AoA-Based Low Complexity Beamforming for Aerial RIS Assisted Communications at mmWaves

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AoA-based Low Complexity Beamforming for Aerial RIS assisted Communications at mmWaves

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***Abstract*—Reconfigurable intelligent surface (RIS) mounted on mobile devices (aerial RIS), such as unmanned aerial vehicles, is envisioned to increase coverage for future millimeter wave (mmWave) communications. In literature, RIS beamformer is designed using channel state information (CSI) or by scanning through gNodeB and user codebooks over different training RIS weight matrices. Both of these methods incur significant training overhead. Also, the traditional subspace-based AoA estimators require several power-intensive radio frequency (RF) chains, which is not practically affordable to users. As a low-power and low-complexity alternative, we propose to design RIS beamforming weights based on angle-of-arrival (AoA) from gNodeB-to-RIS and from RIS-to-user with a single RF chain coupled to an antenna array. The proposed estimator is a combination of maximum likelihood and maximum correlation estimator and uses a single RIS training matrix. The simulation results show that, though a slightly larger number of antenna elements is required for the same spectral efficiency, the proposed approach offers a significantly energy- and computationally-efficient beamforming alternative compared to the existing subspace-based AoA estimator and CSI based techniques.**

***Index Terms*—Angle-of-arrival, energy efficiency, maximum likelihood, reconfigurable intelligent surface (RIS), beamforming.**

I. Introduction

Millimeter wave (mmWave) communication is an emerging 6th Generation technology that offers a wider spectrum with high capacity gains [1]. However, the mmWave channel has high signal attenuation and signals are easily blocked by obstacles. These problems can be addressed effectively by utilizing reconfigurable intelligent surfaces (RISs). A RIS can be configured to relay signals by avoiding obstacles between gNodeB (gNB) and users (UEs), thus providing a virtual line-of-sight (LoS) link and improving system's spectral efficiency (SE) at a lower capital cost [2]. RIS mounted on unmanned aerial vehicles (UAVs), also referred to as aerial RIS, will further provide an added degree of freedom and a better probability of LoS links to overcome the drawbacks of terrestrial mmWave communications [3], for example, to offload traffic from overloaded or damaged gNB. Also, owing to small wavelength at mmWaves, a large number of antenna elements can be packed in a small space, enabling the deployment of an energy-efficient aerial relay system aboard an aircraft.

A. Related work and motivation

Optimal design of RIS phase shifts is critical for maximizing the RIS performance gain. A proposed a deep learning-based approach was proposed in [4] for joint gNB transmit and RIS beamforming design. Setting the phase of RIS elements equal to the inverse of the phase of the cascaded gNB-RIS-UE channel is another method for designing RIS phase shifts. In [5], gNB-RIS channel was approximated as a rank one channel and then singular value decomposition (SVD) of the cascaded channel state information (CSI) was used to derive the gNB and RIS beamformer with the assumption of a single antenna UE. The computational complexity of RIS beamforming using SVD grows exponentially with the size of the channel matrix. Besides, estimation of the cascaded CSI is difficult and incurs a significant training overhead as it consists of estimation of all channel parameters such as angle-of-arrivals (AoAs) and propagation path gain products. In [6], atomic norm minimization was used to sequentially estimate the CSI. Here, the authors used traditional techniques such as multiple signal classification (MUSIC) to estimate AoAs. In [7], a parametric maximum likelihood (ML) channel estimation framework was proposed for estimating the LoS gNB-RIS channel with a reduced number of RIS training matrices. However, the performance of this parametric estimation method is strictly limited by the initial guess of the parameters.

mmWave channel sparsity in angular domain can be used to overcome high training costs of CSI based RIS beamforming. Besides, a preliminary gNB-RIS-UE link can be created using AoA information rather than waiting to acquire the complete CSI and then transmitting data. RIS phase shifts can then be updated by concurrently acquiring complete CSI and transmitting data to UE over the preliminary link. To that end, a discrete Fourier transform (DFT) codebook was used in [8] to estimate AoA to design RIS beamformer. In [9], RIS beamformer was designed based on the required angle of reflection to serve a UE. It was demonstrated that estimating the phase difference between two RIS elements is sufficient for designing the RIS beamformer. Moreover, a closed-form expression was derived for phase shift calculation using only three measurements. This study was limited to a single-path RIS to UE channel model as the phase estimation algorithm fails in a multipath environment. In [2], modified MUSIC based estimator was proposed to estimate AoAs. The limitation of MUSIC is that, for better accuracy a large number of radio frequency (RF) chains are required. Because mmWave RF chains waste a lot of power, decreasing number of RF chains at UE is critical [10]. Random beamforming was proposed in [11], which provides nearly acceptable long-run performance while requiring no CSI or AoA information.

B. Contribution and Significance

Although reducing the number of RF chains at mmWaves is crucial, estimating channel parameters with a limited number

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of RF chains is difficult. To this end, as an energy-efficient alternative, in this letter we propose a reduced complexity design of RIS beamformer and UE combiner based on AoA estimated using single RF chain at UE. The key contributions are as follows. (1) We present an aerial RIS-assisted mmWave communication system that improves spectral efficiency for serving the obstructed UEs with a ULA connected to a single RF chain, optimizing the RIS weight vector and UE steering vector based on estimated AoA. (2) We establish the gNB-RIS-UE link process in two steps: first estimating AoA at the UE using an ML estimator in conjunction with a maximum correlation estimator, and then optimizing RIS weights and UE beam steering vector by exploiting antenna geometry for maximum rate support. (3) We compare the achievable spectral efficiency, hardware power consumption, and computation complexity of our proposed RIS beamforming design with benchmark RIS beamforming designs through simulations. Our proposed technique offers a practical and affordable beamforming alternative for resource-constrained UEs at mmWaves compared to existing solutions in the literature.

II. System Model

Consider a wideband mmWave communication system consisting of a gNB, an aerial RIS (RIS mounted on UAV), and a mobile UE, as shown in Fig. 1. The direct link from gNB to UE is assumed to be in a complete outage. The aerial RIS is used to establish a link from gNB to UE. Let (x_g, y_g, z_g), (x_r, y_r, z_r), and (x_u, y_u, z_u) respectively denote the coordinates of gNB, aerial RIS, and UE. The gNB and aerial RIS coordinates are considered fixed, and the coordinates of UE are unknown. Both gNB and UE are equipped with a ULA placed along the z-axis having N_G and N_U antenna elements connected to a single RF unit, respectively. The RIS consists of a linear array of N_R reflecting elements placed along x-axis.

The wideband mmWave bandwidth B is divided into N_c subcarriers. Assuming pointed beam at gNB, gNB-aerial RIS channel on subcarrier n is modeled with a single LoS path as

$$\mathbf{H}[n] = \sqrt{N_G N_R G_0}\, h[n]\mathbf{a}_r(\theta_h)\mathbf{a}_t^H(\phi_h) \in \mathbb{C}^{N_R \times N_G} \quad (1)$$

where G_0 is per antenna element gain and $h[n] = \bar{h}e^{-j2\pi\Delta f \tau_h (n-1)}$. Here, $\bar{h}$ is complex channel gain with

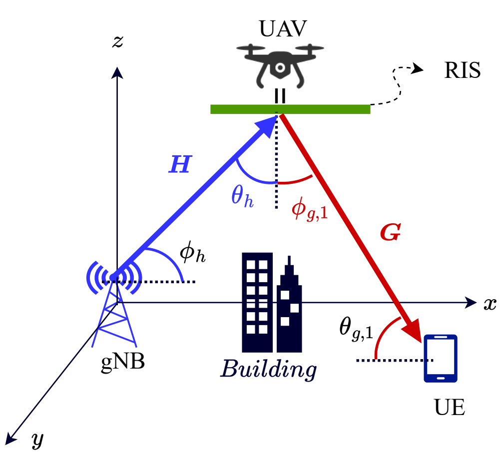

Figure 1: 3D view of placement of gNB, aerial RIS, and UE.

Ricean distribution, τ_h is time delay from gNB to RIS, Δf is subcarrier spacing, θ_h is AoA at aerial RIS from gNB, and ϕ_h is the angle-of-departure (AoD) from gNB. $\mathbf{a}_t(\cdot)$ and $\mathbf{a}_r(\cdot)$ denote transmit and receive array factor, respectively. In general, the array response vector at an offset angle ϕ from the broadside of a ULA having N antenna elements is

$$\mathbf{a}(\phi) = \frac{1}{\sqrt{N}}\left[1, e^{-j\frac{2\pi d}{\lambda}\sin\phi}, \ldots, e^{-j\frac{2\pi d}{\lambda}(N-1)\sin\phi}\right]^T \quad (2)$$

where d is inter-element spacing and λ is carrier wavelength. Similarly, aerial RIS to UE channel over subcarrier n is

$$\mathbf{G}[n] = \sum_{l=1}^{L}\sqrt{N_U N_R G_0}\, g_l[n]\mathbf{a}_r(\theta_{g,l})\mathbf{a}_t^H(\phi_{g,l}) \in \mathbb{C}^{N_U \times N_R} \quad (3)$$

where $g_l[n] = \bar{g}_l e^{-j2\pi\Delta f \tau_{g,l}(n-1)}$. Here, $\bar{g}_l$ is complex channel gain parameter, $\tau_{g,l}$ is time delay from RIS to UE, $\theta_{g,l}$ is AoA at the user, and $\phi_{g,l}$ is AoD from RIS, respectively, of l^{th} multipath. L is the total number of multipath components, out of which 1 is LoS with Ricean fading and remaining are non-LoS with Rayleigh fading. Further, from Fig. 1, for LoS component $L = 1$, $\theta_h = 90^o - \phi_h$ and $\phi_{g,1} = 90^o - \theta_{g,1}$. Let $\phi_h = \phi$ and $\theta_{g,1} = \theta_1$, then

$$\begin{aligned} \mathbf{H}[n] &= \sqrt{N_G N_R G_0}\, h[n]\mathbf{a}_r(90^o - \phi)\mathbf{a}_t^H(\phi) \\ \mathbf{G}[n] &= g_1[n]\mathbf{a}_r(\theta_1)\mathbf{a}_t^H(90^o - \theta_1) \\ &\quad + \sum_{l=2}^{L}\sqrt{N_U N_R G_0}\, g_l[n]\mathbf{a}_r(\theta_{g,l})\mathbf{a}_t^H(\phi_{g,l}). \end{aligned} \quad (4)$$

III. Spectral Efficiency (SE) Analysis

Let $\omega_m \in \mathcal{P} \equiv \{2p\pi/b_R | p = 0, 1, \ldots, (b_R - 1)\}$ denote the phase shift associated the m^{th} passive RIS element. Here $m = \{1, \ldots, N_R\}$ and b_R is the number of bits controlling possible discrete phase shifts at each RIS element. We define

$$\mathbf{\Theta} = \mathrm{diag}(e^{j\omega_1}, \ldots, e^{j\omega_{N_R}}) \in \mathbb{C}^{N_R \times N_R}. \quad (5)$$

Let $\mathbf{b} \in \mathbb{C}^{N_T \times 1}$ denote the analog steering vector of the gNB steered at an offset angle Ω_g given by

$$\mathbf{b} = \frac{1}{\sqrt{N_T}}\left[1, e^{-j\frac{2\pi d}{\lambda}\sin\Omega_g}, \ldots, e^{-j\frac{2\pi d}{\lambda}(N_T-1)\sin\Omega_g}\right]^T. \quad (6)$$

Further, since UAV is controlled by a ground controller, here the gNB, its location is assumed to be known at the gNB. Thus, gNB steers its beam towards the LoS link to RIS, i.e.,

$$\Omega_g = \tan\left(\sqrt{(x_r - x_g)^2 + (y_r - y_g)^2}/(z_r - z_g)\right)^{-1}. \quad (7)$$

Also, let $\mathbf{v}(\Omega) \in \mathbb{C}^{N_U \times 1}$ denote the analog steering vector of UE ULA steered at an offset angle Ω given as

$$\mathbf{v}(\Omega) = \frac{1}{\sqrt{N_U}}\left[1, e^{-j\frac{2\pi d}{\lambda}\sin\Omega}, \ldots, e^{-j\frac{2\pi d}{\lambda}(N_U-1)\sin\Omega}\right]^T. \quad (8)$$

Then the signal received at UE over subcarrier n is

$$y[n] = \mathbf{v}^H(\Omega)\mathbf{G}[n]\mathbf{\Theta}\mathbf{H}[n]\mathbf{b}x[n] + w[n] \quad (9)$$

where $x[n]$ is the transmitted signal of power P on subcarrier n and $w[n] \sim \mathcal{CN}(0, \sigma^2)$ is additive complex white Gaussian noise with zero mean and σ^2 variance. Then, SE is given as

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$$\text{SE} = \sum_{n=1}^{Nc} \log_2\left(1 + |y[n]|^2/\sigma^2\right) \text{ bits/s/Hz}. \tag{10}$$

Let P_{LNA}, P_{LPF}, P_{ADC}, and P_{LO} denote power of low noise amplifier, low pass filter, analog-to-digital converter, and local oscillator, respectively. Then power consumption in N_{RF} RF chains connected to N_U antenna elements at UE is [12]

$$P_{RF} = N_{RF}(P_{ADC} + P_{LPF}) + N_U P_{LNA} + P_{LO} \text{ mW}. \tag{11}$$

IV. Beamforming Design

The optimizing variables in (9) are $\mathbf{\Theta}$ and $\mathbf{v}$. Further, given gNB-to-RIS link information, the process of establishing a gNB-to-UE path via RIS involves two stages. The first stage is the estimation of AoA at the UE using a single RIS training matrix $\mathbf{\Theta}^t$. The second stage is the design of optimal RIS phase shifts $\mathbf{\Theta}^*$ and a steering vector $\mathbf{v}^*$ at the UE by utilizing the estimated AoA. Additionally, the effectiveness of the proposed estimator is demonstrated through the derivation of the Cramer-Rao lower bound (CRLB).

A. AoA estimation

This section deals with estimation of AoA at UE, i.e, $\hat{\theta}$. Let $x[n] = \text{x}\ \forall n$ where x is a pilot signal of power P and $\mathbf{\Theta}^t$ be the RIS training matrix. Then the signal received by UE beam steered at an angle Ω_j on subcarrier n is given as

$$y_j[n] = \mathbf{v}^H(\Omega_j)\mathbf{G}[n]\mathbf{\Theta}^t\mathbf{H}[n]\mathbf{b}\text{x} + w[n] \tag{12}$$

Let $\alpha_l[n] = N_R G_0 \sqrt{N_U N_G} h[n] g_l[n]$ and $\boldsymbol{\alpha}_l = [\alpha_l[1], \ldots \alpha_l[N_c]]$, then the received signal vector over all N_c subcarriers is

$$\mathbf{y}_j = \mathbf{v}^H(\Omega_j)\mathbf{a}_r(\theta_1)\mathbf{a}_t^H(90^o - \theta_1)\mathbf{\Theta}^t\mathbf{a}_r(90^o - \phi)\mathbf{a}_t^H(\phi)\mathbf{b}\boldsymbol{\alpha}_1\text{x} + \sum_{l=2}^{L}\mathbf{v}^H(\Omega_j)\mathbf{a}_r(\theta_{g,l})\mathbf{a}_t^H(\phi_{g,l})\mathbf{\Theta}^t\mathbf{a}_r(90^o - \phi)\mathbf{a}_t^H(\phi)\mathbf{b}\boldsymbol{\alpha}_l\text{x} + \mathbf{w}. \tag{13}$$

Consider that the UE sweeps the area by steering its beam sequentially in M directions. The UE codebook consists of vectors corresponding to steering angles in set $\mathbf{\Omega} = \{0, \ldots, (j-1)90^o/2^b, \ldots, 90^o\}$, where b is the number of bits controlling the size of UE codebook. Then the collection of received signals from all M directions is

$$\begin{aligned}\mathbf{Y} &= \mathbf{V}^H\mathbf{a}_r(\theta_1)\mathbf{a}_t^H(90^o - \theta_1)\mathbf{\Theta}^t\mathbf{a}_r(90^o - \phi)\mathbf{a}_t^H(\phi)\mathbf{b}\boldsymbol{\alpha}_1\text{x} \\ &\quad + \sum_{l=2}^{L}\mathbf{V}^H\mathbf{a}_r(\theta_{g,l})\mathbf{a}_t^H(\phi_{g,l})\mathbf{\Theta}^t\mathbf{a}_r(90^o - \phi)\mathbf{a}_t^H(\phi)\mathbf{b}\boldsymbol{\alpha}_l\text{x} + \mathbf{W} \\ &= \mathbf{d}(\theta_1, \phi)\boldsymbol{\alpha}_1 + \sum_{l=2}^{L}\mathbf{d}(\theta_{g,l}, \phi)\boldsymbol{\alpha}_l + \mathbf{W} \in \mathbb{C}^{M \times N_c}.\end{aligned} \tag{14}$$

Here $\mathbf{V} = [\mathbf{v}(\Omega_1), \ldots, \mathbf{v}(\Omega_M)]$ denotes UE codebook and $\mathbf{d}(\theta, \phi) = \mathbf{V}^H\mathbf{a}_r(\theta)\mathbf{a}_t^H(90^o - \theta)\mathbf{\Theta}^t\mathbf{a}_r(90^o - \phi)\mathbf{a}_t^H(\phi)\mathbf{b}\text{x}$ denotes dictionary matrix for LoS link. The $\hat{\theta}$ of LoS link of RIS-UE channel is estimated using ML estimator as

$$(\mathcal{P}1): \hat{\theta} = \underset{\theta, \phi \in [0, \pi/2]}{\text{argmin}} \|\mathbf{Y} - \mathbf{d}(\theta, \phi)\mathbf{d}^\dagger(\theta, \phi)\mathbf{Y}\|^2. \tag{15}$$

Here, $\mathbf{d}^\dagger$ represents Moore–Penrose inverse of $\mathbf{d}$. Problem $\mathcal{P}1$ is a non-convex problem sensitive to initial points. Therefore, for an initial guess, we choose the points $\{\theta_0, \phi_0\}$ that gives maximum correlation with an atom of dictionary $\mathbf{d}$, i.e,

$$(\mathcal{P}2): \{\theta_0, \phi_0\} = \underset{\theta, \phi \in [0, \pi/2]}{\text{argmin}} \|\mathbf{d}^H(\theta, \phi)\mathbf{Y}\|^2. \tag{16}$$

B. RIS training matrix design

To estimate $\hat{\theta}$ from the cascaded gNB-RIS-UE channel, the RIS training matrix $\mathbf{\Theta}^t$ needs to designed optimally to minimize AoA estimation error while reducing training overhead. Here t is the number of RIS training matrices. There can be $t = (2^{b_R})^{N_R}$ possible combinations of $\mathbf{\Theta}^t$ when each RIS element is controlled b_R bits. Therefore, to reduce the training overhead we propose to use only one RIS training matrix, i.e., $t = 1$. Without any additional phase shifts at RIS, the incident rays arriving at an angle ϕ will reflect at the same angle from the RIS. Therefore, the phase shift between the incident and the reflected ray provided by m^{th} RIS element is $\beta_m = (2\pi d/\lambda)\sin\phi(m-1)$. The reflection angle at RIS elements can be altered to deflect the incident rays from individual RIS elements into the desired direction θ_m. Therefore, an additional phase required at m^{th} element to focus the incoming signal in the desired direction is [9]

$$\omega_m^t = \frac{2\pi d}{\lambda}(m-1)(\sin\phi - \sin\theta_m) = \frac{2\pi d}{\lambda}(m-1)\sin\tilde{\theta}_m. \tag{17}$$

Here, $\{\tilde{\theta}_m\}_{\forall m}$ have normal truncated Gaussian distribution in the range $[-\pi/2, \pi/2]$. Thus, $\mathbf{\Theta}^t = \text{diag}\{e^{\omega_1^t}, \ldots, e^{\omega_{N_R}^t}\}$ is used for AoA estimation in Section IV-A.

C. RIS beamforming and UE beamforming

Next, from estimated AoA, we design RIS and UE beamformer. From (17), the optimal additional phase shift required to converge the arriving signal at RIS from an angle $90^0 - \phi$ to the desired AoD at $90^o - \hat{\theta}$ is

$$\omega_m^* = (2\pi d/\lambda)(\cos\phi - \cos\hat{\theta})(m-1). \tag{18}$$

The UE should steer its beam towards $\hat{\theta}$ to align it with the RIS beamformed signal. Therefore, $\mathbf{v}^* = \mathbf{a}(\hat{\theta})$.

D. CRLB derivation

Here, we derive the CRLB of the proposed estimator. During AoA estimation, the vector of signals received from M sweep directions over n^{th} subcarrier is

$$\mathbf{y}[n] = \sum_{l=1}^{L}\mathbf{V}^H\mathbf{a}_r(\theta_{g,l})\mathbf{a}_t^H(\phi_{g,l})\mathbf{\Theta}^t\mathbf{a}_r(90^o - \phi)\mathbf{a}_t^H(\phi)\mathbf{b}\alpha_l[n]\text{x} + \mathbf{w}[n] = \boldsymbol{\mu}[n] + \mathbf{w}[n] \tag{19}$$

where $\boldsymbol{\mu}[n]$ denotes noiseless part of signal. Given $\Omega_g = \phi$, substituting $\mathbf{a}_t^H(\phi)\mathbf{b} = \mathbf{I}_{N_G}$ in (19), we have

$$\boldsymbol{\mu}[n] = \sum_{l=1}^{L}\mathbf{V}^H\mathbf{a}_r(\theta_{g,l})\mathbf{a}_t^H(\phi_{g,l})\mathbf{\Theta}^t\mathbf{a}_r(90^o - \phi)\alpha_l[n]\text{x}. \tag{20}$$

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$$\mu_j[n]=\sum_{l=1}^{L}\frac{\alpha_l[n]x}{N_UN_R}\sum_{n=1}^{N_U}e^{-j\pi(n-1)(\sin\theta_{g,l}-\sin\Omega_j)}\sum_{m=1}^{N_R}e^{-j\left(\pi(m-1)(\cos\phi-\sin\phi_{g,l})-\omega_m^t\right)}=\sum_{l=1}^{L}\frac{\alpha[n]_lx}{N_UN_R}\bar{\mu}_{1,l}\bar{\mu}_{2,l} \tag{21}$$

$$\frac{\partial[\mu_j[n]]_s}{\partial\theta_{g,l}}=\frac{-j\pi\alpha_l[n]x}{N_UN_R}\left(\bar{\mu}_{2,l}\sum_{n=1}^{N_U}(n-1)\cos\theta_{g,l}e^{-j\pi(n-1)(\sin\theta_{g,l}-\sin\Omega_j)}+\bar{\mu}_{1,l}\sum_{m=1}^{N_R}(m-1)\nu_{g,l}\cos\theta_{g,l}e^{-j\left(\pi(m-1)(\cos\phi-\sin\phi_{g,l})-\omega_m^t\right)}\right) \tag{22}$$

$$\frac{\partial[\mu_j[n]]_s}{\partial\phi}=\sum_{l=1}^{L}\frac{j\pi\alpha_l[n]x\sin\phi}{N_UN_R}\bar{\mu}_{1,l}\sum_{m=1}^{N_R}(m-1)e^{-j\left(\pi(m-1)(\cos\phi-\sin\phi_{g,l})-\omega_m^t\right)} \tag{23}$$

For UE beam steered at Ω_j, the j^{th} row of vector $\boldsymbol{\mu}[n]$ is given by (21). Let the vector of unknown parameters be given by $\boldsymbol{\zeta}=[\theta_1,\ldots,\theta_L,\phi,\{Re\{\alpha_l\}\},\{Im\{\alpha_l\}\}]$. Thus, for M snapshots, corresponding to different steering directions, the Fisher information matrix $\boldsymbol{J}(\boldsymbol{\zeta})$ is defined as

$$\boldsymbol{J}(\boldsymbol{\zeta})=\frac{2}{\sigma^2}\sum_{s=1}^{M}\sum_{n=1}^{N_c}Re\left\{\frac{\partial[\boldsymbol{\mu}[n]]s}{\partial\boldsymbol{\zeta}}\left(\frac{\partial[\boldsymbol{\mu}[n]]_s}{\partial\boldsymbol{\zeta}}\right)^H\right\} \tag{24}$$

where the derivative of $\boldsymbol{\mu}$ with respect to θ and ϕ are given by (22) and (23), respectively. In (22) $\nu_{g,l}=1$ for $l=1$ and 0 otherwise. Additionally,

$$\begin{aligned}\frac{\partial[\boldsymbol{\mu}[n]]_s}{\partial Re(\alpha_l[n])}&=\mathbf{V}^H\mathbf{a}_r(\theta_{g,l})\mathbf{a}_t^H(\theta_{g,l})\boldsymbol{\Theta}^t\mathbf{a}_t(90^o-\phi)\mathrm{x}\\ \frac{\partial[\boldsymbol{\mu}[n]]_s}{\partial Im(\alpha_l[n])}&=j\mathbf{V}^H\mathbf{a}_r(\theta_{g,l})\mathbf{a}_t^H(\phi_{g,l})\boldsymbol{\Theta}^t\mathbf{a}_R(90^o-\phi)\mathrm{x}.\end{aligned} \tag{25}$$

Using (24), CRLB$(\hat{\theta})=\sqrt{\mathrm{diag}[\boldsymbol{J}^{-1}(\boldsymbol{\zeta})]_{1,1}}$.

V. Complexity Analysis and Training Overhead

This section compares the computational complexity of RIS beamforming design based on CSI and RIS design based on proposed estimated AoA at the UE.

The cascaded CSI estimation process consists of estimation of AoA at the UE, AoA at the RIS, and cascaded path gains. For each pilot transmission, the AoA estimated using the MUSIC algorithm has a computational complexity of $\mathcal{O}((1+N_U)^{3.5})$ [6]. The path gains are estimated using least square method that has a complexity of $\mathcal{O}(Z(Lt^2+LN_Rt+t^3))$ [6], where Z denotes the number of iterations required for convergence of estimator and t denotes the number of pilot transmissions. Thereafter, RIS phase shifts are determined using SVD of $\mathbf{HG}$ that has a computational complexity of $\mathcal{O}(N_R^2N_U)$. Thus, the overall complexity of RIS beamforming using CSI based approach is $\mathcal{O}(N_c(1+N_U)^{3.5}+Z(Lt^2+LN_Rt+t^3)+N_R^2N_U)$. On contrary, the proposed approach estimates AoA using one RIS training matrix. The complexity of $\mathcal{P}1$ is upper bounded by the complexity of determining $\mathbf{d}^\dagger\in\mathbb{C}^{M\times1}$ which is $\mathcal{O}(M^3+M^2+MN_U+N_R)$. Also, $\mathcal{P}2$ has complexity $\mathcal{O}(MN_c)$. Therefore, the total complexity of the proposed method is $\mathcal{O}(Z(M^3+M^2+MN_U+N_R)+Z'MN_c)$, where Z and Z' respectively denote the number of iterations required for convergence of $\mathcal{P}1$ and $\mathcal{P}2$. Also, Table I compares the training overhead of parameter estimation process of proposed AoA based RIS beamforming method with two other CSI based benchmarks.

Table I: Training overhead in parameter estimation

Scheme	Training overhead
Proposed	$\lceil M/N_{RF}\rceil$
Atomic norm minimization based CSI estimation [6]	$\lceil M/N_{RF}\rceil+t\lceil L/N_{RF}\rceil; t\geq 30$
ML based CSI estimation [7]	$t\lceil M/N_{RF}\rceil; t\geq 10$

VI. Result and Discussion

This section demonstrates the performance of proposed RIS beamforming and UE steering design based on the estimated AoA at UE. The simulation parameters are as follows: carrier frequency $f_c=28$ GHz, $N_c=100$, $P=0.6$ W, $\Delta f=30$ KHz, $\sigma^2=N_o\Delta f$, $N_o=-184$ dBm/s/Hz, $N_G=64$, $G_0=9$ dBi, $N_R=200$, $d=\lambda/2$, $[x_g,y_g,z_g]=[0,0,10]$ m, $[x_r,y_r,z_r]=[150,0,80]$ m, $[x_u,y_u,z_u]=[180,20,2]$ m, $b_R=5, b=6$, $L=3$, and Ricean parameter $\kappa=18$ dB.

The root mean squared error (RMSE) of $\hat{\theta}$ obtained using proposed estimator and corresponding CRLB as a function transmit power P are shown in Fig. 2(a). The value of RMSE decreases with increasing gNB transmit power and N_U. Furthermore, because of the non-convex nature of $\mathcal{P}1$ and high error in initial AoA estimate using the maximum correlation estimator, there is a large gap between RMSE and CRLB values. We have also compared the RMSE of the proposed estimator with that of MUSIC estimator employing $N_{RF}=N_U$ to estimate AoA [2]. Though the RMSE of MUSIC estimator is small, deployment of such subspace-based estimators are not suitable at mmWaves because of requirement of large number of power-hungry RF chains, as shown in Fig. 2(b). Further, Fig. 2(c) and Fig. 2(d) compares the computational cost of the RIS beamforming along with computations required for parameter estimation of the proposed AoA based design to that of the benchmarks. It is observed that the complexity of AoA estimation using MUSIC grows exponentially as function of N_U. Also, the complexity of CSI based approach grows rapidly as N_R increases due to the increased computations required in SVD of cascaded gNB-RIS-UE channel.

Fig. 3 compares the SE achieved using the proposed beamforming design with the competitive AoA based benchmarks for designing RIS and UE beamformers: random RIS beamforming [11], beamforming based on AoA estimated by searching through DFT codebook at UE [8], beamforming based on perfect AoA information, and beamforming based on AoA estimated using MUSIC estimator (considering single RF chain for data communication) [2]. As observed in Fig.

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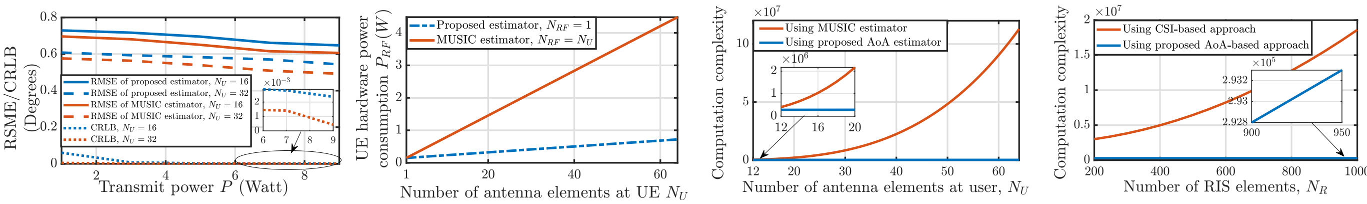

(a) (b) (c) (d)

Figure 2: (a) Comparison of RMSE of $\hat{\theta}$ with CRLB. (b) Comparison of power consumption in RF chains at UE using proposed estimator versus MUSIC estimator (using (11)). (c) Comparison of computational complexity of using the proposed estimator over MUSIC [2] in an AoA based approach as a function of N_U, with $N_R = 200$. (d) Comparison of the computational complexity of beamforming using the proposed AoA against the CSI based approach [6] as a function of N_R, with $N_U = 16$.

3, beamforming at RIS and UE with $\hat{\theta}$ gives a better SE compared to the cases when AoA estimation is done at the UE by merely sweeping beam in the UE codebook directions or designing RIS beamformer randomly. Since MUSIC employs a high number of RF chains and hence has a reduced AoA estimation error, it outperforms the proposed AoA estimator for the same number of N_U. However, as noted from Figs. 2(b)-(c), the power and hardware complexity of the proposed methodology over MUSIC estimator is significantly lower. It is also notable that SE obtained by designing beamformers based on AoA estimated using MUSIC and the proposed estimator improves with increasing N_U values due to a drop in RMSE. Furthermore, Fig. 3 provides a benchmark comparison with CSI based RIS and UE beamforming. *It should be noted that by increasing the number of antenna elements N_U connected to a single RF chain, our proposed system can achieve the same SE as that using CSI based technique.*

VII. Conclusion

In this letter to address the difficulty of estimating channel parameters with a limited number of RF chains, which is crucial at mmWaves, we proposed an energy-efficient and reduced complexity design of RIS beamformer and UE combiner based on AoA estimated using a single RF chain at UE. This is in contrast to the CSI-based approach for RIS beamforming and UE combiner which has high computational cost and training overhead of CSI estimation and SVD of cascaded channel. Though subspace-based estimator, such as MUSIC, offers smaller RMSE, it cannot be employed in a system with single RF chain. Instead, in our approach we proposed an estimator to estimate AoA at the UE utilizing single RF chain and using only one RIS training matrix. Simulation-based performance results demonstrated that our approach achieved an improved SE compared to the cases when AoA is estimated using DFT codebook at UE and when RIS weights are designed randomly. The proposed AoA based RIS beamforming using single RF chain connected to a large antenna array can be considered as a significantly cost-efficient alternative to the existing approaches in the literature with comparable SE performance. Future work includes improving AoA estimation accuracy using multiple RIS training matrices and other prior information, and extending the proposed method to include multiple RF chains at the UE connected to a large antenna array in a sub-array hybrid architecture.

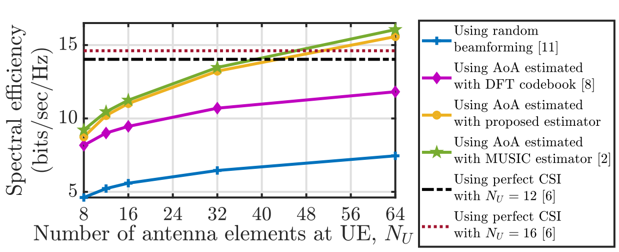

Figure 3: Comparison of SE achieved using the proposed AoA based RIS beamforming with the benchmark schemes given $N_R = 200$.

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