the effective



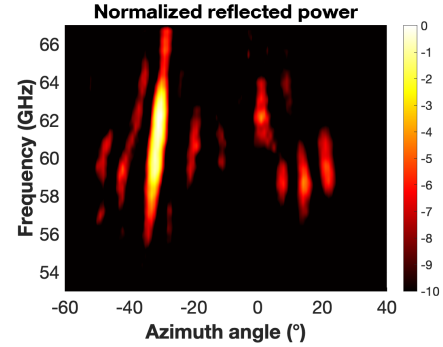
(a) Measurement results for the benchmark phase shift optimization, which requires approximately 148 ms to achieve the SNR_{thr} . This algorithm only focuses on maximizing the final receive power.



(b) Measurement results for the proposed phase shift optimization, which requires approximately 60 ms to achieve the SNR_{thr} . This algorithm minimizes the transition time while reaching the SNR_{thr} .



(c) Measured received signal power after sufficient time. The benchmark algorithm maximizes the final received power regardless of the transition time.



(d) Measured received signal power after sufficient time. The proposed algorithm accelerates the transition time at the cost of a slight reduction in the final received power; however, it remains well above the required SNR_{thr} .

Fig. 12. A comparison between the two algorithms is conducted when applying new phase shifts on the LC-RIS to serve the second user. In this analysis, SNR_{thr} is set to be 10 dB higher than the noise power.

data rate R is zero for both algorithms due to insufficient time for reconfiguration, as implied by (41). As T_s increases, the proposed algorithm achieves a higher data rate for the users. In the limit as $T_s \rightarrow \infty$, both algorithms converge to $R \rightarrow \log_2(1 + \text{SNR}_{\text{thr}})$. Furthermore, Fig. 9 shows that increasing the maximum phase shift range from 2π to 4π significantly improves the effective data rate. This demonstrates a key insight of our work: even though a 2π phase range is sufficient for arbitrary beamforming at the steady state, a larger physical range (ω_{max}) directly accelerates the reconfiguration time, as modeled in (20) and (21). A larger ω_{max} provides a greater dynamic range for the phase transition, increasing the rate of phase change and thus reducing the time T_c required to reach any desired intermediate phase. This highlights a crucial design trade-off where increasing the LC-RIS phase shifter length can be leveraged to overcome the inherent slow response time of the material. Additionally, increasing the maximum phase shift range from 2π to 4π further improves the effective data rate in both algorithms.

F. Validation of the LoS-Dominant Link Assumption

Figure 10 shows the improvement in effective data rate (R) of the proposed transition-aware design (Algorithm 1) over the transition-unaware benchmark as a function of the Rician K -factor. The red curve reports the gain when the RIS knows only the LOS component (derived from user location), whereas the blue curve represents an upper bound

obtained with full CSI including both LOS and non-LOS components. The gap to the upper bound decreases as the K -factor increases. For the parameters considered, the data rate achieved by the LOS-based scheme approaches the full-CSI performance for $K \gtrsim 3$ dB. Such K -factor values are typical in mmWave environments [25], which supports the LOS-based design adopted in this work.

Figure 10 also confirms that Algorithm 1 is not limited to using only LOS link information; it can also improve the effective data rate when full CSI knowledge is available. Specifically, our algorithm operates by taking an initial target phase-shift configuration, denoted as $\mathbf{W}^{(0)} = [\phi_1, \dots, \phi_K]$, which is assumed to be pre-computed based on the available CSI (whether LOS-based or full). Our algorithm then takes this $\mathbf{W}^{(0)}$ as input and computes a new set of phase shifts $\mathbf{W} = [\omega_1, \dots, \omega_K]$ that satisfies the QoS constraint while minimizing the transition time from the previous state.

V. EXPERIMENTAL RESULTS

In this section, we evaluate the proposed algorithm using a small-scale proof of concept implementation of an LC-RIS. The experimental LC-RIS comprises $30 \times 25 = 750$ elements, with the same voltage applied across all elements within a column for 1D beamforming.⁶ This configuration allows the LC-RIS to reflect waves toward different azimuth angles but

⁶The reason is the experimental hardware limitations, otherwise, our proposed algorithm can also be used in 2D.

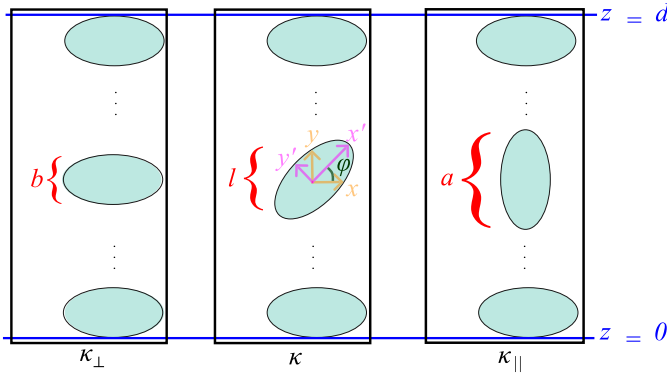


Fig. 13. A rotated ellipse shape LC molecule with semi-major and semi-minor diameters a and b , respectively.

not varying elevation angles. These bias voltages are applied with a 1 kHz square wave. We used DAC60096 EVM with 12 bits from Texas Instruments providing us ± 10.5 V. As shown in Fig. 11, we used two 25dBi V-band horn antennas from MI-wave where the Tx is fixed at an azimuth angle of 30° and the distance of 1 m relative to the LC-RIS, while the Rx can rotate for any azimuth angle with a fixed distance of 55 cm from the LC-RIS. For this example, the receiver is set to an angle of -30° . The measurements have been performed with a PNA-X N5247A from Keysight Technologies.

We evaluate two types of phase-shifter optimization strategies: benchmark and the proposed method. As our work focuses in addressing the dynamic reconfiguration latency of an LC-RIS, we consider time reduction in satisfying the QoS as the performance metric same as Section IV-D. We evaluate our method against the most powerful alternative: a benchmark algorithm that is transition-unaware but is otherwise optimal for the static case, meaning it only tries to maximize the final SNR without accounting for the transition time [5], [24], [45]. Consider a scenario with two receivers located at $\theta_1 = -30^\circ$ and $\theta_2 = 30^\circ$, and a transmitter located at $\theta = 0^\circ$. During a transition, the LC-RIS must shift its focus from serving the first receiver to the second. This requires a seamless adjustment of phase shifters to ensure fast transition. With linear phase shift design, as can be seen from Fig. 12a, it takes about 148 ms to received power reaches the required SNR_{thr} which is 10 dB higher than noise power (note that the fabrication, i.e., the rubbing, of the alignment layer is not ideal which results in a longer response time compared to the reported values in Fig. 4). However, with the proposed phase shifter design, Fig. 12b, the required time reduces to only 60 ms at the cost of the reducing in final received SNR. These results suggest the effectiveness of the proposed modeling and algorithm design in practice. It is important to note that the starting SNR for the second user is higher in the proposed method (Fig. 12a) compared to the benchmark (Fig. 12b). This is not an artifact of the experiment but a direct result of the transition-aware algorithm. The benchmark, being unaware of the subsequent user, generates a configuration for the first user that is solely optimal for that user's link. In contrast, our algorithm finds a configuration for the first user that not only satisfies its SNR requirement but is also 'closer'

to the optimal configuration for the second user, thereby preconditioning the RIS for a faster transition. Consequently, the higher initial SNR is an integral part of the overall performance gain demonstrated by our approach. The final received SNR for benchmark and the proposed algorithms are plotted in Figs. 12c and 12d, respectively. Although the setup is designed for $f = 60$ GHz, the received power spans a frequency range of 53 – 67 GHz. Outside $f = 60$ GHz, beam splitting effects are observed, which may be considered as a potential direction for future research.

VI. CONCLUSION AND FUTURE DIRECTIONS

In this paper, we have studied LC-RISs as a scalable cost-effective solution to realize large RISs required at mmWave band to generate sufficient link budget. The primary challenge of LC-RISs is however their slow response. To address this challenge, we have first developed a physics-driven model to characterize the response time of LC-RIS. Based on this model, we proposed an efficient algorithm for fast reconfiguration of LC-RISs to serve K users in a TDMA framework. Simulation results validated the effectiveness of the proposed transition-aware design, showcasing its significant performance advantages over existing transition-unaware benchmarks in the literature. DSpecif-ically, our analysis and results revealed that the proposed algorithm reduces the transition time by selecting phase configurations that minimize the differential phase shift required between consecutively served MUs. This approach effectively preconditions the RIS state, leading to a higher starting SNR for the subsequent user and a smaller transition range, thereby satisfying the QoS requirement significantly faster than conventional transition-unaware designs. Furthermore, the experimental validation confirmed the effectiveness of the algorithm in a proof of concept implementation of LC-RIS in an indoor application.

This paper lays the necessary foundation for several extensions to fully investigate the potential of LC-RISs. Although our study focused on the TDMA protocol, extending the transition-aware design to multi-user scenarios with hybrid beamforming at the BS requires a joint optimization of interference mitigation and reconfiguration latency, which constitutes a promising research direction [27]. Furthermore, while the proposed algorithm is inherently robust to location estimation errors, it has been primarily developed for LOS-dominated environments. In scenarios where non-LOS links cannot be neglected, robust designs that account for imperfect full CSI are required. Finally, while our simulations highlighted the impact of scheduling order and LC phase-shifter length on performance, a more comprehensive analysis of these factors is left for future work.

APPENDIX A PROOF OF LEMMA 1

Lemma 1 deals with solving a modified heat PDE that incorporates a nonlinear external source $\varepsilon_0 \Delta \varepsilon E^2 \Phi(\varphi)$. In the following, we solve this PDE by analyzing two specific cases:

- The electric field decreases from its maximum value (E_{max}) to zero (decaying in time).

- The electric field increases from zero to $E_{\max}$ (rising in time).

A. Case 1: Decaying in Time

When $E = 0$, the nonlinear term $\varepsilon_0 \Delta \varepsilon E^2 \Phi(\varphi)$ in (15) vanishes, reducing the PDE to a standard heat equation:

$$\gamma_1 \frac{\partial \varphi}{\partial t} = \bar{K} \frac{\partial^2 \varphi}{\partial z^2} \quad (42a)$$

$$\text{C1, C2, and C3.} \quad (42b)$$

Using the method of separation of variables [47], the general solution of (42a) subject to C1 and C2 is:

$$\varphi(z, t) = \sum_{n=1}^{\infty} A_n \sin\left(\frac{n\pi z}{d}\right) e^{-\frac{\bar{K}}{\gamma_1} \left(\frac{n\pi}{d}\right)^2 t}, \quad (43)$$

where the coefficients A_n are determined by the initial condition:

$$A_n = \frac{2}{d} \int_0^d \varphi(z, 0) \sin\left(\frac{n\pi z}{d}\right) dz. \quad (44)$$

Due to the initial condition in C3: $\varphi(z, 0) = A_1 \sin\left(\frac{\pi z}{d}\right)$, we find that $A_n = 0$ for all $n > 1$. Hence, the solution simplifies to:

$$\varphi(z, t) = 0 + (A_1 - 0) \sin\left(\frac{\pi z}{d}\right) e^{-\frac{t}{\tau_{\text{mol}}}}, \quad (45)$$

where $\tau_{\text{mol}}^- = \frac{\gamma_1}{\bar{K}} \left(\frac{d}{\pi}\right)^2$.

B. Case 2: Rising in Time

When $E = E_{\max}$, the nonlinear term $\varepsilon_0 \Delta \varepsilon E^2 \Phi(\varphi)$ becomes significant. This PDE has been solved in [21], [22] with the following solution:

$$\varphi(z, t) = \frac{\pi}{2} \sqrt{\frac{1}{1 + \left(\frac{\pi^2}{4A_1^2} - 1\right) \exp\left(-\frac{t}{\tau_r}\right)}} \sin\left(\frac{\pi z}{d}\right), \quad (46)$$

where $\tau_r = \frac{\gamma_1}{2|\varepsilon_0 \Delta \varepsilon E_{\max}^2 - \frac{\pi^2}{d^2} \bar{K}|} \propto E_{\max}^{-2}$ (recall $E_{\max} \gg \frac{\pi^2 \bar{K}}{d^2 \varepsilon_0 \Delta \varepsilon}$). We now prove that the following approximation instead of (46) has an error bounded by $\mathcal{O}(E_{\max}^{-2})$:

$$\varphi(z, t) = \left(\frac{\pi}{2} - \left(\frac{\pi}{2} - A_1\right) e^{-\frac{t}{\tau_{\text{mol}}^+}}\right) \sin\left(\frac{\pi z}{d}\right), \quad (47)$$

where $\tau_{\text{mol}}^+ = \alpha \tau_r$, and $\alpha \geq 1$ is a tuning parameter (in this case, we know $\varphi(z, \infty) = \frac{\pi}{2} \sin\left(\frac{\pi z}{d}\right)$ after applying $E = E_{\max}$ [21], [22]).

Error Analysis: To analyze the error, we compute the integral of the absolute difference between (46) and (47) over time. For an upper bound, we consider $z = \frac{d}{2}$, and $\alpha = 1$:

$$\int_0^{\infty} \frac{\pi}{2} \left(1 - \sqrt{\frac{1}{1 + \left(\frac{\pi^2}{4A_1^2} - 1\right) \exp\left(-\frac{t}{\tau_{\text{mol}}^+}\right)}} + \left(\frac{\pi}{2} - A_1\right) e^{-\frac{t}{\tau_{\text{mol}}^+}} dt \right) \quad (48)$$

$\gamma E_t \triangleq \sqrt{\frac{\bar{K}}{\varepsilon_0 \Delta \varepsilon}} \frac{\pi}{d}$ represents the Freedericksz threshold field necessary to initiate rotation of the LC molecules.

Evaluating this integral yields:

$$\tau_{\text{mol}}^+ \left(\underbrace{\log\left(\frac{\frac{\pi}{2A_1} + 1}{\frac{\pi}{2A_1} - 1}\right) + \log\left(\frac{\frac{\pi^2}{4A_1^2} - 1}{4}\right)}_{\beta} + \frac{\pi}{2} - A_1 \right). \quad (49)$$

Hence, the error is bounded by $\tau_{\text{mol}}^+ \beta$, where $\tau_{\text{mol}}^+ \propto E_{\max}^{-2}$. This confirms that the approximation in (47) holds with an error of $\mathcal{O}(E_{\max}^{-2})$. This concludes the proof.

APPENDIX B

PROOF OF PROPOSITION 1

To characterize the time-dependent behavior of the LC phase shifter, we begin by considering (7). In this equation, ℓ_{lc} and $\kappa_{\perp}$ remain constant over time, while $\kappa(t)$ is time-dependent. In general, $\kappa(t)$ is given by:

$$\kappa(t) = \frac{1}{d} \int_0^d \left(\left(\frac{l(\varphi(z, t))}{a} \right) \kappa_{\text{lc}} + \left(\frac{a - l(\varphi(z, t))}{a} \right) \kappa_0 \right) dz, \quad (50)$$

where a represents the semi-major axis of the ellipsoidal shape LC, $l(\varphi(z, t))$ is the effective length of the LC region contributing to the phase shift, κ_{lc} and κ_0 are the wave numbers in LC material and vacuum, respectively, see Fig. 13. As shown in the left figure of Fig. 13, to compute $\kappa_{\perp}$, we substitute $l(\varphi(z, t)) = b$, $\forall z$ in (50), where b is the semi-minor axis of the LC ellipsoid. Then, we can express the dynamic phase shift function in (7) as:

$$\omega(t) = \ell_{\text{lc}} \frac{1}{d} \int_0^d \left(\left(\frac{l(\varphi(z, t)) - b}{a} \right) (\kappa_{\text{lc}} - \kappa_0) \right) dz. \quad (51)$$

In the following, we first derive the function of $l(\varphi)$. Then, we establish that the phase shift generated by each LC-RIS phase shifter evolves as a summation on exponential functions of time. We begin by proving that the effective length of an LC molecule, depicted in Fig. 13, is given by:

$$l(\varphi) = b \sqrt{\frac{1}{1 + \left(\frac{b^2}{a^2} - 1\right) \sin^2(\varphi)}}, \quad (52)$$

C. Derivation of $l(\varphi)$

The result can be derived using the following coordinate transformation:

$$\begin{bmatrix} x \\ y \end{bmatrix} = \begin{bmatrix} \cos(\varphi) & \sin(\varphi) \\ -\sin(\varphi) & \cos(\varphi) \end{bmatrix} \begin{bmatrix} x' \\ y' \end{bmatrix}.$$

From Fig. 13, we know that the effective length is $l = 2|y|_{x=0}$. Substituting $x = 0$ into the transformation equations, we obtain:

$$y' = -\cot(\varphi) x' \quad (53a)$$

$$y = (-x' \sin(\varphi) + y' \cos(\varphi)). \quad (53b)$$

Combining these two equations yields:

$$x' = \frac{-y \sin(\varphi)}{2}, \quad y' = \frac{y \cos(\varphi)}{2}. \quad (54)$$

Next, we substitute these expressions into the equation of the ellipse:

$$\frac{4x'^2}{a^2} + \frac{4y'^2}{b^2} = 1. \quad (55)$$

After substitution and simplification, we solve for y and find:

$$y = \pm \frac{ab}{2\sqrt{a^2 \cos^2(\varphi) + b^2 \sin^2(\varphi)}}. \quad (56)$$

Since $l = 2|y|$, the effective length becomes:

$$l = b \sqrt{\frac{1}{1 + \left(\frac{b^2}{a^2} - 1\right) \sin^2 \varphi}}. \quad (57)$$

This concludes the derivation of $l(\varphi)$.

D. Expansion of the Phase Shift $\omega(t)$

Using the result from the first part and Lemma 1, we expand $l(\varphi(z, t))$ around $\varphi(z, \infty)$ by exploiting Taylor series as:

$$l(\varphi(z, t)) = b \left(1 + \sum_{p=0}^{\infty} C_p (\varphi(z, t) - \varphi(z, \infty))^p \right). \quad (58)$$

where C_p are the coefficients of the expansion. Firstly, substitute (17) in (58). Secondly, substitute the result into (51) and interchange the summation and integral, we get:

$$\omega(t) = \ell_{lc} \sum_{p=0}^{\infty} e^{-\frac{pt}{\tau_{mol}}} \underbrace{\frac{C_p}{ad} \int_0^d (\varphi(z, 0) - \varphi(z, \infty))^p (\kappa_{lc} - \kappa) dz}_{\text{A function of } p}. \quad (59)$$

This finally yields to:

$$\omega(t) = \ell_{lc} \sum_{p=0}^{\infty} D_p e^{-\frac{pt}{\tau_{mol}}}, \quad (60)$$

where D_p are constant coefficients. This concludes the proof.

APPENDIX C PROOF OF LEMMA 2

By substituting the phase shifter vector ($\mathbf{v}$) in SNR definition (29), we have

$$\text{SNR}_k(\mathbf{u}_k^{(p)}) = \sum_{n=1}^N \sum_{m=1}^N e^{-j[\omega_k]_n [\mathbf{M}_k(\mathbf{u}_k^{(p)})]_{n,m}} e^{j[\omega_k]_m}. \quad (61)$$

Because LOS link is dominant on both BS-RIS and RIS-MU, we can decompose $[\mathbf{M}_k(\mathbf{u}_k^{(p)})]_{n,m} = e^{j[\phi_k(\mathbf{u}_k^{(p)})]_n} e^{-j[\phi_k(\mathbf{u}_k^{(p)})]_m}$ where phase shifter $[\phi_k(\mathbf{u}_k^{(n)})]_p, \forall n$ can be derived by (3) [30]. Let us expand SNR around the phase shifter $[\omega]_p$ where all other phase shifters for any $m \neq p$ are fixed. So

$$\begin{aligned} \text{SNR}_k(\mathbf{u}_k^{(p)}) &= e^{[\omega]_p - [\phi_k(\mathbf{u}_k^{(p)})]_p} \sum_{m \neq p} e^{-[\omega]_m + [\phi_k(\mathbf{u}_k^{(p)})]_m} \\ &\quad + e^{-[\omega]_p + [\phi_k(\mathbf{u}_k^{(p)})]_p} \sum_{m \neq p} e^{[\omega]_m - [\phi_k(\mathbf{u}_k^{(p)})]_m} \\ &\quad + 1 + \sum_{m \neq p} e^{[\omega]_m - [\phi_k(\mathbf{u}_k^{(p)})]_m} \end{aligned}$$

$$\sum_{m \neq p} e^{-[\omega]_m + [\phi_k(\mathbf{u}_k^{(p)})]_m}. \quad (62)$$

Let us define $[r_k]_p e^{j[\theta_k]_p} \triangleq \sum_{m \neq p} e^{[\omega]_m - [\phi_k(\mathbf{u}_k^{(p)})]_m}$, where $[r_k]_p$ and $[\theta_k]_p$ are amplitude and phase, respectively. For sufficiently large N , $[\theta_k]_p$ tends to be zero $\forall p$.⁸ Therefore, (62) can be decomposed as $c + [r_k]_p \cos([\omega_k]_p - [\phi_k(\mathbf{u}_k^{(n)})]_p)$ where c is a constant for k th user. Substituting this result into (35) gives us:

$$\begin{aligned} [L_k] &= C_k + [\xi_{k+1}]_p [\mathbf{t}_{k+1}]_p + [\xi_k]_p [\mathbf{t}_k]_p \\ &\quad - \underbrace{[\lambda_k]_p [r_k(\mathbf{u}_k^{(n)})]_p}_{\text{New } [\lambda_k]_p} \cos([\omega_k]_p - \phi_k(\mathbf{u}_k^{(p)})). \end{aligned} \quad (63)$$

This concludes the proof.

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⁸This result can be derived by evaluating the integral over $[0, 2\pi]$, $\forall m \neq p$. However, due to space constraints, the detailed derivation is omitted.

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