HBF-PDVG: Hybrid Beamforming and User Selection for UL MU-MIMO mmWave Systems

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***Abstract*—In this paper, a novel hybrid beamforming (HBF) design utilizing a pre-defined virtual grouping (PDVG) algorithm is proposed for the uplink (UL) multi-user (MU) multiple-input multiple-output (MIMO) millimeter wave (mmWave) systems. The computational complexity of achieving traditional MU-MIMO beamforming (BF) methods in the mmWave systems becomes significant due to the hardware constraints. Thus, the proposed HBF-PDVG algorithm aims to reduce the feedback overhead by employing a three-phase algorithm. The essential idea is to perform the user selection independently from the HBF design by utilizing the PDVG algorithm after completing the analog BF training and then mitigating the remaining MU interference among the selected users using digital BF. Simulation results demonstrate the significant sum-rate benefits of the HBF-PDVG algorithm compared with zero-forcing (ZF) and random selection algorithm methods.**

I. INTRODUCTION

Recently, Internet traffic has changed from very light website browsing and file transfer to a very rich amount of content uploaded by the users because of the recent emergence of applications, such as ultra-high definition video wireless streaming. Millimeter wave (mmWave) system is considered one of the key enabling technologies for future mobile networks to meet the considerable growth of wireless traffic because of the large bandwidths in the mmWave spectrum [1]–[6]. However, directional antenna using large antenna arrays and beamforming (BF) techniques are essential in the mmWave systems because of the propagation characteristics of the mmWave. In order to attain high quality communication links, massive multiple-input multiple-output (MIMO) systems can be packed in a small physical dimension due to the significantly short wavelength of the mmWave systems. The current IEEE 802.11ad standard, for instance, employs an analog BF in the radio frequency (RF) domain by using antenna phase shifters to modify the phase of the signal transmitted at each antenna.

However, the IEEE 802.11ad is limited to single-stream single-user wireless transmission. In order to enhance the throughput for mmWave bands, the IEEE 802.11 task group ay has been launched to develop modifications to the legacy IEEE 802.11ad standard where it is aimed to support the multi-user (MU) MIMO. MU-MIMO allows a set of links to be transmitted simultaneously to enhance the network throughput by utilizing a large number of antennas. In addition, uplink (UL) MU-MIMO is considered one of the future solutions for the IEEE 802.11ay standard to improve the system efficiency and meet the requirements of the current and upcoming dense deployment usage scenarios.

In order to implement MU-MIMO, BF techniques need to be considered. Employing full-digital BF using a baseband precoder, that has been widely applied in the conventional lower frequency systems, poses challenges in the mmWave systems due to the significant power consumption of the analog-to-digital converters (ADCs) and the large overhead of collecting the complete channel state information (CSI) [6], [7]. Digital/analog hybrid BF (HBF) is proposed to reduce the complexity of the hardware constraints which is a trade-off between performance and complexity. HBF is proposed for both the UL and downlink (DL) scenarios while few UL proposals exist in the literature.

In the UL MU-MIMO, users need to be separated and selected by the access point (AP) to avoid the MU interference. One of the solutions is to coordinate the user/station (STA) selection by the AP in order to have a better spatial characteristic among all the selected STAs [8]. User selection in MU-MIMO was proposed for the lower frequency systems to achieve higher performance where the AP can schedule the most spatially effective simultaneous users [9], [10]. For the mmWave bands, several HBF techniques have been proposed recently by considering in particular the UL scenario in the MU-MIMO mmWave systems [11]–[14]. UL MU-MIMO mmWave system is proposed in [11] for joint HBF design where an orthogonal matching pursuit algorithm is employed for the analog precoder while a mean square error measure is used for the digital BF. In [12], an UL of large-scale MU-MIMO is considered for the mmWave systems. HBF joint optimization problem is addressed to maximize the sum-rate by considering the transmit power, analog phase shifter and receive antenna selection matrix. In [13], a near maximum likelihood detector is proposed to deal with the challenge of the UL MU massive MIMO systems by considering one-bit ADCs where a two-stages are considered to solve the problem. In [14], a HBF for the UL MU-MIMO scenario is proposed where the Gram-Schmidt method and minimum mean square error BF are utilized in the analog and digital stages respectively. However, the digital processing and the complexity of searching the joint optimal solution using the HBF remain challenging in practical mmWave systems.

In this paper, we propose a practical HBF-pre-defined virtual grouping (PDVG) algorithm for the UL MU-MIMO mmWave systems. The complexity of the joint optimization in the HBF mmWave systems grows exponentially with the number of antenna elements and number of STAs in the system [7]. Thus,

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it requires significant BF training and large feedback overhead. The proposed HBF algorithm in this paper, however, implements a simpler design structure by separating the user selection method from the joint optimal HBF design where a three-phase algorithm is utilized in order to reduce the complexity and feedback overhead. The user selection method used in this proposal is performed by exploiting the PDVG algorithm which is a continuation of our previous works [1], [2]. This PDVG algorithm was proposed for the IEEE 802.11 working task group ay, where an orthogonality criterion based on the analog BF information is utilized as a selection metric instead of directly collecting CSI from all STAs. During the first phase of the HBF-PDVG algorithm, the analog beam steering between the AP and each associated STA can be resolved while STAs can be selected independently for simultaneous transmission by the PDVG algorithm during the second phase. Then, during the third phase, the interference resulting from the simultaneous transmissions can be managed by exploiting a digital BF where the method of maximizing the signal-to-interference-plus-noise ratio (SINR) is utilized instead of using the zero-forcing (ZF) method. Performance evaluation and simulation results show that the proposed HBF-PDVG algorithm delivers higher sum-rates compared with existing zero-forcing (ZF) and random selection algorithm methods.

The remainder of this paper is organized as follows: In Section II, the system model is described. In Section III, the details of the proposed three-phases HBF-PDVG algorithm is described. The performance evaluation of the proposed solution is provided in Section IV. Finally, the conclusions are drawn in Section V.

II. System Model

We consider a mmWave system architecture where we use, in what follows, the IEEE 802.11ay mmWave to describe the UL MU-MIMO BF problem, which is also applicable and has the backward compatibility to the legacy IEEE 802.11ad. In this scenario, we consider one AP and K STAs where the AP and STAs are equipped with steerable directional antenna and have the ability of enhanced directional multi-gigabit (EDMG) antenna as defined in the IEEE 802.11ay.

A. IEEE 802.11ay BF Training

The directional MAC access time of the IEEE 802.11ay is formulated into beacon intervals (BIs) as shown in Fig. 1. Beacon transmission interval (BTI), association BF training (A-BFT), announcement transmission interval (ATI) and data transfer interval (DTI) are the access periods of each BI. The BF training for the IEEE 802.11ay is first achieved in the sector level sweep (SLS) and then can be completed optionally in the beam refinement protocol (BRP) during the DTI. This BF training in the SLS is essential to realize appropriate link budget by using analog BF between a transmitter and receiver initially. Then during the A-BFT, the best sector identification (SID) and the signal-to-noise ratio (SNR) are reported for each STA. In the data transfer interval (DTI) period, frame transmissions are exchanged by allocating the service periods (SPs) in different time slots (t) and the length of the SPs time slots is assumed to be equal.

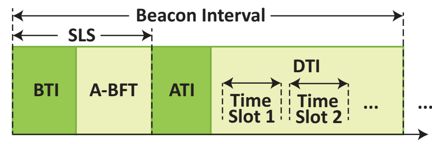

Fig. 1. IEEE 802.11ay directional MAC and BF protocol.

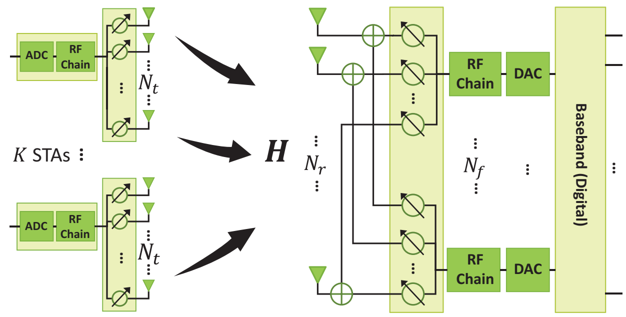

Fig. 2. HBF architecture.

B. HBF Model

As shown in the Fig. 2, an UL MU-MIMO mmWave system is considered in this paper where K STAs are served simultaneously by an AP. N_r and N_t represent the numbers of equipped antennas at the AP and each STA respectively. In our HBF-PDVG algorithm, at the AP, $N_f \geq K$ RF chains are equipped to support multiple STAs simultaneously. For ease of representation, we assume that only K RF chains are adopted out of N_f available RF chains at the AP to serve K STAs, i.e., $N_f = K$. Specifically, the analog BF vectors at i-th STA and i-th RF chain at AP are represented by $\mathbf{u}_i \in \mathbb{C}^{N_t \times 1}$ and $\mathbf{v}_i \in \mathbb{C}^{N_r \times 1}$ respectively where $\mathrm{E}[\|\mathbf{u}_i\|^2] = \mathrm{E}[\|\mathbf{v}_i\|^2] = 1$. Since analog BF is implemented with phase shifters due to the hardware limitation, $\mathbf{v}_i$ is an element of a set of feasible transmitter analog BF vectors $\mathcal{F}_t$ with constant-magnitude entries. Similarly, $\mathbf{u}_i$ is also an element of the set of feasible receiver analog BF vectors $\mathcal{F}_r$. Thus, the discrete time transmitted signal at i-th STA is

$$\mathbf{s}_i = \sqrt{P}\mathbf{u}_i x_i, \tag{1}$$

where x_i denotes the transmitted symbol from the i-th STA which satisfies $\mathrm{E}[x_i] = 0$ and $\mathrm{E}[|x_i|^2] = 1$. Assuming all STAs employ the same transmitting power which is denoted by P. For simplicity, a block-fading channel is assumed in this paper and, thus, the received signal at AP is given by

$$\mathbf{r} = \sum_{i=1}^{K} \sqrt{P}\mathbf{H}_i \mathbf{u}_i x_i + \mathbf{n}, \tag{2}$$

where $\mathbf{H}_i \in \mathbb{C}^{N_r \times N_t}$ denotes the UL channel matrix between i-th STA and AP, and $\mathbf{n} \in \mathbb{C}^{N_r \times 1}$ is the complex Gaussian white noise whose variance is σ^2. For simplicity, the linear antenna array is considered in this paper and, thus, the channel matrix is given by

$$\mathbf{H}_i = \sqrt{\frac{N_t N_r}{L}} \sum_{l=1}^{L} g_l \mathbf{a}_r(\theta_l) \mathbf{a}_t(\phi_l)^*, \tag{3}$$

$$\mathbf{a}_r(\theta_r) = \frac{1}{\sqrt{N_r}} [1, e^{j\frac{2\pi}{\lambda} r \sin(\theta_r)}, ..., e^{j(N_r-1)\frac{2\pi}{\lambda} r \sin(\theta_r)}]^T,$$

$$\mathbf{a}_t(\phi_t) = \frac{1}{\sqrt{N_t}} [1, e^{j\frac{2\pi}{\lambda} r \sin(\phi_t)}, ..., e^{j(N_t-1)\frac{2\pi}{\lambda} r \sin(\phi_t)}]^T,$$

where L is the number of paths. g_l, θ_l and ϕ_l are the channel gain, angle of arrival and angle of departure of the l-th path respectively. λ denotes the mmWave signal wavelength and r represents the distance between antenna elements.

After the procession of analog BF at AP, the processed signal can be written as

$$\begin{aligned} \mathbf{y} &= \mathbf{V}^H \mathbf{r}, \\ &= \mathbf{V}^H \sum_{i=1}^{K} \sqrt{P} \mathbf{H}_i \mathbf{u}_i x_i + \mathbf{V}^H \mathbf{n}, \\ &= \sqrt{P} \mathbf{V}^H \mathbf{H}_i \mathbf{u}_i x_i + \underbrace{\sqrt{P} \sum_{j \neq i}^{K} \mathbf{V}^H \mathbf{H}_j \mathbf{u}_j x_j}_{\text{interference}} + \mathbf{V}^H \mathbf{n}, \end{aligned} \tag{4}$$

where $\mathbf{V} = [\mathbf{v}_1, \mathbf{v}_2, ..., \mathbf{v}_K]$ is the aggregation matrix of all the analog BF vectors at AP. The second item represents the MU interference which is caused by other transmitting STAs in the UL channel.

In order to reduce the MU interference in the UL MU channel, digital BF is adopted after the procession of analog BF. $\mathbf{F} = [\mathbf{f}_1, \mathbf{f}_2, ..., \mathbf{f}_K] \in \mathbb{C}^{K \times K}$ represents the digital BF at AP whose each column has a unit power, i.e., $\|\mathbf{f}_i\|^2 = 1$. For the i-th STA, $\mathbf{f}_i$ is adopted to decode x_i among MU interference. The decoded symbol is given by

$$\hat{x} = \sqrt{P} \mathbf{f}_i^H \mathbf{V}^H \mathbf{H}_i \mathbf{u}_i x_i + \underbrace{\sqrt{P} \sum_{j \neq i}^{K} \mathbf{f}_i^H \mathbf{V}^H \mathbf{H}_j \mathbf{u}_j x_j}_{\text{interference}} + \mathbf{f}_i^H \mathbf{V}^H \mathbf{n}. \tag{5}$$

With the above decoded symbol, the received SINR can be expressed as

$$\gamma_i = \frac{P |\mathbf{f}_i^H \mathbf{V}^H \mathbf{H}_i \mathbf{u}_i|^2}{P |\sum_{j \neq i}^{K} \mathbf{f}_i^H \mathbf{V}^H \mathbf{H}_j \mathbf{u}_j|^2 + |\mathbf{f}_i^H \mathbf{V}^H \mathbf{n}|^2}. \tag{6}$$

Then, the sum-rate can be written as

$$R_{sum} = \sum_{i=1}^{K} \log_2(1 + \gamma_i). \tag{7}$$

Therefore, the design of HBF can be defined as an optimization problem to maximize the sum data rate, which can be given by

$$\begin{aligned} \mathcal{P}1 : \max_{\mathbf{U}, \mathbf{V}, \mathbf{F}} \quad & R_{sum} && \text{(8a)} \\ \text{s.t.} \quad & \mathbf{u}_i \in \mathcal{F}_r, \forall i, && \text{(8b)} \\ & \mathbf{v}_i \in \mathcal{F}_t, \forall i, && \text{(8c)} \\ & \|\mathbf{f}_i\|^2 = 1, \forall i. && \text{(8d)} \end{aligned}$$

Problem $\mathcal{P}1$ is challenging to be solved for the following reasons. First, constraints (9b) and (9c) are finite sets which involve integer constraints. Second, even with fixed analog BF vectors $\mathbf{u}_i$ and $\mathbf{v_i}$, the objective function is still a non-convex function with respect to the digital BF vectors $\mathbf{f}_i$ due to the interference. Problem $\mathcal{P}1$ is a mixed integer programming problem which is a non-convex problem and, thus, realizing the optimal solution can be challenging. To solve this problem, we propose a suboptimal HBF algorithm to design BF efficiently which has a low computation complexity.

III. Three-Phases UL MU-MIMO HBF-PDVG Algorithm

The essential idea of the proposed HBF-PDVG algorithm is to improve the HBF design process by considering an algorithm with three phases as summarized in Algorithm 1. First, the desired beam steering between the AP and each STA is maximized using the analog BF, which is similar to the single user analog BF training in the IEEE 802.11ad standard. During this phase, the SID information is exchanged among associated STAs and AP. Second, the set of STA group is selected for MU transmission by utilizing the PDVG algorithm [1]–[3] where the coverage area of the AP is divided into two virtual groups (VG_{g1} and VG_{g2}) based on the SID information. In this phase, STA selection can be achieved by separating the user selection problem from the joint optimal HBF design, where zero CSI feedback is required. Third, digital BF is employed at the AP in order to manage the MU interference after selecting the set of STAs (VG_{g1} and VG_{g2}) in the second phase.

A. First Phase: Analog BF

The analog BF of the AP and each associated STA are used to maximize the desired beam direction sequentially. This BF training algorithm is similar to the single user analog BF design problem of the IEEE 802.11ad system where BF training is utilized in different time slots at the A-BFT period neither requiring the CSI nor considering the MU interference. This is because the analog BF training of the IEEE 802.11ad system is a bidirectional and iterative training transmissions where the sector sweep (SSW) frames during the SLS stage are utilized. In order to maximize the channel gain for each K STA by considering the analog BF, each STA and an AP select and find a best SID by choosing $\mathbf{u}_i \in \mathcal{F}_r$ and $\mathbf{v}_i \in \mathcal{F}_t$ as shown in the first phase in Algorithm 1. Thus, during the SLS stage of the 802.11ad BF training, the best sector pair between a transmitter and receiver can be selected by transmitting antenna weight vectors (AWVs) training with quasi-omni antenna pattern. Then, the best SID can be specified where the AWV and SNR are reported.

B. Second Phase: PDVG Algorithm

Analog BF and beam steering is essential in the mmWave system in order to establish a high quality communication link where this is not required in the conventional systems that operate in the lower frequencies. Thanks to this analog BF mechanism, STA selection can be achieved by utilizing the information obtained from the analog BF training. In the MU-MIMO systems, the channel quality rises when the angle among concurrent links is almost orthogonal because it is dependent on the angles. Thus, in order to maintain strong channel orthogonality, STA selection of orthogonal STAs is crucial. By deploying the PDVG algorithm, the STAs that have the most spatial separation can be selected.

After achieving the analog BF training and the best SIDs are reported and exchanged between the AP and STAs, the PDVG algorithm starts by dividing the coverage area into virtual sectors (VSs) based on the best SID information as shown in Fig. 3. Two orthogonal groups VG_{g1} and VG_{g2} are defined in order to divide the STAs into orthogonal groups for the sake of achieving MU transmission without the need of finding an optimal set of STAs by collecting the CSI for all of the STAs. Let p be the total number of SID defined in the system, A be the total number of VSs, and $q = \{1, 2, ..., A/2\}$. Then, based on the SID, VG_{g1} and VG_{g2} can be constructed from the following equations, respectively:

$$\left(\frac{A+2pq-2p}{A}\right) \leq \text{SID} \leq \left(\frac{2pq-p}{A}\right), \tag{9}$$

$$\left(\frac{A+2pq-p}{A}\right) \leq \text{SID} \leq \left(\frac{2pq}{A}\right). \tag{10}$$

If $A = 8$, then the SIDs of VG_{g1} and VG_{g2} can be seen in Fig. 3, where $\{VS_1, VS_3, VS_5, VS_7\} \in VG_{g1}$ and $\{VS_2, VS_4, VS_6, VS_8\} \in VG_{g2}$. Even though STAs can be selected for concurrent transmissions using the concept of the divided VSs, a non-spatial orthogonal (NSO) problem exists when there is more than one STA in the same VS [1], [3]. In such a case, collision and MU interference can be significant and in order to solve the NSO problem, the PDVG algorithm selects only one STA for simultaneous transmissions among all other STAs in the considered VS. The reported SNR during the A-BFT is used as a scheduling mechanism to select only one STA that has the highest SNR among other STAs if the NSO problem exists. The PDVG algorithm selects STAs for VG_{g1} and VG_{g2} and assigns the selected STAs at different time slots (t and $t+1$ respectively). Then, the remaining unselected STAs are evaluated again for selection for VG_{g1} and VG_{g2} and assigned at advanced time slots. This process continues until the remaining unselected STAs cannot be selected for VG_{g1} and VG_{g2}. The associated and unselected STAs can be then scheduled sequentially in different time slots.

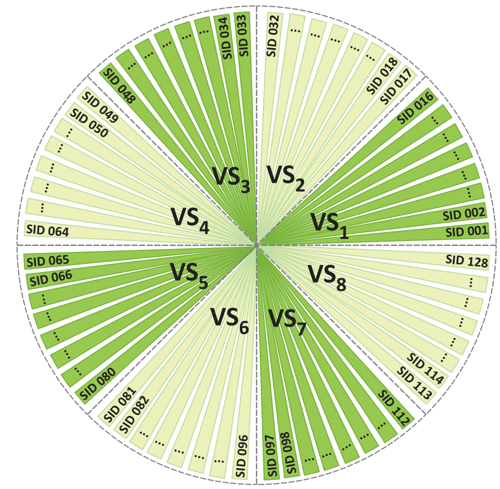

Fig. 3. The SID and the PDVG concept when $A = 8$ and $p = 128$.

Algorithm 1: HBF-PDVG

Input: $\mathcal{F}_r$ AP RF BF codebook
$\mathcal{F}_t$ STA RF BF codebook

First Phase:

The STA and AP have to choose $\mathbf{u}_i \in \mathcal{F}_r$ and $\mathbf{v}_i \in \mathcal{F}_t$ to maximize channel gain, i.e.,

$$\max_{\mathbf{u}_i, \mathbf{v}_i} \quad \|\mathbf{v}_j^H \mathbf{H}_i \mathbf{u}_i\|^2$$
$$\text{s.t.} \quad \mathbf{u}_i \in \mathcal{F}_r, \forall i, \quad \mathbf{v}_i \in \mathcal{F}_t, \forall i.$$

STAs and AP obtain the optimal analog BF, where the best SID is then reported during the A-BFT period

Second Phase:

Repeat

for $SID = 1$ *to* p **do**

if $STA_K \in \left\{\left(\frac{A+2pq-2p}{A}\right) \leq SID \leq \left(\frac{2pq-p}{A}\right)\right\}$ **then**

select STA_K and $\mapsto VG_{g1}$ at time slot t

end

if $STA_K \in \left\{\left(\frac{A+2pq-p}{A}\right) \leq SID \leq \left(\frac{2pq}{A}\right)\right\}$ **then**

select STA_K and $\mapsto VG_{g2}$ at $t+1$

end

if more than one STA in any one of the VSs **then**

select only one STA with the highest SNR

end

Until the remaining unselected STAs cannot be selected **then** allocate unselected STAs sequentially in different t

Third Phase:

For each time slot t, STA i that belongs to VG_{g1} or VG_{g2} estimates the effective channel vector $\bar{\mathbf{h}}_i = \mathbf{V}^H \mathbf{H}_i \mathbf{u}_i$ and feedbacks it to the AP. With all of the effective CSI, AP designs the digital BF for i-th RF chain

$$\mathbf{f}_i = \text{eigenvector}((\hat{\mathbf{H}}_i \hat{\mathbf{H}}_i^H + \frac{\sigma^2}{P}\mathbf{I}_K)^{-1} \bar{\mathbf{h}}_i \bar{\mathbf{h}}_i^H), \quad \text{for } i,$$

where $\hat{\mathbf{H}}_i = [\bar{\mathbf{h}}_1, ..., \bar{\mathbf{h}}_{i-1}, \bar{\mathbf{h}}_{i+1}, ..., \bar{\mathbf{h}}_K]$

AP normalizes $\mathbf{f}_i = \frac{\mathbf{f}_i}{\|\mathbf{f}_i\|}$, for i

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C. Third Phase: Digital BF Design

Although ZF method is widely adopted to design digital BF in traditional DL HBF design, the interference mitigation comes at the cost of energy inefficiency. In the UL channel, the power resources constraint should be considered due to limited on-board battery power. Thus, an energy efficient digital BF method is utilized by maximizing the SINR to design the digital BF instead of using the traditional ZF method.

For the ease of representation, we define the effective channel gain between i-th STA and j-th RF chain at AP as $h_{i,j} = \mathbf{v}_j^H \mathbf{H}_i \mathbf{u}_i \in \mathbb{C}^{1\times 1}$. Similarly, the effective channel vector between i-th STA and AP is represented by $\bar{\mathbf{h}}_i = \mathbf{V}^H \mathbf{H}_i \mathbf{u}_i \in \mathbb{C}^{K\times 1}$. Thus, the definition of SINR in (6) can be rewritten as

$$\gamma_i = \frac{|\mathbf{f}_i^H \bar{\mathbf{h}}_i|^2}{|\mathbf{f}_i^H (\sum_{j\neq i}^K \bar{\mathbf{h}}_j + \frac{1}{\sqrt{P}}\bar{\mathbf{n}})|^2}, \tag{11}$$

where $\bar{\mathbf{n}} \triangleq \mathbf{V}^H \mathbf{n} \in \mathbb{C}^{K\times 1}$ is the processed noise vector.

We aim to design digital BF to maximize SINR which can be formulated as a maximization problem, i.e.,

$$\mathcal{P}2: \max_{\mathbf{F}} \quad \frac{|\mathbf{f}_i^H \bar{\mathbf{h}}_i|^2}{|\mathbf{f}_i^H (\sum_{j\neq i}^K \bar{\mathbf{h}}_j + \frac{1}{\sqrt{P}}\bar{\mathbf{n}})|^2} \tag{12a}$$

$$\text{s.t.} \quad \|\mathbf{f}_i\|^2 = 1, \forall i. \tag{12b}$$

First, we simplify the objective function. Here we define $\hat{\mathbf{H}}_i = [\bar{\mathbf{h}}_1, ..., \bar{\mathbf{h}}_{i-1}, \bar{\mathbf{h}}_{i+1}, ..., \bar{\mathbf{h}}_K] \triangleq \sum_{j\neq i}^K \bar{\mathbf{h}}_j \in \mathbb{C}^{K\times K-1}$ as the aggregated effective channel matrix excludes $\bar{\mathbf{h}}_i$. Then, the (11) can be rewritten as

$$\gamma_i = \frac{|\mathbf{f}_i^H \bar{\mathbf{h}}_i|^2}{|\mathbf{f}_i^H (\hat{\mathbf{H}}_i + \frac{1}{\sqrt{P}}\bar{\mathbf{n}})|^2} = \frac{\mathbf{f}_i^H \bar{\mathbf{h}}_i \bar{\mathbf{h}}_i^H \mathbf{f}_i}{\mathbf{f}_i^H (\hat{\mathbf{H}}_i \hat{\mathbf{H}}_i^H + \frac{\sigma^2}{P}\mathbf{I}_K)\mathbf{f}_i}, \tag{13}$$

which is a Rayleigh quotient form. According to the Rayleigh-Ritz quotient theorem [15], γ_i is upper bounded by

$$\gamma_i \leq \lambda_{max}(\bar{\mathbf{h}}_i \bar{\mathbf{h}}_i^H, \hat{\mathbf{H}}_i \hat{\mathbf{H}}_i^H + \frac{\sigma^2}{P}\mathbf{I}_K), \tag{14}$$

where $\lambda_{max}(\mathbf{A})$ is the largest eigenvalue of matrix $\mathbf{A}$. Thus, the equality holds when $\mathbf{f}_i$ is the eigenvector corresponding to the largest eigenvalue. Thus, the optimal digital BF vector is

$$\mathbf{f}_i = \text{eigenvector}(\bar{\mathbf{h}}_i \bar{\mathbf{h}}_i^H, \hat{\mathbf{H}}_i \hat{\mathbf{H}}_i^H + \frac{\sigma^2}{P}\mathbf{I}_K). \tag{15}$$

If the matrix $\hat{\mathbf{H}}_i \hat{\mathbf{H}}_i^H$ is convertible, the optimal digital BF can be rewritten as

$$\mathbf{f}_i = \text{eigenvector}((\hat{\mathbf{H}}_i \hat{\mathbf{H}}_i^H + \frac{\sigma^2}{P}\mathbf{I}_K)^{-1} \bar{\mathbf{h}}_i \bar{\mathbf{h}}_i^H). \tag{16}$$

Due to unit power constraint of digital BF, we normalize each digital BF vector, i.e., $\mathbf{f}_i = \frac{\mathbf{f}_i}{\|\mathbf{f}_i\|}$.

Remark 1: Higher sum-rate is expected due to the use of the above SINR maximization method compared with the traditional ZF method. In addition, the computation complexity can be significantly reduced since only low-dimension effective CSI is required to obtain the digital BF.

TABLE I
SIMULATION PARAMETERS

Parameters	Value
Central frequency (f)	60 GHz
System bandwidth	2.16 GHz
Noise power spectral density	-134 dBm/MHz
LOS Corridor (n)	1.64
LOS hall (n)	2.17
NLOS hall (n)	3.01
STA antennas (N_t)	16
AP antennas (N_r)	32-128
Number of STAs (K)	8-64
Virtual Sectors (A)	[8, 16]

IV. Performance Evaluation

This section provides the simulation results obtained for the proposed algorithm compared with ZF and random selection methods. We perform Monte-Carlo simulation to evaluate the proposed algorithm for 10,000 times. The performance criterion is the achievable sum-rate of the selected STAs.

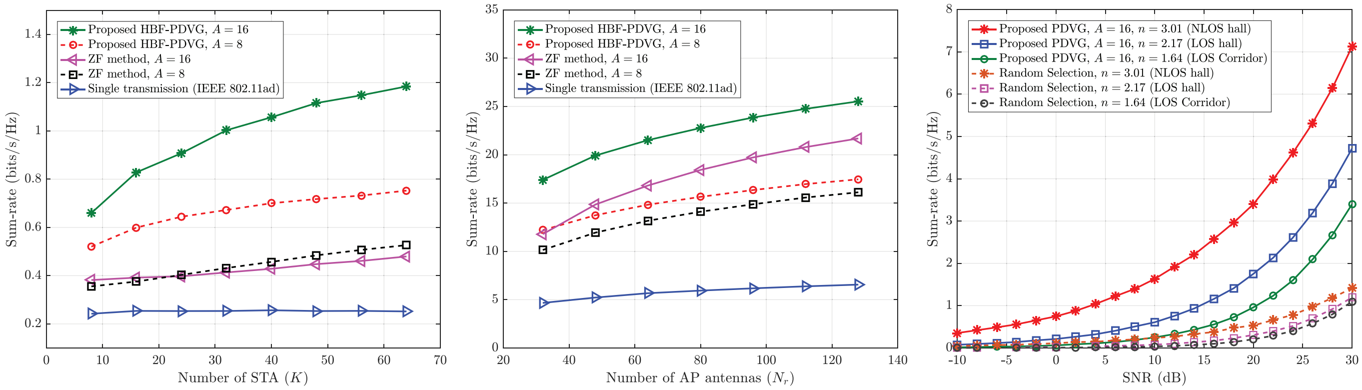
A. Simulation Setup

We consider an indoor mmWave environment where the AP is placed in the center of 5 m radius and K STAs are uniformly randomly distributed in the coverage area. The central frequency (f) is 60 GHz, and the system bandwidth is 2.16 GHz. The channel model defined in [16] is implemented for simulation. Different path loss exponents (n) based on the 60 GHz indoor environment measurements [17] are exploited in this simulation. The considered scenarios of these path loss exponents are line-of-sight (LOS) in a corridor, LOS in a hall, and non-line-of-sight (NLOS) in a hall. The simulation parameters are summarized in Table I.

B. Simulation Results

Fig. 4a shows the sum-rate performance versus the number of STAs (K). We compare our proposed algorithm with the ZF method. It can be observed that our proposed HBF-PDVG algorithm can achieve more sum-rate and this performance is increasing when the number of STAs is increased since higher probability to select STAs for concurrent transmission in VG_{g1} or VG_{g2} for different time slots can be achieved. In addition, the proposed HBF-PDVG algorithm can achieve higher sum-rate when the number of virtual sectors is 16 ($A = 16$) where in this case the spatial multiplexing gain is higher. This significant increase in sum-rate is expected since having 8 VSs to perform the STA selection increases the opportunity for simultaneous transmission. The directivity and beamwidth should be considered when the number of VS is increased since there is a trade-off between the maximum spatial multiplexing gain and beamwidth [3].

Fig. 4b shows the sum-rate performance and the relationship between the sum-rate and the number of AP antennas (N_r), where the number of the STA antenna is $N_t = 16$. As the number N_r increases, the performance gain of the proposed HBF-PDVG algorithm also increases. Likewise in Fig. 4a, a



(a) Sum-rate versus the number of total STAs ($N_r = 36$, SNR = 30 dB).

(b) Sum-rate versus the number of AP antennas (SNR = 60 dB).

(c) Sum-rate versus the SNR for different path loss exponent ($N_r = 64$).

Fig. 4. Performance evaluation results.

higher sum-rate gain of HPF-PDVG algorithm can be achieved compared with the ZF method. Furthermore, higher gain is attained when the number of VSs is 16 because a higher number of simultaneous STAs can be selected, up to eight simultaneous STAs when $A = 16$ while only up to four STAs when $A = 8$. It can be seen that as higher antennas are employed, larger BF gain can be achieved which results in greater sum-rate.

In Fig. 4c, we compare our proposed algorithm with the random selection method where the sum-rate versus SNR for different path loss exponents is evaluated. The proposed HBF-PDVG algorithm outperforms the random selection method and achieves higher sum-rate in all path loss exponent cases since the set of most effective simultaneous STAs is selected separately by using the PDVG algorithm utilizing the analog BF from the first phase. In addition, our proposed HBF-PDVG algorithm adopts a criterion which maximizes the SINR in the third phase to improve the performance instead of using the ZF technique.

V. Conclusion

A novel HBF-PDVG algorithm has been proposed in this paper to attain UL MU-MIMO transmission for the mmWave systems. The proposed algorithm employs HBF design where the algorithm is divided into three phases. The main idea is to select the concurrent STAs independently by utilizing the PDVG algorithm after completing the analog BF training. Digital BF is then exploited to mitigate the residual MU interference among the selected STAs. The simulation results have demonstrated that the proposed HBF-PDVG algorithm has significant sum-rate performance advantages over the random selection and ZF method. For future work, the performance of the proposed algorithm will be investigated and analyzed for different channel scenarios. The inter-cell interference problem will also be investigated for the UL MU-MIMO mmWave scenario when the coverage of several hops are overlapped.

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