Joint Microstrip Selection and Beamforming Design for MmWave Systems with Dynamic Metasurface Antennas

Wei Huang, Haiyang Zhang, Nir Shlezinger, and Yonina C. Eldar

Abstract—Dynamic metasurface antennas (DMAs) provide a new paradigm to realize large-scale antenna arrays for future wireless systems. In this paper, we study the downlink millimeter wave (mmWave) DMA systems with limited number of radio frequency (RF) chains. By using the specific DMA structure, an equivalent mmWave channel model with hybrid beamforming is first explicitly characterized. Based on that, we propose an effective joint microstrip selection and beamforming scheme to accommodate for the limited number of RF chains. A low-complexity digital beamforming solution with channel gain-based microstrip selection is developed, while the analog beamformer is obtained via a coordinate ascent method. The proposed scheme is numerically shown to approach the performance of DMAs without RF chain reduction, verifying the effectiveness of the proposed schemes.

Index terms—Dynamic metasurface antennas, millimeter wave.

I. INTRODUCTION

In order to increase the capacity of wireless communication systems, millimeter wave (mmWave) bands ranging from 30GHz to 300GHz, are regarded as a promising candidate. The small wavelength of mmWave signals allows a large number of antenna elements to be packed in a small area, facilitating multiple-input multiple-output (MIMO) processing with very large arrays. However, realizing large antenna arrays for mmWave communication systems in practice can be challenging. A key difficulty stems from the fact that radio frequency (RF) chains in mmWave are costly in terms of hardware implementation, signal processing complexity, and energy consumption. To overcome this issue, various cost-aware hybrid architectures with limited number of RF chains have been proposed [1]–[3]. However, typical hybrid architectures come at the cost of additional analog circuitry, typically comprised of multiple phase shifters, which can lead to relatively high energy consumption [4].

An alternative large-scale MIMO technology utilizes dynamic metasurface antennas (DMAs) [5]. DMAs consist of multiple waveguides (microstrips), each embedded with many metamaterial antenna elements, which inherently provides analog beamforming capabilities with lower power consumption and cost compared with typical phased array antennas [6]. Moreover, DMAs are typically utilized with sub-wavelength element spacing, allowing to pack a larger number of elements in a given antenna area compared to conventional phased array hybrid antennas. This ability has been exploited to enhance communication capacity and energy efficiency [7], [8].

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II. SYSTEM MODEL AND PROBLEM FORMULATION

A. Dynamic Metasurface Antennas

Here, we give a brief review of DMAs. DMAs are metasurface-based antennas comprised of multiple microstrips, which are one-dimensional arrays of metamaterial elements placed on a waveguide cavity [5]. In such architectures, each RF chain is connected with the port located at the edge of the microstrip. Thus, during transmission, the signals in the microstrip propagating from that port undergo a different path for each element, which results in different propagation delay depending on their location. Define \( \alpha_n \) as the wavenumber along microstrip \( n \), and let \( \rho_{n,m} \) denote the relative location of the...
Fig. 1. Downlink MISO system with DMAs.

\textit{mth element of the \textit{n}th microstrip, which is usually proportional to the distance between the port of microstrip \textit{n} and the \textit{mth} element.}

The element-dependent propagation effect is formulated as

\[ f_{n,m} = e^{-\sigma_{n,m} \left( \beta_{m} + j \omega_{m} \right)}, \quad \forall n, m, \quad (1) \]

where \( \beta_{m} \) is the waveguide attenuation coefficient of element \( m \).

Each metamaterial element acts as a resonant circuit whose frequency response for narrowband signaling is approximated as the Lorentzian-constrained phase weights model [13, 16], given by

\[ g_{n,m} \in \mathbb{G} \triangleq \left\{ \frac{j + e^{j \phi_{n,m}}}{2} \right\}, \quad \forall n, m, \quad (2) \]

Here, \( \phi_{n,m} \) denotes the phase shift of the \textit{mth} element in the \textit{n}th microstrip. From (1) and (2), we observe that the equivalent DMAs frequency response for each radiating element is a product of the propagation response (1), which is dictated by the placing of the elements, and the tunable Lorentzian phase weight (2).

**B. Downlink MmWave Systems with DMAs**

We consider a mmWave downlink multiple-input single output (MISO) system, as illustrated in Fig. 1. Here, a base station (BS) equipped with a DMA having \( U = N \times M \) radiating metamaterial elements serves a single-antenna user, where \( N \) and \( M \) denote the number of microstrips and radiating elements in each microstrip, respectively. For cost-effective implementations, the BS is equipped with \( N_{RF} \) (\( N_{RF} \leq N \)) RF chains, which are connected with the DMA array via a switch network. We adopt the widely utilized geometric block-fading channel model, where the channel from the BS to user, denoted by \( h \in \mathbb{C}^{M \times 1} \), is given by

\[ h = [h_{1}^{H}, \ldots, h_{N}^{H}] \in \mathbb{C}^{M \times 1}, \quad (3) \]

In (3), \( h_{n} \in \mathbb{C}^{M \times 1} \) denotes the channel from the \textit{n}th microstrip to user, and \( L \) is the number of scatterers (paths) from the BS to the user, which is generally much smaller than the number of antennas because of the severe path loss in mmWave band; \( \eta \) denotes the complex-value gain of the \textit{l}th path; \( \theta_{l} \in [0, \pi] \) represents the angle of deviation (AoD) of the \textit{l}th path between the BS and user; while \( a(\theta_{l}) \) is the antenna array response vector corresponding to AoD \( \theta_{l} \) between BS and user, which is given by

\[ a(\theta_{l}) = \begin{bmatrix} 1, \ldots, e^{-j 2 \pi (N-1) d_{s} \sin(\theta_{l})} \end{bmatrix}^{T}, \quad \forall n, m, \quad (4) \]

where \( \lambda \) is the signal wavelength, \( d_{s} \) is the distance between adjacent elements, and \( d_{s} \) denotes the distance between the microstrips. \( d_{s} \) can be interpreted as the spatial frequency of the \textit{n}th element corresponding to AoD \( \theta_{l} \).

Since the outgoing signals propagate inside the microstrips before being radiated, the equivalent MISO channel can be characterized as

\[ h = [h_{1}^{H}, \ldots, h_{N}^{H}]^{H} = h \odot f = \left( \sum_{l=1}^{L} \eta_{l} a(\theta_{l}) \right) \odot f, \quad (5) \]

\[ x_n = w_n s, \quad \forall n, \quad (7) \]

where \( x_{n} \) is the input signal for microstrip \( n \), \( w_{n} \) is the corresponding digital beamforming, and \( s \) is the information-bearing symbol with normalized power. Then, the baseband signals \( \{x_{n}\} \) are fed to the corresponding microstrips via the switch network. Thus, the signal emitting by the \textit{n}th element in microstrip \( n \) is \( f_{n,m} z_{n,m} \), where

\[ z_{n,m} = g_{n,m} \xi_{n} x_{n} = g_{n,m} \xi_{n} w_{n} s, \quad (8) \]

In (8), \( \xi_{n} \in \{0, 1\} \) denotes whether the \textit{n}th microstrip is activated or not. Define the \( N \times 1 \) vectors \( x = [x_{1}, \ldots, x_{N}]^{H} \), \( w = [w_{1}, \ldots, w_{N}]^{H} \), \( g_{n,m} = [g_{n,1}, \ldots, g_{n,M}]^{H} \), representing the configurable weights of the microstrip \( n \). After propagating via the wireless channel, the received signal \( y \) is expressed as

\[ y = h^{H} z + o = h^{H} G w + o, \quad (10) \]

where \( o \) is additive white Gaussian noise with variance \( \sigma^{2} \). Then, the achievable rate is given by

\[ \max \frac{1}{\sigma^{2}} \left\| h^{H} G w \right\|^{2} \quad \text{s.t.} \quad \left\| w \right\|^{2} \leq P, \quad \left\| w \right\| \leq N_{RF}. \quad (11) \]

**C. Problem Formulation**

For a mmWave channel with severe path loss, a large array can be exploited to compensate for the high path loss. Even for a single antenna receiver, the number of transmit elements in one microstrip may not be sufficient, as the emitting energy of back-end antennas may become weaken when the signals are propagating inside the microstrip. It is thus of importance to properly utilize the available RF chains to achieve reliable communications.

Since the number of RF chains at the BS is generally smaller than that of the microstrips in mmWave systems, i.e., \( N_{RF} \leq N \), the RF chains have to selectively connect with some microstrips via the switch network. As such, the number of active microstrips may not exceed \( N_{RF} \). Moreover, as the achievable rate monotonically grows with the SNR, our objective is to maximize the SNR under the RF chains constraint, by joint microstrip selection and beamforming optimization. Therefore, the optimization problem is stated as

\[ \max_{w, \{z_{n,m}\} \in \mathbb{G}} \frac{1}{\sigma^{2}} \left\| h^{H} G w \right\|^{2} \quad \text{s.t.} \quad \left\| w \right\|^{2} \leq P, \quad \left\| w \right\| \leq N_{RF}. \quad (11) \]
feasible set as a circle on the complex plane. To address this, we define the variables $(g_{n,m}) \in \mathcal{G}$. To solve the non-convex problem, we develop an iterative algorithm in the following section.

### III. Joint Microstrip and Beamforming Design

In this section, we develop an algorithm to optimize $w$ and $g_{n,m}$. We adopt an alternating approach, where we maximize the SNR by sequentially fixing one variable and updating the other.

**Optimizing $w$:** For a given $G$, the goal is to jointly select the $N_{RF}$ out of $N$ microstrips and design the corresponding beamforming vector associated with the selected microstrips to maximize the SNR. The resultant subproblem is given by

$$
P_1: \max_w \frac{1}{\sigma^2} ||h^H G w||^2 \tag{12a}$$

s.t. \[ ||w||_2^2 \leq P \] \[ ||w||_0 \leq N_{RF} \] \[ (12b) \] \[ (12c) \]

Note that in $P_1$, the objective $(12a)$ and power constraint $(12b)$ are convex. The main difficulty lies in the non-convex and non-differentiable cardinality constraint $(12c)$, which restricts the number of the selected microstrips to be no more than $N_{RF}$. Despite this, $P_1$ admits a closed-form solution, as stated in the following:

**Proposition 1.** Let $\mathcal{I} \triangleq \{i_1, \ldots, i_{N_{RF}}\}$ denote the indices of the $N_{RF}$ largest entries in the set $\{|h_{n,m}^H g_j|^2\}_{n,m=1}^N$. Accordingly, define $\mathbf{h} \triangleq [h_{i_1}, \ldots, h_{i_{N_{RF}}}^H] \mathbf{G} \triangleq \text{diag}([g_{i_1}, \ldots, g_{i_{N_{RF}}}^H])$, and

$$w = \sqrt{P} \frac{G^H \mathbf{h}}{||G^H \mathbf{h}||_2} \tag{13}$$

Then, it holds that $P_1$ is solved by setting

$$w_n = \begin{cases} |w_{i_j}| & \exists i_j \in \mathcal{I} \text{ such that } n = i_j \\ 0 & \text{otherwise} \end{cases} \tag{14}$$

**Proof:** Due to the page limitations, the detailed proof can be found in [17].

**Optimizing $G$:** For a given beamforming vector $w$ set via Proposition 1, the resulting SNR is given by

$$\text{SNR} = \frac{1}{\sigma^2} ||h^H G w||^2 = \frac{P}{\sigma^2} ||h^H G||^2 = \frac{P}{\sigma^2} \sum_{n=1}^{N_{RF}} \sum_{m=1}^M h_{n,m}^* g_{n,m}^2, \tag{15}$$

where $h_{n,m}$ denotes the $m$th element of vector $h_{n,m}$. Then, the configurable weights optimization subproblem is posed as

$$P_2: \max_{\{g_{n,m}\}} \frac{P}{\sigma^2} \sum_{n=1}^{N_{RF}} \sum_{m=1}^M h_{n,m}^* g_{n,m}^2 \tag{16}$$

Note that in $(16)$, the Lorentzian constraint in $P_2$ characterizes the feasible set as a circle on the complex plane $|g - \frac{1}{2} e^{j\frac{\pi}{2}}| = \frac{1}{2}$, with the circle center at $(0, e^{j\frac{\pi}{2}})$ and radius equals to $\frac{1}{2}$. This non-convex constraint makes the subproblem $P_2$ difficult to be solved directly. To address this, we define the variables $b_{n,m}, \forall m,n$, which are related to the DMA weights via the affine mapping

$$b_{n,m} = 2g_{n,m} - e^{j\frac{\pi}{2}}, \forall m,n \tag{17}$$

The variable $b_{n,m}, \forall m,n$ lies on the unit circle of complex plane with the circle center at the origin of coordinates. With $(17)$, the Lorentzian constraint in $P_2$ is converted to the unit-amplitude constraint. Thus, subproblem $P_2$ is equivalently transformed into

$$P_{21}: \max_{\{b_{n,m}\}} \frac{P}{4\sigma^2} \sum_{n=1}^{N_{RF}} \sum_{m=1}^M (h_{n,m}^* b_{n,m} + h_{n,m}^* z_{n,m})^2 \tag{18}$$

s.t. \[ |b_{n,m}|^2 = 1, \forall m,n \].

While $P_{21}$ is still challenging, it can be tackled via a coordinate ascent-based heuristic algorithm to find a local optimal solution, based on the following proposition.

**Proposition 2.** Any local optimal phase of $(18)$ satisfies

$$\angle b_{n,m} = -\angle (\bar{h}_{n,m}^* \bar{z}_{n,m}) \forall n,m, \tag{19}$$

where $\bar{h}_{n,m} = \sum_{m'\neq m} h_{n,m'}^* b_{n,m'} + \sum_{m=1}^M h_{n,m} e^{j\pi}$. \[ \blacksquare \]

**Proof:** The proof is given in Appendix A.

Using Proposition 2, we can find the local optimal solution of subproblem $P_{21}$ by alternatingly updating each element $b_{n,m}, n \in \mathcal{I}$, with all other elements fixed. Since the convex objective function in $P_{21}$ always increases in each iteration, the iterative algorithm is guaranteed to converge eventually. Then, based on the affine relationship between $g_{n,m}$ and $b_{n,m}$ in $(17)$, we configure the weights of the elements in the active microstrips $(g_{n,m})_{n \in \mathcal{I}}$.

**Algorithm Summary:** The joint optimization of the DMAs weights along with the microstrip selection and beamforming alternates between the individual settings in Propositions 1-2. In each iteration, we take the solutions $w$ and $(g_{n,m})$ into those locations corresponding to the selected microstrips of $w$ and $G$, and set the entries that are non-selected microstrips zeros, respectively. The proposed algorithm to solve problem $(11)$ is summarized as Algorithm 1.

**Algorithm 1** Solving problem $(11)$ w.r.t. $w$ and $(g_{n,m})$

1. **Initialize:** set initial $G$ and threshold $\epsilon > 0$
2. **repeat**
3. **Get** $\mathcal{I}$ and $w$ based on $G$ via Proposition 1;
4. **Set** $(g_{n,m})_{n \in \mathcal{I}}$ via $(19)$ and $(17)$
5. **until** SNR increase is smaller than $\epsilon$
6. **Output:** Beampattern $w$; active set $\mathcal{I}$, and DMA weights $(g_{n,m})$

**Discussion:** Our study is different from the RF chain selection and beamforming design in conventional hybrid architecture with switch network [18], [19]. On the one hand, the element of DMA follows Lorentzian constraint, instead of the unit-module constraint as in conventional hybrid architecture. On the other hand, the RF chain selection is formulated as an integer programming problem, which does not pass a closed-form solution. The proposed Algorithm 1 jointly designs the digital beamformers, DMA weights, and microstrip selection, via alternating optimization. Each step in Algorithm 1 is based on a simple closed-form computation, and it is particularly designed to be simple to implement such that it can be carried out online. The resulting joint design allows to achieve a rate within a relatively small gap from that of costly fully-digital arrays, as we demonstrate in Section IV.

Our joint design gives rise to multiple potential extensions. The fact that the signal processing capabilities of DMAs can be viewed as a form of hybrid precoding indicates that the proposed Algorithm can be extended to accommodate other forms of hybrid architectures based on, e.g., vector modulators [4] and phase shifter networks [20]. Furthermore, while we focus here on narrowband signalling, for wideband signals, DMAs are known to provide additional degrees of freedom in the form of a flexible frequency selective analog...
processing [7]. Moreover, when operating in rapidly time-varying channels, one may wish to limit the number of alternating iterations to a small number. In such cases, emerging model-based deep learning techniques [21] can leverage data to facilitate rapid processing, see, e.g., [22]. We leave these extensions for future study.

IV. NUMERICAL EVALUATIONS

In this section, we evaluate our proposed design. Unless otherwise stated, we consider a planar array located in the $xz$-plane. All simulated architectures have the same-sized array aperture and the array size is $0.3\times0.3m$. For DMAs, $\lambda/5$ spacing between the elements inside each microstrip is considered and the separation between microstrips could be still $\lambda/2$. That is, $N = 10$ and $M = 30$. While for the fully digital architecture and hybrid architecture with partially connected phase shifter, the element spacing is $\lambda/2$, i.e., $N = M = 10$. The system operates at 28GHz and the mmWave channel has 12 clusters. As in [15], we set $\alpha = 0.6[m^{-1}]$ and $\beta = 827.67[m^{-1}]$ to represent the propagation inside the DMA waveguides, representing a microstrip implemented in Duroid 5880.

Our proposed design is compared with the following schemes:

• **Fully digital beamforming:** Each antenna is connected to a dedicated RF chain and the corresponding transmit beamforming uses the maximum ratio transmit (MRT) strategy [23], which optimizes the SNR objective (11a) for such settings.

• **DMAs architecture without RF chain reduction:** Each microstrip is connected with one RF chain, and the digital and analog beamforming are designed as in [15].

• **DMAs architecture with incompact element spacing:** All elements spacing are set to $\lambda/2$, and the digital and analog beamforming are designed using the approach proposed in Section III.

• **Random microstrip selection and beamforming:** Randomly select the given number of microstrips, and then apply the proposed method in III to design the corresponding digital and analog beamforming.

• **Hybrid architecture with partially connected phase shifter:** The number of connected RF chains is same to the random microstrip selection scheme, and the digital and analog beamforming vectors are derived via the alternate optimization approach proposed in [24].

In Fig. 3, we compare the spectrum efficiency of our proposed Algorithm 1 (channel gain-based microstrip selection and beamforming) with the above-mentioned benchmark schemes. For microstrip reduction schemes, the total number of RF chains is $N_{RF} = 3$. It is observed that the performance proposed in Algorithm 1 is better than that of the hybrid architecture with partially connected phase shifter. The gain is attributed to the sub-wavelength feature of DMAs, allowing to pack a larger number of elements in a given physical antenna area compared to phased array antennas. However, for DMAs with incompact element space, the performance of hybrid architecture with partially connected phase shifter is better than that of DMAs. Moreover, we can see that the achievable spectrum efficiency of our proposed scheme can reduce the number of RF chains by 75% and signal processing complexity without compromising on performance.

V. CONCLUSIONS

In this paper, we studied mmWave MISO communications with DMA and limited number of RF chains. An equivalent mmWave channel model with DMA was characterised. Then, based on this model, a joint optimization problem of DMAs weights, RF chain selection matrix and digital beamforming vector is formulated to maximize the SNR. We proposed an alternating optimization algorithm to solve the non-convex problem. Numerical results show that the proposed scheme can reduce the number of RF chains by 75% and signal processing complexity without compromising on performance.

APPENDIX

A. Proof of Proposition 2

Considering the fact that $|b_{n,m}|^2 = 1$ and write it as $e^{j\theta_{n,m}}$, the objective function of (18) can be expanded as

$$
\sum_{n=1}^{N_{RF}} \left| \sum_{m=1}^{M} (h_{n,m}^* e^{j\theta_{n,m}} + h_{n,m}^* e^{j\theta_{m,n}}) + \sum_{m' \neq m} h_{n,m}^* e^{j\theta_{n,m'}} + \sum_{m=1}^{M} h_{n,m}^* e^{j\theta_{m,n}} + h_{n,m}^* e^{j\theta_{n,m}} \right|^2
$$

$$
= \sum_{n=1}^{N_{RF}} |h_{n,m}^* e^{j\theta_{n,m}} + \hat{h}_{n,m}^*|^2 = \sum_{n=1}^{N_{RF}} |h_{n,m}|^2 + |\hat{h}_{n,m}|^2 + 2\Re\{h_{n,m}^* \hat{h}_{n,m}^* e^{j\theta_{n,m}}\}. 
$$

(A.1)

With all elements of variables $\{b_{n,m}\}$ fixed except $b_{n,m}$, we should maximize $\Re\{h_{n,m}^* \hat{h}_{n,m}^* e^{j\theta_{n,m}}\}$, which is equivalent to minimize the angle between $h_{n,m}^* \hat{h}_{n,m}^*$ and $e^{j\theta_{n,m}}$. Thus, we have

$$
\theta_{n,m}^* = -\angle(h_{n,m}^* \hat{h}_{n,m}^*) \quad \forall n, m,
$$

which completes the proof.
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