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Creating Spatial Degrees of Freedom for Long-Range LoS MIMO using Reflect-arrays

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***Abstract*—A fundamental bottleneck for long-range Line-of-Sight (LoS) MIMO is transceiver form factor: for a given carrier frequency, the product of the transmit and receiver apertures required to sustain a given number of spatial degrees of freedom scales quadratically with the link distance. In this paper, we propose and investigate the feasibility of a novel approach for sidestepping this bottleneck by creating large effective apertures using multiple reflect-arrays (RAs) placed near the transceivers. The introduction of an RA between a transmitter and receiver results in an end-to-end gain scaling as $1/d_1^2 d_2^2$, where d_1 is the distance between the transmitter and RA and d_2 is the distance between the RA and the receiver. While this leads to poor scaling when d_1 and d_2 are both comparable to the link range, this problem is alleviated in our proposed setting, in which RAs are deployed near the transmitter and receiver. By benchmarking the link budget of our system model against a SISO link using transceivers with comparable form factors, we provide analytical guidelines for choice of system parameters such as the required RA sizes and the distance of the RAs from the associated transceivers. Simulation results based on detailed modeling of the channel matrices validate our analytical framework. Two key findings are as follows: (a) in order to avoid excessive degradation in link budget relative to the SISO benchmark, we must deploy "large enough" RAs at "small enough" distances making near-field beam focusing between RAs and transceivers essential, (b) coarse (2-bit) quantization of RA phases suffices to implement the required combination of near-field focusing and long-range beamforming with relatively small performance loss. We illustrate our ideas for a 6.4 Gbps link at 1.5 km using 4-fold spatial multiplexing and 800 MHz of bandwidth operating at 28 GHz.**

***Index Terms*—mmWave, LoS MIMO, relect-array , link budget, spatial degrees of freedom, intelligent reflecting surfaces**

I. Introduction

Line of sight (LoS) multi-input multi-output (MIMO) systems in mmWave bands have received significant recent attention due to their attractive (potentially cubic) scaling properties with increase in carrier frequency f_c: the available bandwidth typically scales linearly with f_c, while for a given link distance D, the available spatial degrees of freedom (DoF) for two-dimensional (2D) apertures scales with f_c^2. Specifically, we have [1]

$$DoF_{2D} \approx \frac{A_T A_R}{D^2 \lambda^2} + 1 \tag{1}$$

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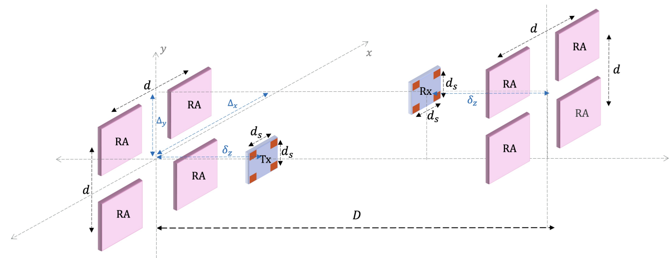

Fig. 1. 4-fold spatially multiplexed reflect-array aided LoS MIMO link

for transmit and receive apertures A_T and A_R, respectively. However, in a standard LoS MIMO system in which these apertures are constrained by transceiver form factors, the available DoF decreases rapidly with link range.

In this paper, we propose a novel approach for long-range LoS MIMO by using reflect-arrays (RAs) [2] placed near the transceivers to synthesize large effective apertures that are no longer constrained by transceiver form factors. An RA is typically a planar surface composed of a large number of sub-λ-sized passive reflecting elements, each with the capability to independently induce a phase (and possibly amplitude) change to the incident signal to suitably direct the reflected wave. By scaling to a large number of elements while eliminating RF chains, RAs have immense potential in terms of enabling system designers to "shape" RF environments at low cost and energy consumption [3]. [1] Our running example for illustrating the proposed concept is a 4-fold spatially multiplexed LoS MIMO system operating at 28 GHz at a link distance of 1.5 km; with 800 MHz bandwidth and Quadrature Phase Shift Keying (QPSK) modulation, this yields data rates in excess of 5 Gbps (after accounting for excess bandwidth and lightweight error control coding). However, a conventional Rayleigh-spaced [1] design using equal apertures at each end requires antenna spacings of 2.84 m for a well-conditioned spatial channel, which would be bulky and expensive. Instead, we propose placing four 2D RAs close to each transceiver (see Figure 1), spaced so as to synthesize large enough virtual apertures while maintaining compact transceiver form factors (e.g., existing 28 GHz platforms with four subarrays for the latter [4]).

[1] RAs are referred to by multiple different names in the literature including intelligent reflecting surfaces (IRS) and metasurfaces.

Our proposed approach represents a "sweet spot" for utilizing RAs in terms of link budget, which is a crucial concern since we target long ranges. The introduction of a RA in a link between transmitter and receiver results in power gain scaling as $1/d_1^2 d_2^2$, where d_1 is the distance between the transmitter and RA and d_2 is the distance between the RA and receiver, with worst-case scaling of $1/D^4$ if d_1 and, d_2 are comparable to the link range D. We largely circumvent this poor scaling by placing RAs close to the transceivers: $d_1 \ll R$ and $d_2 \approx R$ for the RAs creating the virtual transmit aperture, and $d_1 \approx R$ and $d_2 \ll R$ for the RAs creating the virtual receive aperture. By benchmarking our RA-enabled LoS MIMO system against a SISO link with comparable transceiver form factors, we provide analytical design guidelines on RA sizes and phase compensation strategies required to forestall substantial decline in link budget relative to the SISO benchmark. These are supported by BER simulations using a detailed model for the "3-hop" spatial channel. Key findings are as follows: (1) relatively large RA sizes (e.g., 1024 elements) placed at relatively short distances (e.g., 1 m) from the transceivers, requires that the phase settings at each RA must account for the curvature of the wavefront on the "short hop" to the nearby transceiver via *beam focusing* [5] along with the linear phase profiles required for long-range beamforming towards the RAs at the other end; (2) Given the large number of elements in each RA, coarse 2-bit phase quantization incurs negligible performance loss under QPSK modulation.

Related work: While the use of RAs to create spatial DoF has been considered in several recent works [6]–[9], to the best of our knowledge, there is no prior work that enables creation of multiple spatial DoF at long ranges, accounting for the associated link budget considerations. In [6]–[8], RAs are placed *between* transmitter and receiver to create spatial DoF. Unlike our long-range design, the distances of the RAs from transmitter and receiver are comparable, resulting in $O(1/D^4)$ path loss. Thus, these approaches are more appropriate for short ranges. A double-RA design, placing one RA close to the transmitter and receiver, respectively, is considered in [9]. For large enough RAs, it is possible to obtain more than one spatial mode with this approach, but the additional modes are significantly weaker than the dominant mode, and utilizing them requires solving a non-convex optimization problem to maximize the composite channel capacity. In contrast, our approach, by using multiple RAs near the transceiver to create a large virtual aperture, yields multiple spatial modes of comparable strength which can be utilized via phase compensation strategies with simple geometric characterization.

II. System Model

We consider a 4-fold spatially multiplexed LoS MIMO system operating at a carrier frequency of $f_c = 28$ GHz. The transmitter and receiver contain $N_t = N_r = 4$ subarrays, each of which is a square uniform planar array (UPA) containing $n_t = 64$ and $n_r = 64$ elements, respectively. The subarrays are placed at the four corners of a square aperture of side $d_s = 17.5$ cm as shown in Figure 1. A total of $N_t = 4$ RAs containing M reflecting elements each are placed behind the transmitter, one for each transmit subarray. A similar geometry is replicated at the receiver. Each RA is associated with a designated subarray in its nearby transceiver. The required separation between the RAs to support 4 spatial modes at link range $D = 1,500$ m is $d = 2.84$ m. Each subarray at the transmitter sends a unique QPSK stream to its corresponding RA. The RAs at the transmitter side then beamform towards the receiver. Each RA at the transmitter must compensate for two different phase profiles: first, the phase incurred at the RA from its designated transmit subarray, and second, the phase compensation needed to beamform towards the RAs on the receiver side. On the receive side, each RA receives a linear combination of signals sent by the 4 transmit RAs, and sends its received signal back to its designated receive subarray. Thus, each receive RA must apply a phase compensation strategy that includes beamforming towards the distant transmit RAs, and the phase profile corresponding to its designated receive subarray.

The RA planes at the transmitter and receiver side are placed at a distance δ_z from the transceivers. We refer to δ_z as the depth parameter and vary its value between 1 m and 5 m. The RAs are square with an inter-element spacing of $\lambda/2$ and the number of reflecting elements is varied between $M = 64, 256, 1024$ and 4096. We assume perfect channel state information at both transmitter and receiver.

We now characterize the channel matrices for our system. Define $\mathcal{N}_t = \{1, \ldots, N_t\}$, $\mathcal{N}_r = \{1, \ldots, N_r\}$, $\mathcal{N} = \{1, \ldots, n\}$ and, $\mathcal{M} = \{1, \ldots, M\}$ as the sets containing the N_t transmit subarrays, the N_r receive subarrays, the $n = n_t = n_r$ elements in each subarray and the M reflecting elements in each RA, respectively. Let $\mathbf{T}_k \in \mathbb{C}^{M \times n_t}$ and $\mathbf{R}_k \in \mathbb{C}^{M \times n_r}$ denote the LoS channel between the transmit subarray-RA pair and the LoS channel between the receive subarray-RA for stream $k \in \mathcal{N}_t$, respectively. Define $\mathbf{S}_k \in \mathbb{C}^{M \times 1}$ as the complex phase vector corresponding to the phase to be compensated by the RA paired to stream k at the transmit side and $\mathbf{S}_D \in \mathbb{C}^{M \times 1}$ as the complex UPA steering vector at each RA to compensate for the phase required to beamform towards the RAs on the receiver side. Let, $\mathbf{L} \in \mathbb{C}^{MN_r \times MN_t}$ be the LoS channel between the transmit and receive side RAs. The three channels $(\mathbf{T}_k, \mathbf{R}_k$ and $\mathbf{L})$ can then be expressed as

$$\left.\begin{aligned} T_k(l,i) &= \frac{\lambda\sqrt{G_T G_L}}{4\pi d_{Tk}(l,i)} e^{-j(2\pi d_{Tk}(l,i)/\lambda)} \\ &\qquad\qquad l \in \mathcal{M}, i \in \mathcal{N} \\ R_k(l,i) &= \frac{\lambda\sqrt{G_R G_L}}{4\pi d_{Rk}(l,i)} e^{-j(2\pi d_{Rk}(l,i)/\lambda)} \\ &\qquad\qquad l \in \mathcal{M}, i \in \mathcal{N} \\ L(m,l) &= \frac{\lambda\sqrt{G_L G_L}}{4\pi d_L(m,l)} e^{-j(2\pi d_L(m,l)/\lambda)} \\ &\qquad\qquad l = \{1, \ldots, MN_r\}, \\ &\qquad\qquad m = \{1, \ldots, MN_t\} \end{aligned}\right\} \tag{2}$$

where, $G_T/G_R/G_L$ are the gains per transmit, receive and reflecting element respectively. The distances $d_{Tk}(l,i)$ and $d_{Rk}(l,i)$ are the Euclidean distance between reflecting element l and associated transmit and receive subarray element i for

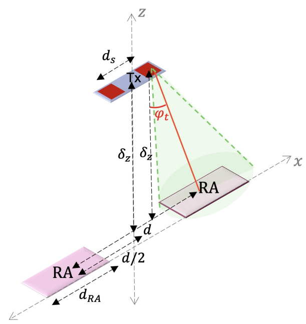

Fig. 2. 8×8 subarray at transmitter sending a unique stream to corresponding RA

stream k and $d_L(m,l)$ is the Euclidean distance between the RA reflecting elements l and m on the receive and transmit sides respectively. Let $\mathbf{a}_t(\theta_t, \phi_t) \in \mathbb{C}^{n \times 1}$ be the UPA steering vector [10] employed by each transmit subarray where, θ_t and ϕ_t are the elevation and azimuth steering angles required to steer the beam towards the associated RA. Figure 2 depicts the azimuth steering angle ϕ_t for one such transmit subarray and RA pair. By the symmetry of our system model, $\theta_t = \phi_t$ and $\mathbf{a}_t(\theta_t, \phi_t)$ is same for the N_t streams. Additionally, each receive subarray also employs a steering vector $\mathbf{a}_r(\theta_r, \phi_r) \in \mathbb{C}^{n \times 1}$ to receive beamform to the associated RA on the receiver side. By symmetry, $\mathbf{a}_r(\theta_r, \phi_r) = \mathbf{a}_t(\theta_t, \phi_t)$ and is same across all N_r streams. The effective channels $\hat{\mathbf{T}}_k \in \mathbb{C}^{M \times 1}$ and $\hat{\mathbf{R}}_k \in \mathbb{C}^{M \times 1}$ between each subarray-RA pair in the transmit and receive side for stream k are then

$$\left.\begin{aligned} \hat{\mathbf{T}}_k &= \mathbf{T}_k \mathbf{a}_t^*(\theta_t, \phi_t) \quad k \in \mathcal{N}_t \\ \hat{\mathbf{R}}_k &= \mathbf{R}_k \mathbf{a}_r^*(\theta_r, \phi_r) \quad k \in \mathcal{N}_r \end{aligned}\right\} \tag{3}$$

The diagonal complex phase matrices for the RAs on the transmit and receive side are denoted by $\mathbf{\Psi}_t \in \mathbb{C}^{MN_r \times MN_t}$ and $\mathbf{\Psi}_r \in \mathbb{C}^{MN_r \times MN_t}$ respectively. The total phase to be compensated by each RA for stream $k \in \mathcal{N}_t$ can be expressed as the addition of the element wise phase compensations required to compensate for the phase from the associated transmit subarray and the phase required to beamform towards the RAs on the receiver side. Alternatively, this is equivalent to the Hadamard product ($\odot$) between $\mathbf{S}_k$ and $\mathbf{S}_D$ as

$$\mathbf{\Psi}_{t_k} = \text{diag}(\mathbf{S}_k \odot \mathbf{S}_D) \tag{4}$$

The combined effective block diagonal phase matrix for the transmit side RAs is then

$$\mathbf{\Psi}_t = \text{diag}\{\mathbf{\Psi}_{t_1}^*, \ldots, \mathbf{\Psi}_{t_{N_t}}^*\} \tag{5}$$

By the symmetry of our system model, $\mathbf{\Psi}_r = \mathbf{\Psi}_t$. Hence, the effective channel seen at the receiver $\mathbf{H} \in \mathbb{C}^{N_r \times N_t}$ is

$$\mathbf{H} = \mathbf{H}_r \mathbf{\Psi}_r \mathbf{L} \mathbf{\Psi}_t \mathbf{H}_t \tag{6}$$

where, $\mathbf{H}_t \in \mathbb{C}^{(MN_t) \times N_t}$ is a block diagonal matrix capturing the combined LoS channel between the N_t transmit subarrays and the corresponding RAs and $\mathbf{H}_r \in \mathbb{C}^{N_r \times (MN_r)}$ is a block diagonal matrix capturing the combined LoS channel between the N_r receive subarrays and the corresponding RAs. $\mathbf{H}_t$ and $\mathbf{H}_r$ are

$$\left.\begin{aligned} \mathbf{H}_t &= \text{diag}\{\hat{\mathbf{T}}_1, \ldots, \hat{\mathbf{T}}_{N_t}\} \\ \mathbf{H}_r &= \text{diag}\{\hat{\mathbf{R}}_1, \ldots, \hat{\mathbf{R}}_{N_r}\} \end{aligned}\right\} \tag{7}$$

where $\hat{\mathbf{T}}_k$ and $\hat{\mathbf{R}}_k$ are defined as described in (3). Finally, the received signal vector $\mathbf{y} \in \mathbb{C}^{N_r \times 1}$ is

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{w} \tag{8}$$

where, $\mathbf{x} \in \mathbb{C}^{N_t \times 1}$ are the N_t unique QPSK streams sent from the transmit subarrays and $\mathbf{w} \sim \mathcal{CN}(0, \sigma^2 \mathbf{I}_{N_r})$ denotes the iid additive complex Gaussian noise terms. Linear minimum mean square equalizer (LMMSE) is applied at the receiver to demodulate the received signal. While (6) describes the generalized form for the effective channel induced by the RA aided LoS MIMO system, specific guidelines related to system design parameters such as the required RA size M and their optimal distance from the transceivers are necessary to effectively construct the RA phase compensation matrices and maximize the overall channel gain. To this end, we detail a quantifiable benchmark for our system in Section III.

III. Geometry and Beam Design

In this section, we develop design guidelines for the required system geometry (RA sizes and distance of RAs from transceivers) and phase compensation strategies.

A. Link Geometry Design

We seek to design the link geometry so that the path losses for our three-hop system should not be (too much) worse than that of a one-hop SISO system described by the journey of a single QPSK stream from a transmit subarray to reach a receive subarray through a direct LoS link. To this end, we compare an upper bound on the per-stream SNR for our system to that of the SISO system.

The upper bound is computed under the following ideal (unrealizable in the near field even with perfect CSI): signals from the elements of a transmit subarray to its designated RA add up coherently at each RA element, and the signals from a receive RA to its designated subarray add up coherently at each subarray element. In addition, we assume ideal far-field beamforming between RAs (which is realizable under perfect CSI). Under these assumptions, the amplitude seen at a single RA element l on the transmit side is given by

$$A_T = n_t |\alpha_T| \tag{9}$$

where

$$\alpha_T = \sqrt{G_T G_L}\big(\lambda/(4\pi d_{T_{li}})\big) e^{-j2\pi d_{T_{li}}/\lambda} \tag{10}$$

and where, G_T denotes the antenna element gain per transmit element, G_L the gain per reflecting element and $d_{T_{li}}$ the distance form transmit element i to the corresponding reflecting element l of the associated RA. The amplitude at each RA element on the receive side is given by

$$A_L = M |\alpha_L| A_T \tag{11}$$

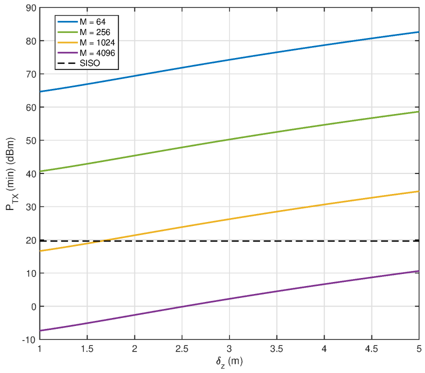

Fig. 3. P_{Tx} dBm for different RA sizes M and depth parameter δ_z with Tx subarray size of 8×8

where,

$$\alpha_L = \sqrt{G_L G_L}\big(\lambda/(4\pi d_{L_{ml}})\big)e^{-j2\pi d_{L_{ml}}/\lambda} \tag{12}$$

and where, $d_{L_{ml}}$ is the distance between RA element l on the transmit side and RA element m on the receiver side. It follows that the amplitude at each subarray element in the receiver is then

$$A_R = M|\alpha_R|A_L \tag{13}$$

where $\alpha_R = \alpha_T$ by the symmetry of our system model. The overall amplitude per stream is therefore

$$A = N_r n_r A_R. \tag{14}$$

For $n_t = n_r = n$, the noise variance per stream is then $N_r n\sigma^2$. The effective SNR per stream for the RA aided LoS MIMO system under the ideal combining assumption is

$$\text{SNR} = \left(N_r n^3 M^4 |\alpha_T|^4 |\alpha_L|^2\right)/\sigma^2 \tag{15}$$

We now compare this unrealizable ideal against the SNR for the ideal SISO system, which is given by

$$\text{SNR}_{\text{SISO}} = \left(n^3 |\alpha_{\text{SISO}}|^2\right)/\sigma^2 \tag{16}$$

where $\alpha_{\text{SISO}} = \alpha_L$ by symmetry. Considering $G_T = G_L$ and $d_{T_{li}} \approx \delta_z$, the effective channel gain per stream obtained by utilizing RAs is found by comparing (15) and (16) as

$$G = N_r \left(M G_T \lambda/(4\pi\delta_z)\right)^4 \tag{17}$$

Thus, the gain G increases with the fourth power of the number M of RA elements (due to the ability of RAs to both gather and direct energy), while decreasing with the fourth power of the normalized depth $\frac{\delta_z}{\lambda}$ (due to the two hops corresponding to the links between the RAs and their designated subarrays at each end). Thus, M must scale linearly with $\frac{\delta_z}{\lambda}$ in order not to incur significant loss in link budget relative to the SISO benchmark. In order to obtain numerical values, we carry out link budget analysis with δ_z values between 1 m and 5 m with the following system parameters:

- antenna element gain covering a hemisphere is 3 dBi
- the $n_t = 64$ transmit subarray gives 18 dB transmit beamforming gain, plus 18 dB power pooling gain and the $n_r = 64$ receive subarray gives 18 dB receive beamforming gain

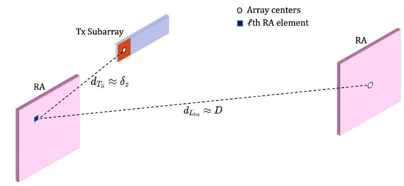

Fig. 4. Geometric representation of phase shifts required by RA element

- the RA at the transmitter and the receiver, each provide a beamforming gain of $10\log_{10}(M)$ respectively
- the noise figure of each RF chain is 7 dB
- thermal noise power for 800 MHz is about -85 dBm and SNR is 10.2 dB for QPSK modulation scheme
- link margin of 10 dB, a total insertion loss of 5 dB and a rain attenuation of 7 dB/Km which corresponds to a rain rate of 50mm/hr [11] (moderate rainfall) is considered

The required receiver sensitivity in this setting is -62 dBm. The link budget analysis concretely illustrates the trade-off between M and δ_z in (17): Figure 3 plots the minimum transmit power per element required for varying RA sizes and depth parameter values δ_z compared to the benchmark SISO system. The SISO system requires a transmit power per element of 20 dBm, which is realizable in low-cost CMOS processes for a 28 GHz carrier. In order to realize our proposed system with similar transmit powers, we see from Figure 3 that the minimal configuration (smaller RA size, largest depth) needed to prevent a substantial decline in link budget compared to the SISO link is $M = 1024$ elements at $\delta_z = 1$ m. As we discuss next, such combinations of large RA size and relative small depth place us in a near-field regime which requires that the RAs perform "beam focusing."

B. Beam Design

Having extracted link geometry design guidelines from the SNR upper bound, we now explore simple strategies for beam design that approach this upper bound, as follows:

Subarray beam: Each transmit subarray implements a linear phase profile under a far field assumption, using a UPA steering vector $\mathbf{a}_t(\theta_t, \phi_t)$ designed to steer the beam to the center of its associated RA. This is easy to implement based on geometric information regarding the relative distance and orientation of subarrays and RAs, but incurs some loss relative to employing the dominant eigenmode of the $n_t \times M$ channel matrix between the subarray and the RA. Each receive subarray implements an analogous center-to-center beamformer pointing towards its designated RA.

RA beam: At the RA, the phase profile of the complex phase compensation vector $\mathbf{S}_k$ for stream k depends on δ_z. For the considered transceiver aperture size d_s, it can be verified that we operate in the *near-field* [12] of the transmit subarray for values of δ_z between 1m and 5 m. This implies that the phase profile to be compensated by the RA from the transmit subarray is quadratic. It becomes necessary therefore to accurately account for the curvature of the wavefront impinging on the RA from the transmitter through *beam-focusing* [5]. Then, the complex quadratic phase to be compensated by reflecting

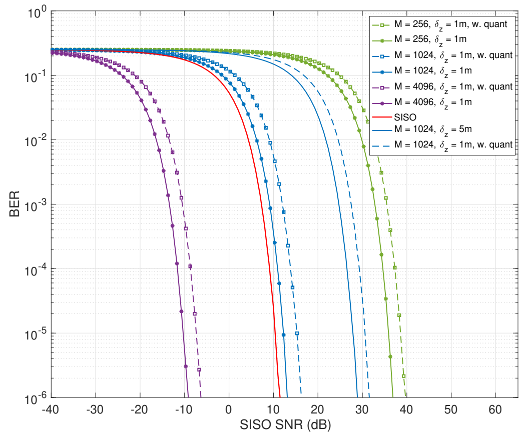

Fig. 5. Bit error rate versus SISO SNR for various RA sizes and depth δ_z

element $l \in \mathcal{M}$ of the RA corresponding to stream k is defined by

$$S_{k_l} = \sum_{i=0}^{n_t-1} e^{-j2\pi d_{Tk_{il}}/\lambda} \mathbf{a}_{t_i}^*(\theta_t, \phi_t) \tag{18}$$

In addition, the phase compensation required at each transmit RA to steer towards the receiver RAs is $\mathbf{S}_D(\theta_D, \phi_D)$, a linear phase profile synthesizing a far-field beam determined by the elevation and azimuth steering angles θ_D and ϕ_D towards the receive plane. Due to the long range considered in our system, the same steering vector is used by all RAs. The effective two-fold phase compensation required by each RA is therefore the sum of two phase profiles: (i) the quadratic phase from the transmit subarray and (ii) the linear phase compensation to beamform towards the RAs on the receiver side and is constructed as shown in (4). A geometric representation of this required phase compensation is presented in Fig. 4 as a function of the relative distances δ_z and D. Phase compensation strategies for RAs at the receive side are entirely analogous.

IV. Results and Discussion

We now present simulation results for the beam design in Section III-B, comparing the bit error rate (BER) averaged over streams against the benchmark SISO system and the predictions from the SNR upper bound in Section III-A. We also explore the performance impact of coarse 2-bit quantization of the phases in the RA elements. Figure 5 shows BER curves for varying RA sizes with $M = 256, 1024$ and 4096 reflecting elements at $\delta_z = 5$ and 1 m. We average over 10^3 realizations, each with random values for the horizontal and vertical receive plane displacement Δ_x and Δ_y (see Figure 1) uniformly drawn from $[-2, 2]$ m. A total insertion loss of 5 dB is incorporated into the channel model used for simulations. Some key design insights are as follows: 1) $M = 1024$ elements at $\delta_z = 1$ m is close to the SISO benchmark as predicted by the analysis in Section III-A, and represents an attractive option for synthesizing the required DoF. 2) For fixed δ_z, the distance between curves is approximately predicted by the M^4 scaling of the gain in (17): $M = 1024$ is 24 dB worse, while $M = 256$ is 48 dB worse than $M = 4096$. 3) For fixed M, we expect the performance at $\delta_z = 1$ m to be 28 dB better than at $\delta_z = 5$ m from the $1/\delta_z^4$ scaling in (17), but the BER curves for $M = 1024$ show, for example, only 18 dB improvement in performance as we decrease δ_z from 5 m to 1 m. An interesting open question is to what extent we can improve upon our center-to-center beam design to more closely approach the SNR upper bound (15) for small δ_z.

V. Conclusion

Our proposed approach for creating spatial DoF highlights the promise of RAs in synthesizing novel wireless environments that would be precluded by conventional designs, while bringing out key challenges to be addressed. Our results show that link budget considerations are paramount when utilizing RAs at long ranges, dictating choice of system parameters (RA sizes and distances from transceivers) and the associated signal processing strategies. In particular, RAs need to implement a combination of short-range beam focusing and long-range beamforming in our proposed system. An important topic for future work, therefore, is the development of efficient phase adaptation/channel estimation algorithms making realistic assumptions on geometric priors and available feedback.

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