Efficient Joint Hybrid Precoding And Analog Combining Scheme For Massive MIMO Systems

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***Abstract*—Hybrid precoding plays an important role in massive MIMO systems for reducing the hardware cost caused by radio frequency (RF) chains. In this paper, an efficient joint hybrid precoding and analog combining (EJHPAC) scheme is proposed for massive MIMO with multiple-antenna user equipment (UE), which applies the phase elimination method to harvest the power gain. Specifically, the problem of analog combining is transformed to a least square problem with constant modulus constraint. Based on it, we adopt the gradient descent projection (GDP) method to the analog combiner and jointly design the related hybrid precoding algorithm, which leads to the proposed EJHPAC algorithm. According to complexity analysis and simulation results, we show that the EJHPAC algorithm has advantages in both spectral efficiency and computational complexity for massive MIMO systems.**

***Index Terms*—Massive MIMO, hybrid precoding, analog combining, gradient descent projection**

I. Introduction

With the development of 5G and B5G technologies, it is promising to extend the spectrum to the millimeter wave (mmWave) band based on massive MIMO [1], [2]. However, in massive MIMO systems, the traditional digital precoding becomes impractical due to the high hardware cost on radio frequency chains [3]. To solve this problem, hybrid precoding algorithm has emerged as an important topic in massive MIMO [4]. To this end, a low-complexity hybrid precoding scheme named as phased-ZF (PZF) algorithm has been proposed in [5], which approaches the performance of fully ZF precoding with a small number of RF chains. Nevertheless, as [6] points out, the existing hybrid precoding algorithms only consider the massive MIMO systems with single-antenna UE, which severely limits their applications in practice. Therefore, hybrid precoding scheme with analog combining has been proposed for various scenarios of interest in massive MIMO [7].

Specifically, the hybrid block diagonalization (Hy-BD) in [8] is designed for massive MIMO with multiple-antenna UE. Unfortunately, as a heuristic method, its analog combining method cannot guarantee the optimality of sum-rate maximization. In this paper, to improve the spectral efficiency, an efficient joint hybrid precoding and analog combining scheme named as EJHPAC is proposed for massive MIMO with multiple-antenna UE. First of all, the theoretical spectral efficiency of hybrid precoding and analog combining is derived. Then, with respect to the analog combining, the original non-convex problem is transformed to a least square optimization with modulus constraint, where GDP method is applied for solving it. This leads to the proposed EJHPAC algorithm while its low complexity cost is also given by computational analysis. Finally, simulation results show that the proposed EJHPAC algorithm achieves competitive performance in massive MIMO systems.

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II. System Model

As shown in Fig. 1, we consider a multi-user massive MIMO downlink system. The base station (BS) is equipped with N_t transmit antennas and K RF chains to transmit K data streams to K users. Different from the work in [5] that only takes the single-antenna UE into account, here, each user we considered is equipped with $N_r > 1$ receive antennas and one RF chain.

Basically, the hybrid precoding is divided into digital baseband precoder and analog RF precoder, respectively denoted by $\mathbf{D} \in \mathbb{C}^{K\times K}$ and $\mathbf{F} \in \mathbb{C}^{N_t\times K}$. Besides, each multi-antenna receiver needs one analog combiner, denoted by $\mathbf{w}_k \in \mathbb{C}^{N_r\times 1}$ of the k-th user. For the digital precoder $\mathbf{D}$, both the phase and amplitude are adjustable. For the analog precoder $\mathbf{F}$ or the analog combiner $\mathbf{w}_k$ of the k-th user, only the phase changes can be made while the amplitude of each element remains constant. Therefore, each element in $\mathbf{F}$ and $\mathbf{w}_k$ has constant modulus constraint, and is normalized such that $|\mathbf{F}(l,k)| = \frac{1}{N_t}$ and $|\mathbf{w}_k(i)| = \frac{1}{N_r}$, where $|\cdot|$ denotes the magnitude. Furthermore, to meet the total transmit power constraint, $\mathbf{D}$ should be normalized to satisfy $||\mathbf{DF}||_F^2 = K$ [5]. Typically, the signal transmitted from BS antennas can be written as:

$$\mathbf{x} = \mathbf{FDs}, \tag{1}$$

where $\mathbf{s} = [s_1, s_2, \cdots, s_K]^H \in \mathbb{C}^{K\times 1}$ is the signal vector for the K users. Here, the k-th element s_k denotes the scalar symbol transmitted to the k-th user. Note that $\mathbb{E}[\mathbf{ss}^H] = \frac{P}{K}\mathbf{I}_K$ is always satisfied, $\mathbb{E}[\cdot]$ denotes the expectation and P is the transmit power at the BS. On the other hand, the received signal of the k-th multi-antenna user can be expressed as:

$$\begin{aligned}\mathbf{y}_k &= \mathbf{w}_k^H(\mathbf{H}_k\mathbf{FDs} + \mathbf{n}_k)\\ &= \mathbf{w}_k^H\mathbf{H}_k\mathbf{FD}s_k + \sum_{j\neq k}\mathbf{w}_k^H\mathbf{H}_k\mathbf{FD}s_j + \mathbf{w}_k^H\mathbf{n}_k.\end{aligned} \tag{2}$$

Here, $\mathbf{H}_k \in \mathbb{C}^{N_r\times N_t}$ is the channel matrix from the BS to the k-th user, $\mathbf{n}_k$ denotes complex additive white Gaussian noise (AWGN) with zero mean and unit variance.

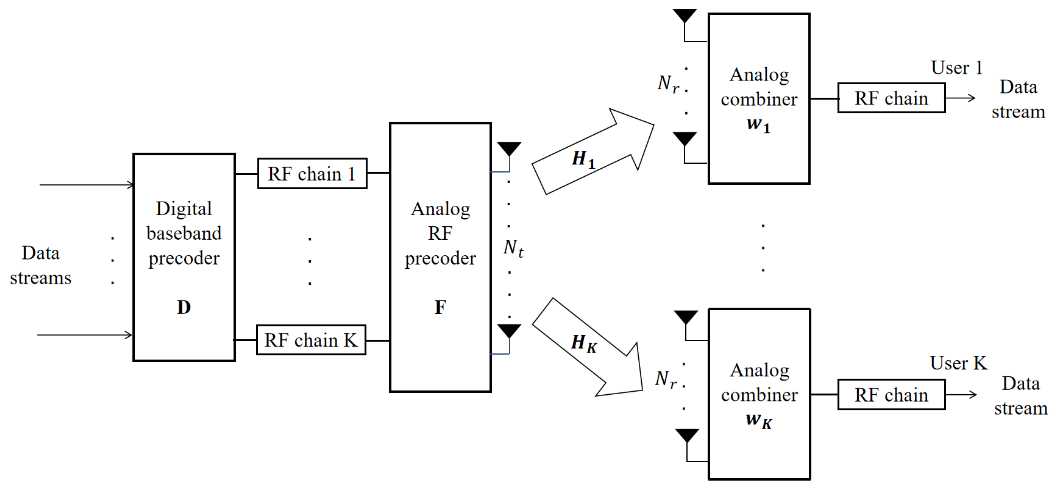

Fig. 1. System diagram of massive MIMO downlink with hybrid precoding and analog combining.

According to (2), the received signal-to-interference-plus-noise ratio (SINR) of the k-th user is given by:

$$\mathrm{SINR}_k = \frac{\frac{P}{K}|\mathbf{w}_k^H \mathbf{H}_k \mathbf{F} \mathbf{d}^k|^2}{1+\sum_{j\neq k}\frac{P}{K}|\mathbf{w}_k^H \mathbf{H}_k \mathbf{F} \mathbf{d}^j|^2}, \tag{3}$$

where $\mathbf{d}^k$ is the k-th column of $\mathbf{D}$. Given Gaussian inputs, the sum spectral efficiency turns out to be:

$$R = \sum_{k=1}^{K} \mathbb{E}\left[\log_2\left(1+\mathrm{SINR}_k\right)\right]. \tag{4}$$

Therefore, to maximize the sum spectral efficiency R, the goal of hybrid precoding and analog combining is:

$$\max_{\mathbf{w}_k,\mathbf{F},\mathbf{D}} \sum_{k=1}^{K} \mathbb{E}\left[\log_2\left(1+\frac{\frac{P}{K}|\mathbf{w}_k^H \mathbf{H}_k \mathbf{F} \mathbf{d}^k|^2}{1+\sum_{j\neq k}\frac{P}{K}|\mathbf{w}_k^H \mathbf{H}_k \mathbf{F} \mathbf{d}^j|^2}\right)\right]. \tag{5}$$

In particular, as shown in [8], the analog combiner $\mathbf{w}_k$ can be obtained by solving:

$$\begin{aligned} &\max_{\mathbf{w}_k} \sum_{k=1}^{K} \left\|\mathbf{w}_k^H \mathbf{H}_k\right\|_1^2, \\ &s.t. \left|\mathbf{w}_k^H(i)\right| = \frac{1}{\sqrt{N_r}}, \forall i,k. \end{aligned} \tag{6}$$

Then, given the analog combiner $\mathbf{w}_k$, the analog precoder $\mathbf{F}$ is designed to harvest the power gain. Finally, the digital precoder $\mathbf{D}$ is designed to eliminate the inter-user interference. However, the problem of (6) is non-convex, rendering it difficult to solve. Though a heuristic algorithm is proposed in [8], considerable performance loss can not be avoided.

III. Hybrid Precoding and Analog Combining Design

A. Theory Analysis

To start with, regarding to the problem of hybrid precoding and analog combining in (5), we first give its theoretical spectral efficiency shown below.

Theorem 1. *When $N_t \times N_r$ tends to infinity, the theoretical spectral efficiency achieved by the joint hybrid precoding and analog combining in Rayleigh fading massive MIMO systems with multiple-antenna UE is*

$$\lim_{N_t\times N_r\to\infty} R = K\log_2\left(1+\frac{\pi}{4}\frac{PN_tN_r}{K}\right). \tag{7}$$

Proof: Typically, when the inter-user interference is eliminated, the sum spectral efficiency R in (4) can be transformed to:

$$R = \sum_{k=1}^{K} \mathbb{E}\left[\log_2\left(1+\frac{P}{K}|\mathbf{w}_k^H \mathbf{H}_k \mathbf{f}^k|^2\right)\right], \tag{8}$$

where $\mathbf{f}^k$ denotes the k-th column of $\mathbf{F}$. Meanwhile, $\mathbf{w}_k$ and $\mathbf{f}^k$ are designed to eliminating the phases of $\mathbf{H}_k$. Given the optimum of phase eliminations $\mathbf{w}_k^*$ and $\mathbf{f}^{k^*}$, we can obtain the following result:

$$\mathbf{w}_k^{*H}\mathbf{H}_k\mathbf{f}^{k^*} = \frac{1}{\sqrt{N_tN_r}}\sum_{i=1}^{N_r}\sum_{l=1}^{N_t}\left|\mathbf{h}_k^l(i)\right|, \forall i,l, \tag{9}$$

where $\mathbf{h}_k^l$ represents the l-th column of $\mathbf{H}_k$. As for Rayleigh fading channels, each element of $\mathbf{H}_k$ is independent and identically distributed (i.i.d.) complex Gaussian random variable with zero mean and unit variance, so that we conclude $|h_k|$ follows Rayleigh distribution with mean $\frac{\sqrt{\pi}}{2}$ and variance $1-\frac{\pi}{4}$. According to the *central limit theorem*, we have:

$$\mathbf{w}_k^{*H}\mathbf{H}_k\mathbf{f}^{k^*} \sim \mathcal{N}\left(\frac{\sqrt{\pi N_tN_r}}{2}, 1-\frac{\pi}{4}\right). \tag{10}$$

Then the spectral efficiency R in (8) can be derived as:

$$\begin{aligned} R =& K\mathbb{E}\left[\log_2\left(1+\frac{P}{K}\left(x+\frac{\sqrt{\pi N_tN_r}}{2}\right)^2\right)\right] \\ =& K\log_2\left(1+\frac{\pi}{4}\frac{PN_tN_r}{K}\right)+K\mathbb{E}\left[\log_2\left(\frac{1+\frac{P}{K}\left(x+\frac{\sqrt{\pi N_tN_r}}{2}\right)^2}{1+\frac{\pi N_tN_r}{4}\frac{P}{K}}\right)\right], \end{aligned} \tag{11}$$

where $x \sim \mathcal{N}(0, 1-\frac{\pi}{4})$. Intuitively, when $N_t \times N_r$ tends to infinity, the second term is limited to zero, completing the proof. ■

As a comparison, the theoretical spectral efficiency achieved by only hybrid precoding in massive MIMO with single-antenna UE has been given as $K\log_2\left(1+\frac{\pi}{4}\frac{PN_t}{K}\right)$ [5]. Consequently, analog combining scheme designed for multiple-antenna UE not only extends hybrid precoding to more practical cases, but also effectively improves the spectral efficiency by extracting the array gain of multiple receive antennas [9], thus making it more promising for massive MIMO systems. Nevertheless, we claim that the spectral efficiency derived in *Theorem 1* is based on the optimal choices of $\mathbf{w}_k$ and $\mathbf{f}^k$. Put it in another way, the R in (7) only serves as an upper bound of the spectral efficiency, and how to approach it chiefly lies on the design of the analog combining and the analog precoding by harvesting the power gain from (5). Next, in order to find $\mathbf{w}_k$ in a better way, we propose an analog combining scheme for approximately solving the problem in (6).

B. Analog Combining Scheme

For notational simplify, here we set $\mathbf{q}_k=\mathbf{w}_k^H\mathbf{H}_k\in\mathbb{C}^{1\times N_t}$ for the k-th user. Then, the problem in (6) can be transformed to maximize the modulus of each element in $\mathbf{q}_k$ with $\mathbf{q}_k(l)=\mathbf{w}_k^H\mathbf{h}_k^l, 1\le l\le N_t$. More precisely, $\mathbf{q}_k(l)$ can be further expressed in an exponential form:

$$\mathbf{q}_k(l)=\frac{1}{\sqrt{N_r}}\sum_{i=1}^{N_r}|\mathbf{h}_k^l(i)|e^{j\theta_{i,l}}e^{j\phi_i},\forall l, \tag{12}$$

where $\theta_{i,l}$ and ϕ_i represent the complex angles of $\mathbf{h}_k^l(i)$ and $\mathbf{w}_k^H(i)$.

Clearly, the modulus of $\mathbf{q}_k(l)$ can be maximized if and only if $\theta_{i,l}+\phi_i$ is small enough, which is known as phase elimination. To this end, $\mathbf{w}_k^H(i)$ is designed for the phase adjustment such that $\mathbf{h}_k^l(i)$ can concentrate on real axis for $1\le l\le N_t$. Here, let us denote the optimization target of $\mathbf{q}_k$ as $\mathbf{q}_k^*\in\mathbb{C}^{1\times N_t}$, so that its l-th element is expressed as:

$$\mathbf{q}_k^*(l)=\frac{1}{\sqrt{N_r}}\sum_{i=1}^{N_r}|\mathbf{h}_k^l(i)|,\forall l. \tag{13}$$

In this way, the non-convex problem in (6) can be transformed to a least square problem with modulus constraint as followed:

$$\begin{aligned}&\min_{\mathbf{w}_k}\sum_{k=1}^{K}\left\|\mathbf{q}_k^*-\mathbf{w}_k^H\mathbf{H}_k\right\|_2^2,\\ &s.t.\left|\mathbf{w}_k^H(i)\right|=\frac{1}{\sqrt{N_r}},\forall i,k.\end{aligned} \tag{14}$$

After that, in order to solve the problem in (14), the GDP method is applied, of which the convergence on solving the least square problem with modulus constraint has been demonstrated in [10]. Specifically, given the r-th iteration $\mathbf{w}_k^{H^{(r)}}$, the $(r+1)$-th iteration computes:

$$\begin{aligned}&\zeta^{(r+1)}=\mathbf{w}_k^{H^{(r)}}+\alpha\big(\mathbf{q}_k^*-\mathbf{w}_k^{H^{(r)}}\mathbf{H}_k\big)\mathbf{H}_k^H;\\ &\mathbf{w}_k^{H^{(r+1)}}=\frac{1}{\sqrt{N_r}}e^{j\angle\zeta^{(r+1)}};\\ &r=r+1,\end{aligned} \tag{15}$$

where $\angle(\cdot)$ denotes the complex angle. Moreover, to aviod introducing extra computational complexity, here we set the initial setup as $\mathbf{w}_k^{H^{(0)}}=(\frac{1}{\sqrt{N_r}},\frac{1}{\sqrt{N_r}},...)$ with step size $\alpha=1$.

C. Efficient Joint Hybrid Precoding and Analog Combining Scheme

As shown in (12), one ϕ_i cannot perfectly match different values of $\theta_{i,l}, 1\le l\le N_t$. Therefore, even if the GDP method converges to the Karush-Kuhn-Tucker (KKT) points of (14), the residue of phase elimination may still exist. To this end, we apply the analog precoding by [8] constructing the intermediate channel as:

$$\mathbf{H}_{int}=\begin{bmatrix}\mathbf{w}_1^H\mathbf{H}_1\\ \vdots\\ \mathbf{w}_K^H\mathbf{H}_K\end{bmatrix}\in\mathbb{C}^{K\times N_t}. \tag{16}$$

Based on it, we construct the analog precoding matrix by the phase elimination, that is:

$$\mathbf{F}(l,k)=\frac{1}{\sqrt{N_t}}e^{j\psi_{l,k}},\forall l,k, \tag{17}$$

where $\psi_{l,k}$ is the phase of the (l,k)-th element of the conjugate transpose of $\mathbf{H}_{int}$, namely $\mathbf{f}^k=(\mathbf{w}_k^H\mathbf{H}_k)^H$ with normalized $|\mathbf{F}(l,k)|=\frac{1}{N_t}$. Obviously, the analog precoding matrix is designed jointly with the analog combining vectors to eliminate the phases of the channel matrix.

On the other hand, to obtain the digital precoding matrix $\mathbf{D}$, we define the equivalent channel vetor of the k-th user as:

$$\tilde{\mathbf{h}}_k=\mathbf{w}_k^H\mathbf{H}_k\mathbf{F}. \tag{18}$$

Then, following the work in [11], we construct as follows the complementary channel matrix of the k-th user:

$$\overline{\mathbf{H}}_k=\left[\widetilde{\mathbf{h}}_1^T,\ldots,\widetilde{\mathbf{h}}_{k-1}^T,\widetilde{\mathbf{h}}_{k+1}^T,\ldots,\widetilde{\mathbf{h}}_K^T\right]^T. \tag{19}$$

In principle, to eliminate the inter-user interference, the k-th column of $\mathbf{D}$ should lie in the null space of $\overline{\mathbf{H}}_k$. Therefore, we perform the singular value decomposition (SVD) with respect to the complementary channel:

$$\overline{\mathbf{H}}_k=\overline{\mathbf{U}}_k\overline{\mathbf{\Sigma}}_k\left[\overline{\mathbf{V}}_k^{K-1},\overline{\mathbf{v}}_k\right]^H, \tag{20}$$

where $\overline{\mathbf{v}}_k$ is the last right singular vector. By doing this, $\overline{\mathbf{v}}_k$ is the null space of $\overline{\mathbf{H}}_k$ with the following relation:

$$\widetilde{\mathbf{h}}_j\overline{\mathbf{v}}_k=\begin{cases}0, & j\neq k,\\ \widetilde{\mathbf{h}}_i\overline{\mathbf{v}}_k, & j=k,\end{cases} \tag{21}$$

so that we can construct the k-th column of $\mathbf{D}$ as:

$$\mathbf{d}^k=\overline{\mathbf{v}}_k. \tag{22}$$

To summarize, the proposed EJHPAC algorithm is shown as *Algorithm 1*.

IV. COMPLEXITY ANALYSIS

In this section, we analyze the computational complexity of the proposed EJHPAC algorithm and then compare it to Hy-BD in [8] and joint scheme in [11].

Algorithm 1 Efficient Joint Hybrid Precoding and Analog Combining (EJHPAC) algorithm

Input: $\mathbf{w}_k^{H^{(0)}}, \alpha, r, \mathbf{H}_1, \mathbf{H}_2, \ \ldots, \mathbf{H}_K$
Output: $\mathbf{F}, \mathbf{D}, \mathbf{w}_k, \forall k \in (1, 2, \ \ldots, K)$

for $k = 1, 2, \ \ldots, K$ **do**
  construct $\mathbf{q}_k^*$ by (13)
  for $r = 0, 1, 2, \ \ldots$ **do**
    get $\mathbf{w}_k^{H^{(r+1)}}$ by (15)
  end for
  construct the k-th column of $\mathbf{F}$ as $\mathbf{f}^k = (\mathbf{w}_k^H \mathbf{H}_k)^H$
end for
normalize $\mathbf{F}$ so that $|\mathbf{F}(l, k)| = \frac{1}{N_t}$
for k =1,2, …,K **do**
  construct the k-th column of $\mathbf{D}$ according to (18)-(22)
end for
normalize $\mathbf{D}$ so that $||\mathbf{DF}||_F^2 = K$

On one hand, the computation of $\mathbf{w}_k$ consists of two parts: computing $\mathbf{q}_k^*$ in (13) and the gradient in (15) at each iteration. Specifically, the computational complexity of $\mathbf{q}_k^*$ is $\mathcal{O}(N_t N_r)$. Computing the gradient includes computing $\mathbf{w}_k^{H^{(r)}} \mathbf{H}_k$ and $\left(\mathbf{q}_k^* - \mathbf{w}_k^{H^{(r)}} \mathbf{H}_k\right) \mathbf{H}_k^H$, and its complexity is $\mathcal{O}(2 N_t N_r)$. Given the number of iteration r, the complexity of computing $\mathbf{w}_k$ is $\mathcal{O}\big((2r+1) N_t N_r\big)$. In contrast, the Hy-BD for constructing $\mathbf{w}_k$ computes $||\mathbf{d}(w)^H \mathbf{H}_k||_1$ for N_r times, of which the complexity is $\mathcal{O}(N_t N_r^2)$. Besides, the computation of $\mathbf{F}$ contains computing $\mathbf{w}_k^H \mathbf{H}_k$ in (16), and its complexity is $\mathcal{O}(K N_t N_r)$. On the other hand, the computational complexity of $\mathbf{D}$ mainly comes from computing the equivalent channel in (18) and the SVD operation in (20), which is $\mathcal{O}\big(K[(K+1) N_t N_r + K^2 (K-1)]\big)$.

Thus, the overall complexity of EJHPAC is $\mathcal{O}\big(K[(2r+K+3) N_t N_r + K^2 (K-1)]\big)$. As a comparison, the main complexity of the joint scheme in [11] is due to the application of the SVD, of which the overall complexity is given as $\mathcal{O}\big(K(N_t^2 N_r + 4N_t^2 + N_r^2)\big)$. Consequently, when the number of iteration of EJHPAC is limited, the complexity of EJHPAC is lower than these algorithms. For a better understanding, the complexity comparison among different algorithms is shown in *Table 1*.

V. Simulation Results

In this section, we evaluate the spectral efficiency achieved by EJHPAC and other existing algorithms. We consider array response vector in an uniform planar array (UPA) for mmWave channels in this study. The number of clusters and propagation paths per cluster are set to $N_c = 8$ and $N_p = 10$, respectively [8]. The truncated *Laplacian* distribution is employed to generate the azimuth and elevation angles of arrival and departure, and the angle spreads are equal to 7.5°. The sector angles in the azimuth domains are set to be 120° for the transmit antenna array, while the receive antenna array is omni-directional due to the relatively smaller antenna array elements [8]. The parameter settings of azimuth angles are also applied to elevation angles.

As shown in Fig. 2, for Rayleigh fading channels, the theoretical spectral efficiency with multiple-antenna UE derived in Theorem 1 is much better than that with single-antenna UE in [5] and weighted minimum mean-square error (WMMSE) precoding algorithm [12], which reveals the improved performance introduced by the multiple receive antennas. Note that the simulation results of EJHPAC is close to the derived theoretical spectral efficiency given in (7). Besides, EJHPAC attains slightly better performance than Hy-BD in [8]. Although the performance of fully ZF precoding is slightly better than that of EJHPAC, fully ZF precoding becomes impractical under the limit of the hardware cost on RF chains.

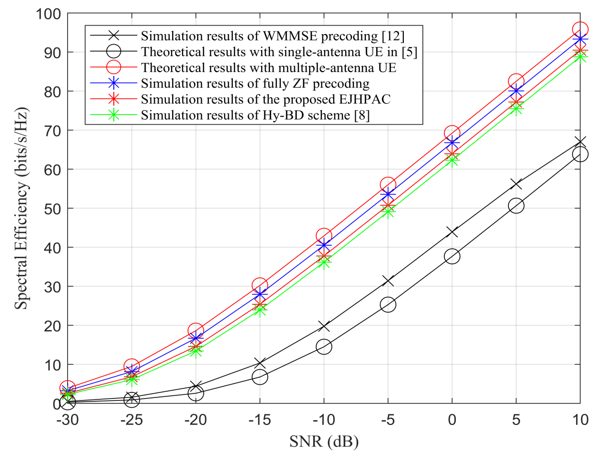

Fig. 2. Spectral efficiency achieved by EJHPAC algorithm and other algorithms with the increasing SNR for 256×16 8-user massive MIMO systems in Rayleigh fading channels.

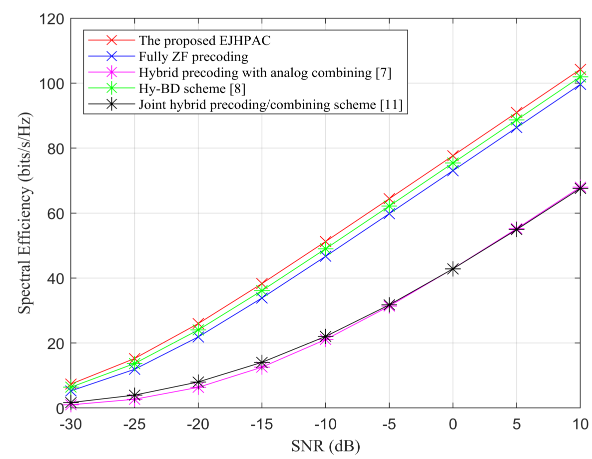

Fig. 3. Spectral efficiency achieved by various precoding schemes with the increasing SNR for 256×16 8-user massive MIMO systems in mmWave channels.

On the other hand, as for mmWave channels, the extended Saleh-Valenzuela channel model in [13] is applied here, i.e.,

$$\mathbf{H} = \sqrt{\frac{N_t N_{\mathrm{r}}}{N_c N_p}} \sum_{i=1}^{N_c} \sum_{j=1}^{N_p} \alpha_{i,j} \boldsymbol{a}_{\mathrm{r}} \left(\phi_{i,j}^r, \theta_{i,j}^r\right) \boldsymbol{a}_t \left(\phi_{i,j}^t, \theta_{i,j}^t\right)^H . \quad (23)$$

Different from the Rayleigh fading channels, mmWave channels can be considered as linear combination of all array response vectors according to (23), which results in the regularity of the phases of channel matrix. For this reason, EJHPAC can effectively eliminate the phases in mmWave channels. Therefore, from Fig. 3 and Fig. 4, we observe that

TABLE I
THE COMPUTATIONAL COMPLEXITIES OF EJHPAC AND OTHER ALGORITHMS

	Overall complexity	64×16 4-user massive MIMO	256×16 8-user massive MIMO
EJHPAC	$\mathcal{O}\big(K[(2r+K+3)N_tN_r+K^2(K-1)]\big)$	$53440(r=3)$	$691712(r=5)$
Hy-BD in [8]	$\mathcal{O}\big(K[(N_r+K+2)N_tN_r+K^2(K-1)]\big)$	90304	855552
Joint scheme in [11]	$\mathcal{O}\big(K(N_t^2N_r+4N_t^2+N_r^2)\big)$	328704	10487808

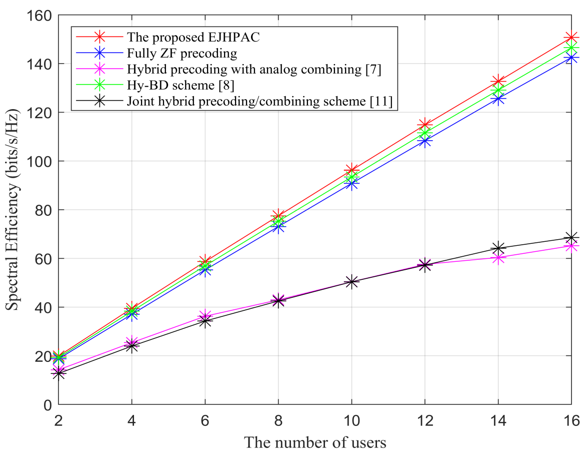

Fig. 4. Spectral efficiency achieved by various precoding schemes with the increasing number of users for 256×16 massive MIMO systems in mmWave channels (SNR = 0dB).

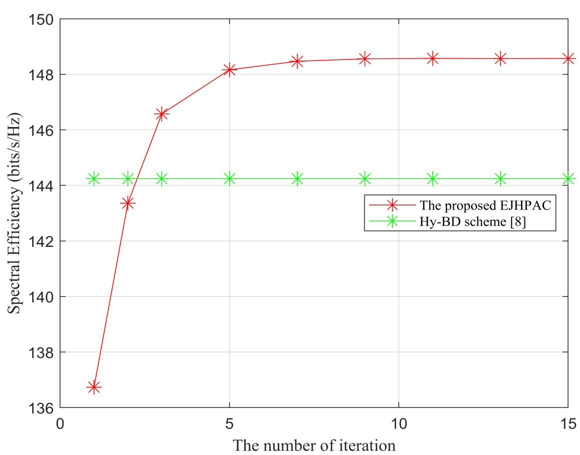

Fig. 5. Spectral efficiency achieved by EJHPAC algorithm with the increasing number of iteration for 256×16 16-user massive MIMO systems in mmWave channels (SNR = 0dB).

EJHPAC achieves better performance than other precoding algorithms, even including fully ZF precoding algorithm. The schemes in [7] and [11] show much lower spectral efficiency because they cannot well obtain larger power gain than the phase elimination in the analog domain. Additionally, it can be seen from Fig. 5 that compared to Hy-BD, the performance of EJHPAC is better while the complexity reduction can be observed from *Table 1*.

VI. CONCLUSION

In this paper we propose an efficient joint hybrid precoding and analog combining scheme named as EJHPAC for massive MIMO systems with multiple-antenna UE. According to theoretical analysis of the spectral efficiency, the problem of analog combining is transformed and the GDP method is chosen to solve it. The analog precoding is jointly designed by phase elimination to achieve the power gain while the digital precoding is designed to eliminate interference. The complexity analysis indicates that EJHPAC is efficient as well. Finally, simulation results confirm that EJHPAC attains better performance than other algorithms.

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