# Energy-Efficient Hybrid Analog and Digital Precoding for mmWave MIMO Systems with Large Antenna Arrays

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Abstract-Millimeter wave (mmWave) MIMO will likely use hybrid analog and digital precoding, which uses a small number of RF chains to avoid energy consumption associated with mixed signal components like analog-to-digital components not to mention baseband processing complexity. However, most hybrid precoding techniques consider a fully-connected architecture requiring a large number of phase shifters, which is also energyintensive. In this paper, we focus on the more energy-efficient hybrid precoding with sub-connected architecture, and propose a successive interference cancelation (SIC)-based hybrid precoding with near-optimal performance and low complexity. Inspired by the idea of SIC for multi-user signal detection, we first propose to decompose the total achievable rate optimization problem with non-convex constraints into a series of simple sub-rate optimization problems, each of which only considers one sub-antenna array. Then, we prove that maximizing the achievable sub-rate of each sub-antenna array is equivalent to simply seeking a precoding vector sufficiently close (in terms of Euclidean distance) to the unconstrained optimal solution. Finally, we propose a low-complexity algorithm to realize SICbased hybrid precoding, which can avoid the need for the singular value decomposition (SVD) and matrix inversion. Complexity evaluation shows that the complexity of SIC-based hybrid precoding is only about 10% as complex as that of the recently proposed spatially sparse precoding in typical mmWave MIMO systems. Simulation results verify the near-optimal performance of SIC-based hybrid precoding.

*Index Terms*—MIMO, mmWave communications, hybrid precoding, energy-efficient, 5G.

# I. INTRODUCTION

The integration of millimeter-wave (mmWave) and multiple-input multiple-output (MIMO) can achieve orders of magnitude increase in rates due to larger bandwidth and greater spectral efficiency [1]. This makes mmWave MIMO as a promising technique for future 5G wireless communication systems [2]. On one hand, the decreased wavelength associated

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with high frequencies of mmWave enables a large antenna array to be packed in small physical dimension [3]. On the other hand, the large antenna array can provide sufficient antenna gain to compensate for the severe attenuation of mmWave signals due to path loss, oxygen absorption, and rainfall effect [4]. Additionally, the large antenna array can also support the transmission of multiple data streams to improve the spectral efficiency through the use of precoding [5].

For MIMO in conventional cellular frequency band (e.g., 2-3 GHz), precoding is entirely realized in the digital domain to cancel the interferences between different data streams. Digital precoding requires an expensive radio frequency (RF) chain (including digital-to-analog converter, up converter, etc.) for every antenna. In mmWave MIMO system with a large number of antennas, it will bring prohibitively high energy consumption and hardware complexity. To solve this problem, mmWave MIMO prefers the more energy-efficient hybrid analog and digital precoding [3], which can significantly reduce the number of required RF chains. Specifically, the transmitted signals are first precoded by the digital precoding of a small dimension to guarantee the performance, and then precoded again by the analog precoding of a large dimension to save the energy consumption and reduce the hardware complexity.

To realize the hybrid precoding in practice, two categories of techniques have been proposed recently. The first category is based on the spatially sparse precoding [6]-[8], which formulates the achievable rate optimization problem as a sparse approximation problem and solves it by orthogonal matching pursuit (OMP) [9] to achieve the near-optimal performance. The second category of hybrid precoding based on codebook is proposed in [10]-[12], which involves an iterative searching procedure among the predefined codebook to find the optimal hybrid precoding matrix. These algorithms are all designed for the hybrid precoding with fully-connected architecture, where each RF chain is connected to all BS antennas via phase shifters. As the number of BS antennas is very large (e.g., 256 as considered in [6]), the fully-connected architecture has two possible limitations. First, it requires thousands of phase shifters like the giant phased array radar to realize the analog precoding [13], which leads to both high energy consumption and hardware complexity. Second, each RF chain will drive hundreds of BS antennas, which is also energy-intensive [13]. By contrast, the hybrid precoding with sub-connected architecture, where each RF chain is connected to only a subset of BS antennas, can reduce the number of required phase

shifters without obvious performance loss [3]. Therefore, the sub-connected architecture is expected to be more energy-efficient and implementation-practical for mmWave MIMO systems. Unfortunately, the design of hybrid precoding with sub-connected architecture is still an open problem [3], [14], as the sub-connected architecture changes the constraints on the original hybrid precoding problem.

In this paper, we propose a successive interference cancelation (SIC)-based hybrid precoding with sub-connected architecture. The contributions of this paper can be summarized as follows.

- 1) Inspired by the idea of SIC derived for multi-user signal detection [15], we propose to decompose the total achievable rate optimization problem with non-convex constraints into a series of simple sub-rate optimization problems, each of which only considers one sub-antenna array. Then, we can maximize the achievable sub-rate of each sub-antenna array one by one until the last sub-antenna array is considered.
- 2) We prove that maximizing the achievable sub-rate of each sub-antenna array is equivalent to seeking a precoding vector which has the smallest Euclidean distance to the unconstrained optimal solution. Based on this fact, we can easily obtain the optimal precoding vector for each sub-antenna array.
- 3) We further propose a low-complexity algorithm to realize the SIC-based precoding, which avoids the need for singular value decomposition (SVD) and matrix inversion. Complexity evaluation shows that the complexity of SIC-based precoding is only about 10% as complex as that of the spatially sparse precoding [6] in typical mmWave MIMO systems, while it can still achieve the near-optimal performance as verified by simulation results.

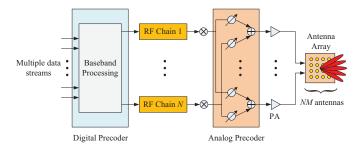
It is worth pointing out that to the best of the authors' knowledge, our work in this paper is the first one that considers the hybrid precoding design with sub-connected architecture.

The rest of the paper is organized as follows. Section II briefly introduces the system model of mmWave MIMO. Section III specifies the proposed SIC-based hybrid precoding, together with the complexity evaluation. The simulation results of the achievable rate are shown in Section IV. Finally, conclusions are drawn in Section V.

Notation: Lower-case and upper-case boldface letters denote vectors and matrices, respectively;  $(\cdot)^T$ ,  $(\cdot)^H$ ,  $(\cdot)^{-1}$ , and  $|\cdot|$  denote the transpose, conjugate transpose, inversion, and determinant of a matrix, respectively;  $||\cdot||_1$  and  $||\cdot||_2$  denote the  $l_1$ - and  $l_2$ -norm of a vector, respectively;  $||\cdot||_F$  denotes the Frobenius norm of a matrix;  $\operatorname{Re}\{\cdot\}$  and  $\operatorname{Im}\{\cdot\}$  denote the real part and imaginary part of a complex number, respectively;  $\mathbb{E}(\cdot)$  denotes the expectation; Finally,  $\mathbf{I}_N$  is the  $N\times N$  identity matrix.

#### II. SYSTEM MODEL

Fig. 1 illustrates two typical architectures for hybrid precoding in mmWave MIMO systems, i.e., the fully-connected architecture as shown in Fig. 1 (a) and the sub-connected architecture as shown in Fig. 1 (b). In both cases the BS has NM antennas but only N RF chains. From Fig. 1, we observe that the sub-connected architecture will likely be more energy-efficient, since it only requires NM phase shifters, while the



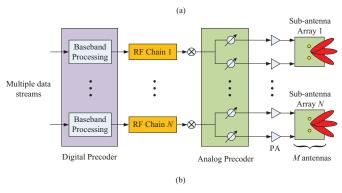


Fig. 1. Two typical architectures of the hybrid precoding in mmWave MIMO systems: (a) Fully-connected architecture, where each RF chain is connected to all BS antennas; (b) Sub-connected architecture, where each RF chain is connected to only a subset of BS antennas.

fully-connected architecture requires  $N^2M$  phase shifters. To fully achieve the spatial multiplexing gain, the BS usually transmits N independent data streams to users employing K receive antennas [3].

In the sub-connected architecture as shown in Fig. 1 (b), N data streams in the baseband are precoded by the digital precoder D. In cases where complexity is a concern, D can be further specialized to be a diagonal matrix as  $\mathbf{D} = \operatorname{diag} [d_1, d_2, \cdots, d_N]$ , where  $d_n \in \mathbb{R}$  for  $n=1,2,\cdots,N$  [3]. Then the role of **D** essentially performs some power allocation. After passing through the corresponding RF chain, the digital-domain signal from each RF chain is delivered to only M phase shifters [16] to perform the analog precoding, which can be denoted by the analog weighting vector  $\bar{\mathbf{a}}_n \in \mathbb{C}^{M \times 1}$ , whose elements have the same amplitude  $1/\sqrt{M}$  but different phases [16]. After the analog precoding, each data stream is finally transmitted by a sub-antenna array with only M antennas associated with the corresponding RF chain. Then, the received signal vector  $\mathbf{y} = [y_1, y_2, \dots, y_K]^T$ at the user in a narrowband system 1 can be presented as

$$y = \rho HADs + n = \rho HPs + n, \tag{1}$$

where  $\rho$  is the average received power;  $\mathbf{H} \in \mathbb{C}^{K \times NM}$  denotes the channel matrix,  $\mathbf{A}$  is the  $NM \times N$  analog precoding matrix comprising N analog weighting vectors  $\{\bar{\mathbf{a}}_m\}_{m=1}^N$  as

A = 
$$\begin{bmatrix} \bar{\mathbf{a}}_1 & \mathbf{0} & \dots & \mathbf{0} \\ \mathbf{0} & \bar{\mathbf{a}}_2 & \mathbf{0} \\ \vdots & & \ddots & \vdots \\ \mathbf{0} & \mathbf{0} & \dots & \bar{\mathbf{a}}_N \end{bmatrix}_{NM \times N},$$
(2)

<sup>1</sup>While mmWave systems are expected to be broadband as in prior work [2], the narrowband system can be regarded as a reasonable first step. The extension to broadband system is an interesting topic of future work.

 $\mathbf{s} = [s_1, s_2, \cdots, s_N]^T$  represents the transmitted signal vector in the baseband, and usually  $\mathbb{E}(\mathbf{s}\mathbf{s}^H) = \frac{1}{N}\mathbf{I}_N$  is assumed for the normalized signal power [6].  $\mathbf{P} = \mathbf{A}\mathbf{D}$  presents the hybrid precoding matrix of size  $NM \times N$ , which satisfies  $\|\mathbf{P}\|_F \leq N$  to meet the total transmit power constraint [6]. Finally,  $\mathbf{n} = [n_1, n_2, \cdots, n_N]^T$  is an additive white Gaussian noise (AWGN) vector, whose entries follow the independent and identical distribution (i.i.d.)  $\mathcal{CN}(0, \sigma^2)$ .

It is known that mmWave channel **H** will not likely follow the rich-scattering model assumed at low frequencies due to the limited number of scatters in the mmWave prorogation environment [2]. In this paper, we adopt the geometric Saleh-Valenzuela channel model for mmWave communications, which was also used in related work in [17] as

$$\mathbf{H} = \gamma \sum_{l=1}^{L} \alpha_{l} \Lambda_{r} \left( \phi_{l}^{r}, \theta_{l}^{r} \right) \Lambda_{t} \left( \phi_{l}^{t}, \theta_{l}^{t} \right) \mathbf{f}_{r} \left( \phi_{l}^{r}, \theta_{l}^{r} \right) \mathbf{f}_{t}^{H} \left( \phi_{l}^{t}, \theta_{l}^{t} \right),$$
(3)

where  $\gamma = \sqrt{\frac{NMK}{L}}$  is a normalization factor, L is the number of effective channel paths corresponding to the limited number of scatters, and we usually have  $L \leq N$  for mmWave communication systems.  $\alpha_l \in \mathbb{C}$  is the gain of the lth path.  $\phi_l^t$  ( $\theta_l^t$ ) and  $\phi_l^r$  ( $\theta_l^r$ ) are the azimuth (elevation) angles of departure and arrival (AoDs/AoAs), respectively.  $\Lambda_t$  ( $\phi_l^t, \theta_l^t$ ) and  $\Lambda_r$  ( $\phi_l^r, \theta_l^r$ ) denote the transmit and receive antenna array gain at a specific AoD and AoA, respectively. For simplicity but without loss of generality,  $\Lambda_t$  ( $\phi_l^t, \theta_l^t$ ) and  $\Lambda_r$  ( $\phi_l^r, \theta_l^r$ ) can be set as one within the range of AoDs/AoAs [18]. Finally,  $\mathbf{f}_t$  ( $\phi_l^t, \theta_l^t$ ) and  $\mathbf{f}_r$  ( $\phi_l^r, \theta_l^r$ ) are the antenna array response vectors depending on the antenna array structures at the BS and the user, respectively. For the uniform linear array (ULA) with U elements, the array response vector can be presented as [13]

$$\mathbf{f}_{\text{ULA}}(\phi) = \frac{1}{\sqrt{U}} \left[ 1, e^{j\frac{2\pi}{\lambda} d \sin(\phi)}, \cdots, e^{j(U-1)\frac{2\pi}{\lambda} d \sin(\phi)} \right]^{T}, \tag{4}$$

where  $\lambda$  denotes the wavelength of the signal, and d is the antenna spacing. Note that here we abandon the subscripts  $\{t,r\}$  in (3) and we also do not include  $\theta$  in (4) since the ULA response vector is independent of the elevation angle. Additionally, when we consider the uniform planar array (UPA) with  $W_1$  and  $W_2$  elements ( $W_1W_2=U$ ) on horizon and vertical, respectively, the array response vector can be given by [13]

$$\mathbf{f}_{\text{UPA}}(\phi, \theta) = \frac{1}{\sqrt{U}} \left[ 1, \cdots, e^{j\frac{2\pi}{\lambda}d(x\sin(\phi)\sin(\theta) + y\cos(\theta))}, \cdots, e^{j\frac{2\pi}{\lambda}d((W_1 - 1)\sin(\phi)\sin(\theta) + (W_2 - 1)\cos(\theta))} \right]^T,$$
(5)

where  $0 \le x \le (W_1 - 1)$  and  $0 \le y \le (W_2 - 1)$ .

# III. SIC-BASED HYBRID PRECODING FOR MMWAVE MIMO SYSTEMS

In this section, we propose a low-complexity SIC-based hybrid precoding to achieve the near-optimal performance. The evaluation of computational complexity is also provided to show its advantages over current solutions. A. Structure of SIC-based hybrid precoding

The final aim of precoding is to maximize the total achievable rate R of mmWave MIMO systems, which can be expressed as [6]

$$R = \log_2\left(\left|\mathbf{I}_N + \frac{\rho}{N\sigma^2}\mathbf{H}\mathbf{P}\mathbf{P}^H\mathbf{H}^H\right|\right). \tag{6}$$

According to the system model (1) in Section II, since the hybrid precoding matrix  $\mathbf{P}$  can be represented as  $\mathbf{P} = \mathbf{A}\mathbf{D} = \mathrm{diag}\left\{\bar{\mathbf{a}}_1, \cdots, \bar{\mathbf{a}}_N\right\} \cdot \mathrm{diag}\left\{d_1, \cdots, d_N\right\}$ , there are three constraints for the design of  $\mathbf{P}$ .

Constraint 1: **P** should be a block diagonal matrix similar to the form of **A** as shown in (2), i.e.,  $\mathbf{P} = \operatorname{diag} \{\bar{\mathbf{p}}_1, \cdots, \bar{\mathbf{p}}_N\}$ , where  $\bar{\mathbf{p}}_n = d_n \bar{\mathbf{a}}_n$  is the  $M \times 1$  non-zero vector of the nth column  $\mathbf{p}_n$  of **P**, i.e.,  $\mathbf{p}_n = \begin{bmatrix} \mathbf{0}_{1 \times M(n-1)}, \ \bar{\mathbf{p}}_n, \ \mathbf{0}_{1 \times M(N-n)} \end{bmatrix}^T$ ;

Constraint 2: The non-zero elements of each column of **P** should have the same amplitude, since the digital precoding matrix **D** is a diagonal matrix, and the amplitude of non-zero elements of the analog precoding matrix **A** is fixed to  $1/\sqrt{M}$ ;

Constraint 3: The Frobenius norm of **P** should satisfy  $\|\mathbf{P}\|_F \leq N$  to meet the total transmit power constraint.

Unfortunately, these non-convex constraints on **P** make maximizing the total achievable rate (6) very difficult to be solved. However, based on the special block diagonal structure of the hybrid precoding matrix **P**, we can observe that the precoding on different sub-antenna arrays are independent. This inspires us to decompose the total achievable rate (6) into a series of sub-rate optimization problems, each of which only considers one sub-antenna array.

In particular, we can divide the hybrid precoding matrix  $\mathbf{P}$  as  $\mathbf{P} = [\mathbf{P}_{N-1} \ \mathbf{p}_N]$ , where  $\mathbf{p}_N$  is the Nth column of  $\mathbf{P}$ , and  $\mathbf{P}_{N-1}$  is an  $NM \times (N-1)$  matrix containing the first (N-1) columns of  $\mathbf{P}$ . Then, the total achievable rate R in (6) can be rewritten as

$$R = \log_{2} \left( \left| \mathbf{I}_{N} + \frac{\rho}{N\sigma^{2}} \mathbf{H} \mathbf{P} \mathbf{P}^{H} \mathbf{H}^{H} \right| \right)$$

$$= \log_{2} \left( \left| \mathbf{I}_{N} + \frac{\rho}{N\sigma^{2}} \mathbf{H} \left[ \mathbf{P}_{N-1} \mathbf{p}_{N} \right] \left[ \mathbf{P}_{N-1} \mathbf{p}_{N} \right]^{H} \mathbf{H}^{H} \right| \right)$$

$$= \log_{2} \left( \left| \mathbf{I}_{N} + \frac{\rho}{N\sigma^{2}} \mathbf{H} \mathbf{P}_{N-1} \mathbf{P}_{N-1}^{H} \mathbf{H}^{H} \right|$$

$$+ \frac{\rho}{N\sigma^{2}} \mathbf{H} \mathbf{p}_{N} \mathbf{p}_{N}^{H} \mathbf{H}^{H} \right| \right)$$

$$\stackrel{(a)}{=} \log_{2} \left( \left| \mathbf{T}_{N-1} \right| \right) + \log_{2} \left( \left| \mathbf{I}_{N} + \frac{\rho}{N\sigma^{2}} \mathbf{T}_{N-1}^{-1} \mathbf{H} \mathbf{p}_{N} \mathbf{p}_{N}^{H} \mathbf{H}^{H} \right| \right)$$

$$\stackrel{(b)}{=} \log_{2} \left( \left| \mathbf{T}_{N-1} \right| \right) + \log_{2} \left( 1 + \frac{\rho}{N\sigma^{2}} \mathbf{p}_{N}^{H} \mathbf{H}^{H} \mathbf{T}_{N-1}^{-1} \mathbf{H} \mathbf{p}_{N} \right), \tag{7}$$

where (a) is obtained by defining the auxiliary matrix  $\mathbf{T}_{N-1} = \mathbf{I}_N + \frac{\rho}{N\sigma^2}\mathbf{H}\mathbf{P}_{N-1}\mathbf{P}_{N-1}^H\mathbf{H}^H$ , and (b) is true due to the fact that  $|\mathbf{I} + \mathbf{X}\mathbf{Y}| = |\mathbf{I} + \mathbf{Y}\mathbf{X}|$  by defining  $\mathbf{X} = \mathbf{T}_{N-1}^{-1}\mathbf{H}\mathbf{p}_N$  and  $\mathbf{Y} = \mathbf{p}_N^H\mathbf{H}^H$ . Note that the second term  $\log_2\left(1 + \frac{\rho}{N\sigma^2}\mathbf{p}_N^H\mathbf{H}^H\mathbf{T}_{N-1}^{-1}\mathbf{H}\mathbf{p}_N\right)$  on the right side of (7) is the achievable sub-rate of the Nth sub-antenna array, while the first term  $\log_2\left(|\mathbf{T}_{N-1}|\right)$  shares the same form as (6). This observation implies that we can further decompose

 $\log_2(|\mathbf{T}_{N-1}|)$  using the similar method in (7) as

$$\log_2\left(|\mathbf{T}_{N-2}|\right) + \log_2\left(1 + \frac{\rho}{N\sigma^2}\mathbf{p}_{N-1}^H\mathbf{H}^H\mathbf{T}_{N-2}^{-1}\mathbf{H}\mathbf{p}_{N-1}\right).$$

Then, after N such decompositions, the total achievable rate R in (6) can be presented as

$$R = \sum_{n=1}^{N} \log_2 \left( 1 + \frac{\rho}{N\sigma^2} \mathbf{p}_n^H \mathbf{H}^H \mathbf{T}_{n-1}^{-1} \mathbf{H} \mathbf{p}_n \right), \tag{8}$$

where we have  $\mathbf{T}_n = \mathbf{I}_N + \frac{\rho}{N\sigma^2}\mathbf{H}\mathbf{P}_n\mathbf{P}_n^H\mathbf{H}^H$  and  $\mathbf{T}_0 = \mathbf{I}_N$ . From (8), we observe that the total achievable rate optimization problem can be transformed into a series of sub-rate optimization problems of sub-antenna arrays, which can be optimized one by one. After that, inspired by the idea of SIC for multiuser signal detection [15], we can optimize the achievable sub-rate of the first sub-antenna array and update the matrix  $\mathbf{T}_1$ . Then, the similar method can be utilized to optimize the achievable sub-rate of the second sub-antenna array. Such procedure will be executed until the last sub-antenna array is considered. Fig. 2 shows the diagram of the proposed SIC-based hybrid precoding. Next, we will discuss how to optimize the achievable sub-rate of each sub-antenna array.

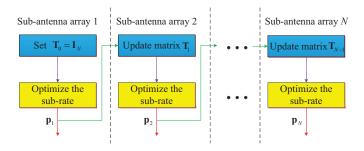


Fig. 2. Diagram of the proposed SIC-based hybrid precoding.

## B. Solution to the sub-rate optimization problem

In this subsection, we focus on the sub-rate optimization problem of the nth sub-antenna array, which can be directly applied to other sub-antenna arrays. According to (8), the sub-rate optimization problem of the nth sub-antenna array by designing the nth precoding vector  $\mathbf{p}_n$  can be stated as

$$\mathbf{p}_{n}^{\text{opt}} = \underset{\mathbf{p}_{n} \in \mathcal{F}}{\operatorname{arg \, max} \, \log_{2} \left( 1 + \frac{\rho}{N\sigma^{2}} \mathbf{p}_{n}^{H} \mathbf{G}_{n-1} \mathbf{p}_{n} \right), \quad (9)$$

where  $\mathbf{G}_{n-1}$  is defined as  $\mathbf{G}_{n-1} = \mathbf{H}^H \mathbf{T}_{n-1}^{-1} \mathbf{H}$ ,  $\mathcal{F}$  is the set of all feasible vectors satisfying the three constraints described in Section III-A. Note that the nth precoding vector  $\mathbf{p}_n$  only has M non-zero elements from the (M(n-1)+1)th one to the (Mn)th one. Therefore, the sub-rate optimization problem (9) can be equivalently written as

$$\bar{\mathbf{p}}_{n}^{\text{opt}} = \operatorname*{arg\,max}_{\bar{\mathbf{p}}_{n} \in \bar{\mathcal{F}}} \log_{2} \left( 1 + \frac{\rho}{N\sigma^{2}} \bar{\mathbf{p}}_{n}^{H} \bar{\mathbf{G}}_{n-1} \bar{\mathbf{p}}_{n} \right), \tag{10}$$

where  $\bar{\mathcal{F}}$  includes all possible  $M \times 1$  vectors satisfying *Constraint 2* and *Constraint 3*,  $\bar{\mathbf{G}}_{n-1}$  of size  $M \times M$  is the corresponding sub-matrix of  $\mathbf{G}_{n-1}$  by only keeping the rows and columns of  $\mathbf{G}_{n-1}$  from the (M(n-1)+1)th one to the (Mn)th one, which can be presented as

$$\mathbf{\bar{G}}_{n-1} = \mathbf{R}\mathbf{G}_{n-1}\mathbf{R}^H = \mathbf{R}\mathbf{H}^H\mathbf{T}_{n-1}^{-1}\mathbf{H}\mathbf{R}^H, \quad (11)$$

where  $\mathbf{R} = \begin{bmatrix} \mathbf{0}_{M \times M(n-1)} & \mathbf{I}_M & \mathbf{0}_{M \times M(N-n)} \end{bmatrix}$  is the corresponding selection matrix.

Define the singular value decomposition (SVD) of the Hermitian matrix  $\bar{\mathbf{G}}_{n-1}$  as  $\bar{\mathbf{G}}_{n-1} = \mathbf{V} \mathbf{\Sigma} \mathbf{V}^H$ , where  $\mathbf{\Sigma}$  is an  $M \times M$  diagonal matrix containing the singular values of  $\bar{\mathbf{G}}_{n-1}$  in a decreasing order, and  $\mathbf{V}$  is an  $M \times M$  unitary matrix. It is known that the optimal unconstrained precoding vector of (10) is the first column  $\mathbf{v}_1$  of  $\mathbf{V}$ , i.e., the first right singular vector of  $\bar{\mathbf{G}}_{n-1}$  [6]. According to the constraints mentioned in Section III-A, we cannot directly choose  $\bar{\mathbf{p}}_n^{\text{opt}}$  as  $\mathbf{v}_1$  since the elements of  $\mathbf{v}_1$  do not obey the constraint of same amplitude (i.e., *Constraint 2*). To find a practical solution to the sub-rate optimization problem (10), we need to further convert (10) into another form, which is given by the following **Proposition 1**.

**Proposition 1**. The optimization problem (10)

$$\bar{\mathbf{p}}_n^{\text{opt}} = \operatorname*{arg\,max}_{\bar{\mathbf{p}}_n \in \bar{\mathcal{F}}} \log_2 \left( 1 + \frac{\rho}{N\sigma^2} \bar{\mathbf{p}}_n^H \bar{\mathbf{G}}_{n-1} \bar{\mathbf{p}}_n \right)$$

is equivalent to the following problem

$$\bar{\mathbf{p}}_n^{\text{opt}} = \underset{\bar{\mathbf{p}}_n \in \bar{\mathcal{F}}}{\text{arg min}} \left\| \mathbf{v}_1 - \bar{\mathbf{p}}_n \right\|_2^2, \tag{12}$$

where  $\mathbf{v}_1$  is the first right singular vector of  $\mathbf{\bar{G}}_{n-1}$ .

Proof: See Appendix A.

**Proposition 1** indicates that we can find a feasible precoding vector  $\bar{\mathbf{p}}_n$ , which is sufficiently close (in terms of Euclidean distance) to the optimal but unpractical precoding vector  $\mathbf{v}_1$ , to maximize the achievable sub-rate of the nth sub-antenna array. Since  $\bar{\mathbf{p}}_n = d_n \bar{\mathbf{a}}_n$  according to (1), the target  $\|\mathbf{v}_1 - \bar{\mathbf{p}}_n\|_2^2$  in (12) can be rewritten as

$$\|\mathbf{v}_{1} - \mathbf{\bar{p}}_{n}\|_{2}^{2}$$

$$= (\mathbf{v}_{1} - d_{n}\mathbf{\bar{a}}_{n})^{H} (\mathbf{v}_{1} - d_{n}\mathbf{\bar{a}}_{n})$$

$$= \mathbf{v}_{1}^{H}\mathbf{v}_{1} + d_{n}^{2}\mathbf{\bar{a}}_{n}^{H}\mathbf{\bar{a}}_{n} - 2d_{n}\operatorname{Re}(\mathbf{v}_{1}^{H}\mathbf{\bar{a}}_{n})$$

$$\stackrel{(a)}{=} 1 + d_{n}^{2} - 2d_{n}\operatorname{Re}(\mathbf{v}_{1}^{H}\mathbf{\bar{a}}_{n})$$

$$= (d_{n} - \operatorname{Re}(\mathbf{v}_{1}^{H}\mathbf{\bar{a}}_{n}))^{2} + (1 - [\operatorname{Re}(\mathbf{v}_{1}^{H}\mathbf{\bar{a}}_{n})]^{2}), \quad (13)$$

where (a) is obtained based on the facts that  $\mathbf{v}_1^H \mathbf{v}_1 = 1$  and  $\bar{\mathbf{a}}_n^H \bar{\mathbf{a}}_n = 1$ , since  $\mathbf{v}_1$  is the first column of the unitary matrix  $\mathbf{V}$  and each element of  $\bar{\mathbf{a}}_n$  has the same amplitude  $1/\sqrt{M}$ .

From (13), we observe that the distance between  $\bar{\mathbf{p}}_n$  and  $\mathbf{v}_1$  consists of two parts. The first one is  $\left(d_n - \operatorname{Re}\left(\mathbf{v}_1^H \bar{\mathbf{a}}_n\right)\right)^2$ , which can be minimized to zero by choosing  $d_n = \operatorname{Re}\left(\mathbf{v}_1^H \bar{\mathbf{a}}_n\right)$ . The second one is  $\left(1 - \left[\operatorname{Re}\left(\mathbf{v}_1^H \bar{\mathbf{a}}_n\right)\right]^2\right)$ , which can be minimized by maximizing  $\left|\operatorname{Re}\left(\mathbf{v}_1^H \bar{\mathbf{a}}_n\right)\right|$ . Note that both  $\bar{\mathbf{a}}_n$  and  $\mathbf{v}_1$  have a fixed power of one, i.e.,  $\mathbf{v}_1^H \mathbf{v}_1 = 1$  and  $\bar{\mathbf{a}}_n^H \bar{\mathbf{a}}_n = 1$ . Therefore, the optimal  $\bar{\mathbf{a}}_n^{\text{opt}}$  to maximize  $\left|\operatorname{Re}\left(\mathbf{v}_1^H \bar{\mathbf{a}}_n\right)\right|$  is

$$\bar{\mathbf{a}}_n^{\text{opt}} = \frac{1}{\sqrt{M}} e^{j \text{angle}(\mathbf{v}_1)},\tag{14}$$

where  $angle(\mathbf{v}_1)$  denotes the phase vector of  $\mathbf{v}_1$ , i.e., each element of  $\bar{\mathbf{a}}_n^{\text{opt}}$  shares the same phase as the corresponding

element of  $\mathbf{v}_1$ . Accordingly, the optimal choice of  $d_n^{\text{opt}}$  is

$$d_n^{\text{opt}} = \text{Re}\left(\mathbf{v}_1^H \bar{\mathbf{a}}_n\right) = \frac{1}{\sqrt{M}} \text{Re}\left(\mathbf{v}_1^H e^{j \text{angle}(\mathbf{v}_1)}\right) = \frac{\|\mathbf{v}_1\|_1}{\sqrt{M}}.$$
(15)

Based on (14) and (15), the optimal solution  $\bar{\mathbf{p}}_n^{\text{opt}}$  to the optimization problem (12) (or equivalently (10)) can be obtained by

$$\bar{\mathbf{p}}_n^{\text{opt}} = d_n^{\text{opt}} \bar{\mathbf{a}}_n^{\text{opt}} = \frac{1}{M} \|\mathbf{v}_1\|_1 e^{j \text{angle}(\mathbf{v}_1)}.$$
 (16)

It is worth pointing out that  $\mathbf{v}_1$  is the first column of the unitary matrix  $\mathbf{V}$ , each element  $v_i$  of  $\mathbf{v}_1$  (for  $i=1,\cdots,M$ ) has the amplitude less than one. Therefore, we have  $\|\mathbf{\bar{p}}_n^{\mathrm{opt}}\|_2^2 \leq 1$ . Note that for all sub-antenna arrays, the optimal solution  $\mathbf{\bar{p}}_n^{\mathrm{opt}}$  for  $n=1,2,\cdots,N$  have a similar form. Thus, we can conclude that

$$\|\mathbf{P}^{\text{opt}}\|_F^2 = \|\operatorname{diag}\left\{\bar{\mathbf{p}}_1^{\text{opt}}, \cdots, \bar{\mathbf{p}}_N^{\text{opt}}\right\}\|_F^2 \le N,$$
 (17)

which demonstrates that the total transmit power constraint (*Constraint 3*) is satisfied.

After we have acquired  $\bar{\mathbf{p}}_n^{\text{opt}}$  for the nth sub-antenna array, the matrices  $\mathbf{T}_n = \mathbf{I}_N + \frac{\rho}{N\sigma^2}\mathbf{H}\mathbf{P}_n\mathbf{P}_n^H\mathbf{H}^H$  (8) and  $\bar{\mathbf{G}}_n = \mathbf{R}\mathbf{H}^H\mathbf{T}_n^{-1}\mathbf{H}\mathbf{R}^H$  (11) can be updated. Then, the method described above for the nth sub-antenna array can be reused again to optimize the achievable sub-rate of the (n+1)th sub-antenna array. To sum up, solving the sub-rate optimization problem of the nth sub-antenna array consists of the following three steps.

Step 1: Execute the SVD of  $\bar{\mathbf{G}}_{n-1}$  to obtain  $\mathbf{v}_1$ ;

Step 2: Let  $\bar{\mathbf{p}}_n^{\text{opt}} = \frac{1}{M} \|\mathbf{v}_1\|_1 e^{j \text{angle}(\mathbf{v}_1)}$  as the optimal solution to the current nth sub-antenna array;

Step 3: Update matrices  $\mathbf{T}_n = \mathbf{I}_N + \frac{\rho}{N\sigma^2}\mathbf{H}\mathbf{P}_n\mathbf{P}_n^H\mathbf{H}^H$  and  $\bar{\mathbf{G}}_n = \mathbf{R}\mathbf{H}^H\mathbf{T}_n^{-1}\mathbf{H}\mathbf{R}^H$  for the next (n+1)th sub-antenna array.

Note that although we can obtain the optimal solution  $\bar{\mathbf{p}}_n^{\text{opt}}$  by the method above, we need to compute the SVD of  $\bar{\mathbf{G}}_{n-1}$  (Step I) and the matrix  $\bar{\mathbf{G}}_n$  (Step 3) involving the matrix inversion of large size, which leads to high computational complexity as well as high hardware complexity. To this end, next we will propose a low-complexity algorithm to obtain  $\bar{\mathbf{p}}_n^{\text{opt}}$  to avoid the complicated SVD and matrix inversion.

#### C. Low-complexity algorithm to obtain the optimal solution

We start by considering how to avoid the SVD involving high computational complexity as well as a large number of divisions, which are difficult to be implemented in hardware. We can observe from  $Step\ I$  that the SVD of  $\mathbf{\bar{G}}_{n-1}$  does not need to be computed to acquire  $\mathbf{\Sigma}$  and  $\mathbf{V}$ , as only the first column  $\mathbf{v}_1$  of  $\mathbf{V}$  is enough to obtain  $\mathbf{\bar{p}}_n^{\text{opt}}$ . This observation inspires us to exploit the simple power iteration algorithm [19], which is used to compute the largest eigenvalue and the corresponding eigenvector of a diagonalizable matrix. Since  $\mathbf{\bar{G}}_{n-1}$  is a Hermitian matrix, it follows that: 1)  $\mathbf{\bar{G}}_{n-1}$  is also a diagonalizable matrix; 2) The singular values (right singular vectors) of  $\mathbf{\bar{G}}_{n-1}$  are same as the eigenvalues (eigenvectors). Therefore, the power iteration algorithm can be also utilized to

Input: (1)  $\bar{\mathbf{G}}_{n-1}$ ;

(2) Initial solution  $\mathbf{u}^{(0)}$ ;

(3) Maximum number of iteration Sfor  $1 \leq s \leq S$ 1)  $\mathbf{z}^{(s)} = \bar{\mathbf{G}}_{n-1}\mathbf{u}^{(s-1)}$ 2)  $m^{(s)} = \arg\max_{z_i^{(s)}} \left| z_i^{(s)} \right|$ 3) if  $1 \leq s \leq 2$   $n^{(s)} = m^{(s)}$ else  $n^{(s)} = \frac{m^{(s)}m^{(s-2)} - \left(m^{(s-1)}\right)^2}{m^{(s)} - 2m^{(s-1)} + m^{(s-2)}}$ end if

4)  $\mathbf{u}^{(s)} = \frac{\mathbf{z}^{(s)}}{n^{(s)}}$ end for

Output: (1) The largest singular value  $\Sigma_1 = n^{(S)}$ (2) The first singular vector  $\mathbf{v}_1 = \frac{\mathbf{u}^{(s)}}{\|\mathbf{u}^{(s)}\|_2}$ 

Algorithm 1: Power iteration algorithm

compute  $\mathbf{v}_1$  as well as the largest singular value  $\Sigma_1$  of  $\bar{\mathbf{G}}_{n-1}$  with low complexity.

More specifically, as shown by the pseudo-code in **Algorithm 1**, the power iteration algorithm starts with an initial solution  $\mathbf{u}^{(0)} \in \mathbb{C}^{M \times 1}$ , which is usually set as  $[1,1,\cdots,1]^T$  without loss of generality [19]. In each iteration, it first computes the auxiliary vector  $\mathbf{z}^{(s)} = \bar{\mathbf{G}}_{n-1}\mathbf{u}^{(s-1)}$  (s is the number of iterations) and then extracts the element of  $\mathbf{z}^{(s)}$  having the largest amplitude as  $m^{(s)}$ . After that,  $\mathbf{u}^{(s)}$  is updated as  $\mathbf{u}^{(s)} = \frac{\mathbf{z}^{(s)}}{m^{(s)}}$  for the next iteration. The power iteration algorithm will stop until the number of iterations reaches the predefined number S. Finally,  $m^{(S)}$  and  $\mathbf{u}^{(S)}/\|\mathbf{u}^{(S)}\|_2$  will be output as the largest singular value  $\Sigma_1$  and the first right singular vector  $\mathbf{v}_1$  of  $\bar{\mathbf{G}}_{n-1}$ , respectively.

According to [19], we know that

$$m^{(s)} = \Sigma_1 \left[ 1 + \mathcal{O}\left( \left( \frac{\Sigma_2}{\Sigma_1} \right)^s \right) \right],$$
 (18)

where  $\Sigma_2$  is the second largest singular value of  $\bar{\mathbf{G}}_{n-1}$ . From (18), we can conclude that  $m^{(s)}$  will converges to  $\Sigma_1$  as long as  $\Sigma_1 \neq \Sigma_2$ . Similarly, when  $\Sigma_1 \neq \Sigma_2$ ,  $\mathbf{u}^{(s)}/\|\mathbf{u}^{(s)}\|_2$  will also converge to  $\mathbf{v}_1$ , i.e.,

$$\lim_{s \to \infty} m^{(s)} = \Sigma_1, \quad \lim_{s \to \infty} \frac{\mathbf{u}^{(s)}}{\left\|\mathbf{u}^{(s)}\right\|_2} = \mathbf{v}_1. \tag{19}$$

Although the power iteration algorithm is convergent, its convergence rate may be slow if  $\Sigma_1 \approx \Sigma_2$  based on (18). To solve this problem, we propose to utilize the Aitken acceleration method [20] to further increase the convergence rate of the power iteration algorithm. Specifically, we can compute

$$\begin{cases}
 n^{(s)} = m^{(s)}, & \text{for } 1 \le s \le 2, \\
 n^{(s)} = \frac{m^{(s)}m^{(s-2)} - (m^{(s-1)})^2}{m^{(s)} - 2m^{(s-1)} + m^{(s-2)}}, & \text{for } 2 < s \le S.
\end{cases}$$
(20)

Then,  $\mathbf{u}^{(s)}$  and  $\Sigma_1$  will be correspondingly changed to  $\mathbf{u}^{(s)} = \frac{\mathbf{z}^{(s)}}{n^{(s)}}$  and  $\Sigma_1 = n^{(S)}$ , respectively.

Next, we will focus on how to reduce the complexity to compute the matrices  $\mathbf{T}_n = \mathbf{I}_N + \frac{\rho}{N\sigma^2}\mathbf{H}\mathbf{P}_n\mathbf{P}_n^H\mathbf{H}^H$ 

and  $\bar{\mathbf{G}}_n = \mathbf{R}\mathbf{H}^H\mathbf{T}_n^{-1}\mathbf{H}\mathbf{R}^H$ , which involve the complicated matrix-to-matrix multiplication and matrix inversion of large size. In particular, with some standard mathematical manipulations, the computation of  $\bar{\mathbf{G}}_n$  can be significantly simplified as shown by the following Proposition 2.

**Proposition 2.** The matrix  $\mathbf{\bar{G}}_n = \mathbf{R}\mathbf{H}^H\mathbf{T}_n^{-1}\mathbf{H}\mathbf{R}^H$ , where  $\mathbf{T}_n = \mathbf{I}_N + \frac{\rho}{N\sigma^2}\mathbf{H}\mathbf{P}_n\mathbf{P}_n^H\mathbf{H}^H$ , can be simplified as

$$\bar{\mathbf{G}}_n \approx \bar{\mathbf{G}}_{n-1} - \frac{\frac{\rho}{N\sigma^2} \Sigma_1^2 \mathbf{v}_1 \mathbf{v}_1^H}{1 + \frac{\rho}{N\sigma^2} \Sigma_1},\tag{21}$$

where  $\Sigma_1$  and  $\mathbf{v}_1$  are the largest singular value and first right singular vector of  $\mathbf{G}_{n-1}$ , respectively.

*Proof:* See Appendix B.

**Proposition 2** implies that we can simply exploit  $\Sigma_1$  and  $\mathbf{v}_1$  that have been obtained by **Algorithm 1** as described above to update  $\bar{\mathbf{G}}_n$ , which only involves one vector-to-vector multiplication instead of the complicated matrix-to-matrix multiplication and matrix inversion. Note that the evaluation of computational complexity will be discussed in detail in Section III-E.

# D. Summary of the proposed SIC-based hybrid precoding

Based on the discussion so far, the pseudo-code of the proposed SIC-based hybrid precoding can be summarized in **Algorithm 2**, which can be explained as follows. The proposed SIC-based hybrid precoding starts by computing the largest singular value  $\Sigma_1$  and first right singular vector  $\mathbf{v}_1$  of  $\mathbf{\bar{G}}_{n-1}$ , which is achieved by Algorithm 1. After that, according to Section III-B, the optimal precoding vector for the nth subantenna array can be obtained by utilizing  $v_1$ . Finally, based on **Proposition 2**,  $\bar{\mathbf{G}}_n$  can be updated with low complexity for the next iteration. This procedure will be executed until the last (Nth) sub-antenna array is considered. Finally, after N iterations, the optimal digital, analog, and hybrid precoding matrix D, A, and P can be obtained, respectively.

Input: 
$$\bar{\mathbf{G}}_0$$
for  $1 \leq n \leq N$ 

1) Compute  $\mathbf{v}_1$  and  $\Sigma_1$  of  $\bar{\mathbf{G}}_{n-1}$  by Algorithm 1

2)  $\bar{\mathbf{a}}_n^{\mathrm{opt}} = \frac{1}{\sqrt{M}} e^{j\mathrm{angle}(\mathbf{v}_1)}$ ,  $d_n^{\mathrm{opt}} = \frac{\|\mathbf{v}_1\|_1}{\sqrt{M}}$ ,
 $\bar{\mathbf{p}}_n^{\mathrm{opt}} = \frac{1}{M} \|\mathbf{v}_1\|_1 e^{j\mathrm{angle}(\mathbf{v}_1)}$  (14)-(16)

3)  $\bar{\mathbf{G}}_n = \bar{\mathbf{G}}_{n-1} - \frac{\frac{\rho}{N\sigma^2} \Sigma_1^2 \mathbf{v}_1 \mathbf{v}_1^H}{1 + \frac{\rho}{N\sigma^2} \Sigma_1}$  (Proposition 2) end for

Output: (1)  $\mathbf{D} = \mathrm{diag}\left\{d_1^{\mathrm{opt}}, \cdots, d_N^{\mathrm{opt}}\right\}$ 
(2)  $\mathbf{A} = \mathrm{diag}\left\{\bar{\mathbf{a}}_1^{\mathrm{opt}}, \cdots, \bar{\mathbf{a}}_N^{\mathrm{opt}}\right\}$ 
(3)  $\mathbf{P} = \mathbf{A}\mathbf{D}$ 

Algorithm 2: SIC-based hybrid precoding

It is worth pointing out that the idea of SIC-based hybrid precoding can be also extended to the combining at the user following the similar logic in [6]. When the number of RF chains at the BS is smaller than that at the user, we first compute the optimal hybrid precoding matrix P according to Algorithm 2, where we assume that the combining matrix Q = I. Then, given the effective channel matrix HP, we

can similarly obtain the optimal hybrid combining matrix Q by referring to **Algorithm 2**, where the input  $\mathbf{G}_0$  and the optimal unconstrained solution  $v_1$  should be correspondingly replaced. Conversely, when the number of RF chains at the BS is larger than that at the user, we can assume P = I and obtain the optimal hybrid combining matrix Q. After that, the optimal precoding matrix P can be acquired given the effective channel matrix QH. Additionally, to further improve the performance, we can combine the above method with the "Ping-pong" algorithm [16], which involves an iteration procedure between the BS and the user, to jointly explore the optimal hybrid precoding and combining matrices pair. Further discussion about hybrid combining will be left for our future work.

#### E. Complexity evaluation

In this subsection, we provide the complexity evaluation of the proposed SIC-based hybrid precoding in terms of the required numbers of complex multiplications and divisions. From Algorithm 2, we can observe that the complexity of SIC-based hybrid precoding comes from the following four parts:

- 1) The first one originates from the computation of  $\bar{\mathbf{G}}_0 = \mathbf{R}\mathbf{H}^H\mathbf{H}\mathbf{R}^H$  according to (11). Note that  $\mathbf{R}$  is a selection matrix and **H** has the size  $K \times NM$ . Therefore, this part involves  $KM^2$  times of multiplications without any division.
- 2) The second one is from executing Algorithm 1. It can be observed that in each iteration we need to compute a matrixto-vector multiplication  $\mathbf{z}^{(s)} = \mathbf{\bar{G}}_{n-1}\mathbf{u}^{(s-1)}$  together with the Aitken acceleration method (20). Therefore, we totally require  $S(M^2+2)-4$  and (2S-2) times of multiplications and divisions, respectively.
- 3) The third one stems from acquiring the optimal solution  $\bar{\mathbf{p}}_n^{\text{opt}}$  in step 2 of **Algorithm 2**. We can find that this part is quite simple, which only needs 2 times of multiplications without any division, since  $\mathbf{v}_1$  has been obtained and  $\frac{1}{\sqrt{M}}$  is a fixed constant.
- 4) The last one comes from the update of  $\bar{\mathbf{G}}_n$ . According to Proposition 2, we know that this part mainly involves a outer product  $\mathbf{v}_1\mathbf{v}_1^H$ . Thus, it requires  $\hat{M}^2$  times of multiplications with only one division.

To sum up, the proposed SIC-based hybrid precoding approximately requires  $M^2(NS+K)$  times of multiplications and 2NS times of divisions. It is worth pointing out that the recently proposed spatially sparse precoding [6] requires  $\mathcal{O}\left(N^4M+N^2L^2+N^2M^2L\right)$  times of multiplications and  $\mathcal{O}(2N^3)$  times of divisions, where L is the number of effective channel paths as defined in (3). Considering the typical mmWave MIMO system with N=8, M=8, K=16,L=3 [6], we can observe that the complexity of SIC-based hybrid precoding is about  $4 \times 10^3$  times of multiplications and  $10^2$  times of divisions, where we set S=5 that is enough to guarantee the performance as will be verified in Section IV. By contrast, the complexity of the spatially sparse precoding is about  $5 \times 10^4$  times of multiplications and  $10^3$  times of divisions. Therefore, the proposed SIC-based hybrid precoding enjoys much lower complexity, which is only about 10% as complex as that of the spatially sparse precoding.

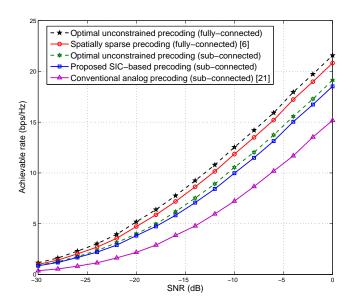


Fig. 3. Achievable rate comparison for an  $NM \times K = 64 \times 16$  (N=8) mmWave MIMO system.

## IV. SIMULATION RESULTS

In this section, we provide the simulation results of the achievable rate to evaluate the performance of the proposed SIC-based hybrid precoding. We compare the performance of SIC-based hybrid precoding with the recently proposed spatially sparse precoding [6] and the optimal unconstrained precoding based on the SVD of the channel matrix, which are both with fully-connected architecture. Additionally, we also include the conventional analog precoding and the optimal unconstrained precoding (i.e.,  $\bar{\mathbf{p}}_n^{\text{opt}} = \mathbf{v}_1$ ) which are both with sub-connected architecture [21] as benchmarks for comparison.

The simulation parameters are described as follows. We generate the channel matrix according to the channel model [17] described in Section II. The number of effective channel paths is L=3. The carrier frequency is set as 28GHz [10]. Both the transmit and receive antenna arrays are ULAs with antenna spacing  $d=\lambda/2$ . Since the BS usually employs the directional antennas to eliminate interference and increase antenna gain [2], the AoDs are assumed to follow the uniform distribution within  $\left[-\frac{\pi}{6},\frac{\pi}{6}\right]$ . Meanwhile, due to the random position of users, we assume that the AoAs follow the uniform distribution within  $\left[-\pi,\pi\right]$ , which means the omnidirectional antennas are adopted by users. Furthermore, we set the maximum number of iterations S=5 to run **Algorithm 2**. Finally, the signal-to-noise ratio (SNR) is defined as  $\frac{\rho}{e^2}$ .

Fig. 3 shows the achievable rate comparison in mmWave MIMO system, where  $NM \times K = 64 \times 16$  and the number of RF chains is N=8. We can observe from Fig. 3 that the proposed SIC-based hybrid precoding outperforms the conventional analog precoding with sub-connected architecture in whole simulated SNR range. Meanwhile, Fig. 3 also verifies the near-optimal performance of SIC-based hybrid precoding, since it can achieve about 99% of the rate achieved by the optimal unconstrained precoding with sub-connected architecture.

Fig. 4 compares the achievable rate in mmWave MIMO

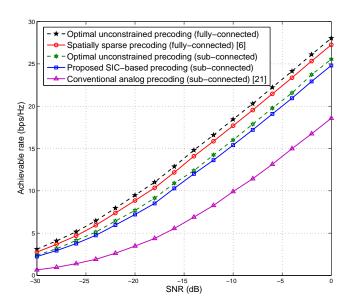


Fig. 4. Achievable rate comparison for an  $NM \times K = 128 \times 32$  (N=16) mmWave MIMO system.

system with  $NM \times K = 128 \times 32$  and N = 16, where we can observe similar trends as those from Fig. 3. More importantly, Fig. 3 and Fig. 4 show that the performance of SIC-based hybrid precoding is also close to the spatially sparse precoding and the optimal unconstrained precoding with fully-connected architecture. For example, when SNR = 0 dB, our method can achieve more than 90% of the rate achieved by the near-optimal spatially sparse precoding in both simulated mmWave MIMO configurations. Considering the low energy consumption and computational complexity of the proposed SIC-based hybrid precoding as analyzed before, we can further conclude that SIC-based hybrid precoding can achieve much better trade-off among the performance, energy consumption, and computational complexity.

Fig. 5 provides a achievable rate comparison in mmWave MIMO systems against the numbers of BS and user antennas, where NM = K and the number of RF chains is fixed to N = 8. We can find that the performance of the proposed SIC-based hybrid precoding can be improved by increasing the number of BS and user antennas, which involves much lower energy consumption than increasing the number of energy-intensive RF chains [13].

Fig. 6 shows the achievable rate comparison against the numbers of user antennas K, where NM=64 and N=8. We can imply from Fig. 6 that the performance loss of SIC-based hybrid precoding due to the sub-connected architecture can be compensated by increasing the number of user antennas K. For example, the achievable rate of SIC-based hybrid precoding when K=30 is the same as that of the spatially sparse precoding when K=20. Note that in this case, the required number of phase shifters of SIC-based hybrid precoding is NM=64 and each RF chain only needs to drive 8 BS antennas, while for the spatially sparse precoding, the number of required phase shifters is  $N^2M=512$  and each RF chain has to drive 64 BS antennas. By contrast, the cost of increasing the number of user antennas K will be negligible since the power consumption of user antenna

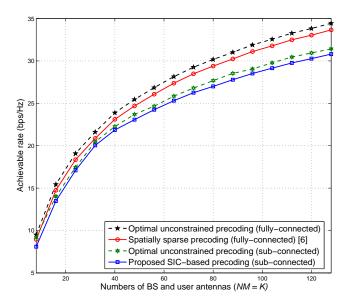


Fig. 5. Achievable rate comparison against the numbers of BS and user antennas (NM = K), where N = 8.

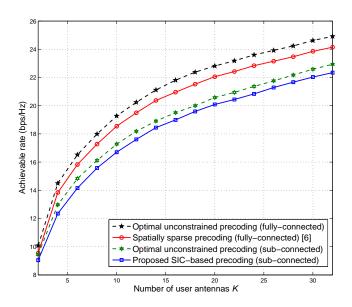


Fig. 6. Achievable rate comparison against the number of user antennas K, where NM=64 and N=8.

is usually small [13]. Therefore, we can conclude that the proposed SIC-based hybrid precoding is more energy-efficient and implementation-practical.

### V. CONCLUSIONS

In this paper, we proposed a SIC-based hybrid precoding with sub-connected architecture for mmWave MIMO systems. We first showed that the total achievable rate optimization problem with non-convex constraints can be decomposed into a series of sub-rate optimization problems, each of which only considers one sub-antenna array. Then, we proved that the sub-rate optimization problem of each sub-antenna array can be solved by simply seeking a precoding vector sufficiently close to the unconstrained optimal solution. Finally, a low-complexity algorithm was proposed to realize SIC-based precoding without the complicated SVD and matrix

inversion. Complexity evaluation showed that the complexity of the proposed SIC-based hybrid precoding is only about 10% as complex as that of the recently proposed spatially sparse precoding with fully-connected architecture in typical mmWave MIMO system. Simulation results verified the near-optimal performance of SIC-based hybrid precoding, and implied that the performance loss induced by sub-connected architecture can be compensated by increasing the number of antennas. This may be a reasonable tradeoff versus increasing the number of phase shifters as required in the fully-connected architecture. Our further work will focus on the limited feedback scenario, where the channel state information may be not perfect and the angles of phase shifters are quantified.

# APPENDIX A PROOF OF PROPOSITION 1

Define the target of the optimization problem (10) as

$$R_n = \log_2 \left( 1 + \frac{\rho}{N\sigma^2} \bar{\mathbf{p}}_n^H \bar{\mathbf{G}}_{n-1} \bar{\mathbf{p}}_n \right), \tag{22}$$

and the SVD of  $\bar{\mathbf{G}}_{n-1}$  as  $\bar{\mathbf{G}}_{n-1} = \mathbf{V} \Sigma \mathbf{V}^H$ . Then, by separating the matrices  $\Sigma$  and  $\mathbf{V}$  into two parts:

$$\Sigma = \begin{bmatrix} \Sigma_1 & 0 \\ 0 & \Sigma_2 \end{bmatrix}, \quad \mathbf{V} = [\mathbf{v}_1 \ \mathbf{V}_2], \tag{23}$$

 $R_n$  in (22) can be rewritten as

$$R_{n} = \log_{2} \left( 1 + \frac{\rho}{N\sigma^{2}} \bar{\mathbf{p}}_{n}^{H} \bar{\mathbf{G}}_{n-1} \bar{\mathbf{p}}_{n} \right)$$

$$= \log_{2} \left( 1 + \frac{\rho}{N\sigma^{2}} \bar{\mathbf{p}}_{n}^{H} \mathbf{V} \mathbf{\Sigma} \mathbf{V}^{H} \bar{\mathbf{p}}_{n} \right)$$

$$= \log_{2} \left( 1 + \frac{\rho}{N\sigma^{2}} \right)$$

$$\times \bar{\mathbf{p}}_{n}^{H} [\mathbf{v}_{1} \mathbf{V}_{2}] \begin{bmatrix} \Sigma_{1} & 0 \\ 0 & \Sigma_{2} \end{bmatrix} [\mathbf{v}_{1} \mathbf{V}_{2}]^{H} \bar{\mathbf{p}}_{n}$$

$$= \log_{2} \left( 1 + \frac{\rho}{N\sigma^{2}} \bar{\mathbf{p}}_{n}^{H} \mathbf{v}_{1} \Sigma_{1} \mathbf{v}_{1}^{H} \bar{\mathbf{p}}_{n} \right)$$

$$+ \frac{\rho}{N\sigma^{2}} \bar{\mathbf{p}}_{n}^{H} \mathbf{V}_{2} \Sigma_{2} \mathbf{V}_{2}^{H} \bar{\mathbf{p}}_{n} \right). \tag{24}$$

Since we aim to find a vector  $\bar{\mathbf{p}}_n$  sufficiently "close" to  $\mathbf{v}_1$ , it is reasonable to assume that  $\bar{\mathbf{p}}_n$  is approximately orthogonal to the matrix  $\mathbf{V}_2$ , i.e.,  $\bar{\mathbf{p}}_n^H \mathbf{V}_2 \approx \mathbf{0}$  [6]. Then, (24) can be simplified as

$$R_{n} \approx \log_{2} \left( 1 + \frac{\rho \Sigma_{1}}{N\sigma^{2}} \bar{\mathbf{p}}_{n}^{H} \mathbf{v}_{1} \mathbf{v}_{1}^{H} \bar{\mathbf{p}}_{n} \right)$$

$$\stackrel{(a)}{=} \log_{2} \left( 1 + \frac{\rho \Sigma_{1}}{N\sigma^{2}} \right)$$

$$+ \log_{2} \left( 1 - \left( 1 + \frac{\rho \Sigma_{1}}{N\sigma^{2}} \right)^{-1} \frac{\rho \Sigma_{1}}{N\sigma^{2}} \left( 1 - \bar{\mathbf{p}}_{n}^{H} \mathbf{v}_{1} \mathbf{v}_{1}^{H} \bar{\mathbf{p}}_{n} \right) \right)$$

$$\stackrel{(b)}{\approx} \log_{2} \left( 1 + \frac{\rho \Sigma_{1}}{N\sigma^{2}} \right) + \log_{2} \left( \bar{\mathbf{p}}_{n}^{H} \mathbf{v}_{1} \mathbf{v}_{1}^{H} \bar{\mathbf{p}}_{n} \right)$$

$$(25)$$

where (a) is obtained by using the formula  $\mathbf{I} + \mathbf{X}\mathbf{Y} = (\mathbf{I} + \mathbf{X}) \left( \mathbf{I} - (\mathbf{I} + \mathbf{X})^{-1} \mathbf{X} (\mathbf{I} - \mathbf{Y}) \right)$  [6], where

we define  $\mathbf{X} = \frac{\rho \Sigma_1}{N\sigma^2}$  and  $\mathbf{Y} = \mathbf{\bar{p}}_n^H \mathbf{v}_1 \mathbf{v}_1^H \mathbf{\bar{p}}_n$ ; (b) is valid by employing the high SNR approximation [22], i.e.,

$$\left(1 + \frac{\rho \Sigma_1}{N\sigma^2}\right)^{-1} \frac{\rho \Sigma_1}{N\sigma^2} \approx 1.$$
(26)

From (25), we can observe that maximizing  $R_n$  is equivalent to maximizing  $\bar{\mathbf{p}}_n^H \mathbf{v}_1 \mathbf{v}_1^H \bar{\mathbf{p}}_n = \|\bar{\mathbf{p}}_n^H \mathbf{v}_1\|_2^2$ , the square of inner product between two vectors  $\bar{\mathbf{p}}_n$  and  $\mathbf{v}_1$ . Note that  $\mathbf{v}_1$  is a fixed vector. Therefore, exploring a vector  $\bar{\mathbf{p}}_n$ , which has the largest projection on  $\mathbf{v}_1$ , will lead to the smallest Euclidean distance to  $\mathbf{v}_1$  as well. Based on this fact, we can conclude that the optimization problem (10) is equivalent to the following problem

$$\bar{\mathbf{p}}_{n}^{\text{opt}} = \underset{\bar{\mathbf{p}}_{n} \in \bar{\mathcal{F}}}{\operatorname{arg\,min}} \left\| \mathbf{v}_{1} - \bar{\mathbf{p}}_{n} \right\|_{2}^{2}. \tag{27}$$

# APPENDIX B PROOF OF PROPOSITION 2

We first consider the matrix  $\mathbf{T}_n = \mathbf{I}_N + \frac{\rho}{N\sigma^2}\mathbf{H}\mathbf{P}_n\mathbf{P}_n^H\mathbf{H}^H$ , which should be inversed to compute  $\bar{\mathbf{G}}_n$  (11). By partitioning  $\mathbf{P}_n$  as  $\mathbf{P}_n = [\mathbf{P}_{n-1}\ \mathbf{p}_n]$ ,  $\mathbf{T}_n$  can be rewritten as

$$\mathbf{T}_{n} = \mathbf{I}_{N} + \frac{\rho}{N\sigma^{2}} \mathbf{H} \mathbf{P}_{n} \mathbf{P}_{n}^{H} \mathbf{H}^{H}$$

$$= \mathbf{I}_{N} + \frac{\rho}{N\sigma^{2}} \mathbf{H} \left[ \mathbf{P}_{n-1} \ \mathbf{p}_{n} \right] \left[ \mathbf{P}_{n-1} \ \mathbf{p}_{n} \right]^{H} \mathbf{H}^{H}$$

$$= \mathbf{I}_{N} + \frac{\rho}{N\sigma^{2}} \mathbf{H} \mathbf{P}_{n-1} \mathbf{P}_{n-1}^{H} \mathbf{H}^{H} + \frac{\rho}{N\sigma^{2}} \mathbf{H} \mathbf{p}_{n} \mathbf{p}_{n}^{H} \mathbf{H}^{H}$$

$$= \mathbf{T}_{n-1} + \frac{\rho}{N\sigma^{2}} \mathbf{H} \mathbf{p}_{n} \mathbf{p}_{n}^{H} \mathbf{H}^{H}. \tag{28}$$

Then, by utilizing the Sherman-Morrison formula [19, Eq 2.1.4]

$$\left(\mathbf{A} + \mathbf{u}\mathbf{v}^{T}\right)^{-1} = \mathbf{A}^{-1} - \frac{\mathbf{A}^{-1}\mathbf{u}\mathbf{v}^{T}\mathbf{A}^{-1}}{1 + \mathbf{v}^{T}\mathbf{A}^{-1}\mathbf{u}},\tag{29}$$

 $\mathbf{T}_n^{-1}$  can be presented as

$$\mathbf{T}_{n}^{-1} = \left(\mathbf{T}_{n-1} + \frac{\rho}{N\sigma^{2}}\mathbf{H}\mathbf{p}_{n}\mathbf{p}_{n}^{H}\mathbf{H}^{H}\right)^{-1}$$

$$= \mathbf{T}_{n-1}^{-1} - \frac{\frac{\rho}{N\sigma^{2}}\mathbf{T}_{n-1}^{-1}\mathbf{H}\mathbf{p}_{n}\mathbf{p}_{n}^{H}\mathbf{H}^{H}\mathbf{T}_{n-1}^{-1}}{1 + \frac{\rho}{N\sigma^{2}}\mathbf{p}^{H}\mathbf{H}^{H}\mathbf{T}_{n-1}^{-1}\mathbf{H}\mathbf{p}_{n}}.$$
(30)

Substituting (30) into  $\mathbf{G}_n = \mathbf{H}^H \mathbf{T}_n^{-1} \mathbf{H}$ , we have

$$\mathbf{G}_{n} = \mathbf{H}^{H} \mathbf{T}_{n}^{-1} \mathbf{H}$$

$$= \mathbf{H}^{H} \left( \mathbf{T}_{n-1}^{-1} - \frac{\frac{\rho}{\sigma^{2}} \mathbf{T}_{n-1}^{-1} \mathbf{H} \mathbf{p}_{n} \mathbf{p}_{n}^{H} \mathbf{H}^{H} \mathbf{T}_{n-1}^{-1}}{1 + \frac{\rho}{\sigma^{2}} \mathbf{p}_{n}^{H} \mathbf{H}^{H} \mathbf{T}_{n-1}^{-1} \mathbf{H} \mathbf{p}_{n}} \right) \mathbf{H}$$

$$= \mathbf{G}_{n-1} - \frac{\frac{\rho}{\sigma^{2}} \mathbf{G}_{n-1} \mathbf{p}_{n} \mathbf{p}_{n}^{H} \mathbf{G}_{n-1}}{1 + \frac{\rho}{\sigma^{2}} \mathbf{p}_{n}^{H} \mathbf{G}_{n-1} \mathbf{p}_{n}}.$$
(31)

Then, according to (11),  $\bar{\mathbf{G}}_n$  can be obtained by

$$\bar{\mathbf{G}}_{n} = \mathbf{R}\mathbf{G}_{n}\mathbf{R}^{H}$$

$$= \mathbf{R}\left(\mathbf{G}_{n-1} - \frac{\frac{\rho}{N\sigma^{2}}\mathbf{G}_{n-1}\mathbf{p}_{n}\mathbf{p}_{n}^{H}\mathbf{G}_{n-1}}{1 + \frac{\rho}{N\sigma^{2}}\mathbf{p}_{n}^{H}\mathbf{G}_{n-1}\mathbf{p}_{n}}\right)\mathbf{R}^{H}$$

$$= \bar{\mathbf{G}}_{n-1} - \frac{\frac{\rho}{N\sigma^{2}}\bar{\mathbf{G}}_{n-1}\bar{\mathbf{p}}_{n}\bar{\mathbf{p}}_{n}^{H}\bar{\mathbf{G}}_{n-1}}{1 + \frac{\rho}{N\sigma^{2}}\bar{\mathbf{p}}_{n}^{H}\bar{\mathbf{G}}_{n-1}\bar{\mathbf{p}}_{n}}.$$
(32)

Note that in Section III-B, we have obtained the optimal solution  $\bar{\mathbf{p}}_n^{\text{opt}}$  which is sufficiently close to  $\mathbf{v}_1$ . Thus, (32) can be well approximated by replacing  $\bar{\mathbf{p}}_n$  with  $\mathbf{v}_1$  as

$$\bar{\mathbf{G}}_{n} = \bar{\mathbf{G}}_{n-1} - \frac{\frac{\rho}{N\sigma^{2}}\bar{\mathbf{G}}_{n-1}\bar{\mathbf{p}}_{n}\bar{\mathbf{p}}_{n}^{H}\bar{\mathbf{G}}_{n-1}}{1 + \frac{\rho}{N\sigma^{2}}\bar{\mathbf{p}}_{n}^{H}\bar{\mathbf{G}}_{n-1}\bar{\mathbf{p}}_{n}}$$

$$\approx \bar{\mathbf{G}}_{n-1} - \frac{\frac{\rho}{N\sigma^{2}}\bar{\mathbf{G}}_{n-1}\mathbf{v}_{1}\mathbf{v}_{1}^{H}\bar{\mathbf{G}}_{n-1}}{1 + \frac{\rho}{N\sigma^{2}}\mathbf{v}_{1}^{H}\bar{\mathbf{G}}_{n-1}\mathbf{v}_{1}}$$

$$\stackrel{(a)}{=} \bar{\mathbf{G}}_{n-1} - \frac{\frac{\rho}{N\sigma^{2}}\Sigma_{1}^{2}\mathbf{v}_{1}\mathbf{v}_{1}^{H}}{1 + \frac{\rho}{N\sigma^{2}}\Sigma_{1}}, \tag{33}$$

where (a) is true due to fact that  $\mathbf{v}_1^H \mathbf{\bar{G}}_{n-1} = \Sigma_1 \mathbf{v}_1^H$ , since  $\mathbf{\bar{G}}_{n-1}$  is an Hermitian matrix.

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