Wideband Hybrid Precoder Design in MU-MIMO based on Channel Angular Information

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***Abstract*—Millimeter wave (mmWave) is one of the key technology in the fifth generation mobile communication system. In mmWave system, hybrid precoding is applied to reduce the number of radio frequency (RF) chains. Although narrow band multiuser hybrid precoding is widely studied, only a few works focus on wideband hybrid precoding. The main problem is that all subcarrier share the same analog precoding matrix but the channel information in frequency domain are different. In this paper, we propose a wideband hybrid precoding method based on channel angular information in multiuser-MIMO (MU-MIMO) scenario. Since the channel state information (CSI) in time domain are the same for all subcarriers, we first construct a autocorrelation channel matrix by angular information, which is available for all subcarriers. Then, we propose an iterative method using manifold optimization for analog precoders and combiners to achieve higher sum data rate. Simulation results show that our proposed method is superior to the existing hybrid precoding method and has a close performance to digital ZF precoding.**

***Index Terms*—Hybrid precoding, wideband, autocorrelation channel matrix, angular information, manifold optimization**

I. Introduction

Millimeter wave (mmWave) is one of the key technology in the fifth generation mobile communication system to provide higher throughput. Compare with the frequency band used in LTE, mmWave has much wider spectral range and less interference source, so more bandwidth is allowed for users. Besides, shorter wavelength of mmWave could reduce the size of antenna, which is helpful for the deployment of large-scale antenna array in the base station (BS) and user equipments (UEs) [1].

However, there are some challenges in mmWave communication. Because of the high pathloss, mmWave has poor scattering and diffraction abilities [2]. Furthermore, in full digital precoding scheme, each antenna connects to a radio frequency (RF) chain, while this structure is hard to achieve since large number of RF chains in massive-MIMO system will induce a great deal of power consumption and high cost. To deal with this problem, hybrid precoding structure is introduced, where the analog precoder conducts phase-only precoding with phase shifters and the baseband adjusts both phase and amplitude.

Existing researches have come up with some hybrid precoding methods for different scenarios. For example, in single-user MIMO (SU-MIMO) system, the main idea is optimizing the analog and digital precoding matrix to approximate the optimal precoding matrix in full digital precoding [3], [4]. [5] propose a alternating minimization approach based on manifold optimization which could approach the performance of digital precoding. But this method has high computation complexity due to the complex gradient function in each iteration. For MU-MIMO system in [6], [7], [8], analog precoder is used to improve the desired signal power and the digital precoder could suppress the interferences between users.

These works could effectively improve the sum data rate in single carrier system, but not suitable for OFDM system. Because in OFDM system, the channel state information (CSI) for all subcarriers are different but the analog precoding matrix is the same. [9] proposes a spatial covariance matrix (SCM) based hybrid precoding method in MU-MISO-OFDM system, where the SCM is constructed by the statistical channel information. This matrix is available for all subcarriers and it is used to calculate analog precoding matrix. But this method relies on sufficient RF chains, it could not perform well when the number of RF chains is nearly equal to the number of users because the analog precoding could not ensure the desired signal power of each user. Besides, although the change of SCM is much slower than instantaneous CSI, frequency domain channel for all subcarriers are still required to calculate the SCM, which will lead to high complexity of channel estimation .

In this paper, we propose a hybrid precoding method for MU-MIMO-OFDM system, where multiple antennas is considered in receivers. In mmWave system, with the sparse nature of mmWave channel, limited rays and clusters arrive at receivers. So it is much easier to obtain channel angular information than perfect CSI in all subcarriers, since the frequency domain channel in all subcarriers share the same angular information and complex gain in each ray. With the existing angle estimation method [10], angular information could be acquired by fast beamforming training method without much complexity. On this basis, analog beamforming in the BS and UEs are designed depends on the angle of departure (AoD) and angle of arrival (AoA).

In details, we will first analysis the relevance of the CSI in frequency domain among all subcarriers and construct the autocorrelation channel matrix which is available for all subcarriers by angular information. Then, we propose a hybrid precoding and analog combining method using the steepest descend algorithm based on manifold optimization. Finally, simulation results show that our proposed method is superior to the SCM based precoding and has a close performance to digital ZF precoding.

II. SYSTEM MODEL

A. System Model

A single cell MU-MIMO-OFDM mmWave system is shown in Fig. 1, where a base station (BS) transmit N_s data streams to K user equipments (UEs) by N_{RF} RF chains and N_t antennas ($N_s \leq N_{\text{RF}} << N_t$). Each UE has N_r antennas and one RF chain, so the BS could only transmit one data stream for each UE ($N_s = K$) and only analog combiner is allowed in receiver. For OFDM system, we consider all UEs use the same L subcarriers. Data streams in the l-th subcarrier are first precoded by different baseband digital precoding matrix $\mathbf{F}_{\text{BB}}(l) \in \mathbb{C}^{N_{\text{RF}} \times N_s}$. Then the frequency domain data are transformed to the time domain by inverse fast fourier transformation (IFFT). In the end, data in all subcarriers are precoded by the same analog precoder $\mathbf{F}_{\text{RF}} \in \mathbb{C}^{N_t \times N_{\text{RF}}}$, since analog precoder is achieved by phase shifters, so the elements in $\mathbf{F}_{\text{RF}}$ have the same modulo value, satisfying $|[\mathbf{F}_{\text{RF}}]_{(m,n)}| = \frac{1}{\sqrt{N_t}}$.

In UE k, the received signal is first combined by beamforming vector $\mathbf{w}_k \in \mathbb{C}^{N_r \times 1}$, satisfying $|[\mathbf{w}_k]_{(m)}| = \frac{1}{\sqrt{N_r}}$. Then the data will be transfered to frequency domain by fast fourier transformation(FFT).

For the downlink data transmission, the received signal in UE k in subcarrier l is represented as

$$y_k(l) = \mathbf{w}_k^H \mathbf{H}_k(l) \mathbf{F}_{\text{RF}} \mathbf{f}_{\text{BB},k}(l) s_k(l) + \sum_{i \neq k} \mathbf{w}_k^H \mathbf{H}_k(l) \mathbf{F}_{\text{RF}} \mathbf{f}_{\text{BB},i}(l) s_i(l) + \mathbf{w}_k^H \mathbf{n}_k(l). \quad (1)$$

For user k in subcarrier l, $\mathbf{H}_k(l) \in \mathbb{C}^{N_r \times N_t}$ denotes the channel matrix in frequency domain, $\mathbf{f}_{\text{BB},k}(l)$ is the digital precoder, $\mathbf{F}_{\text{BB}}(l) = [\mathbf{f}_{\text{BB},1}(l), ..., \mathbf{f}_{\text{BB},K}(l)]$, $s_k(l)$ is the transmission data and $\mathbf{n}_k(l) \sim \mathcal{CN}(0, \sigma_{\text{N}}^2 \mathbf{I})$ is the additive white Gaussian noise vector. In (1), the first term is the expected signal and the second term denotes the inter-user interference.

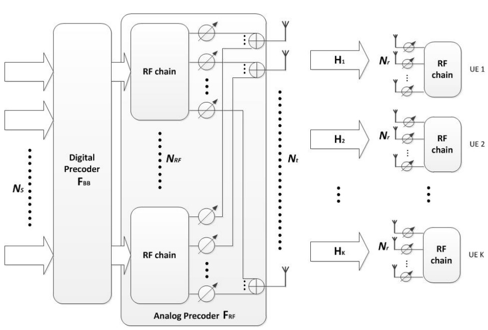

Fig. 1. Fully-connected hybrid precoding architecture.

B. Channel Model

In this paper, we use the 3GPP TR 38.900 channel model for mmWave [11], 2-D channel model is considered for Uniform Linear Array(ULA) in our scenario. In this channel model, a transmitting beam is separated into several clusters and these clusters will be further divided into some rays with various AoD, AoA, complex gain and time delay. The channel response is the superposition of all the impulse response of each ray. So the channel for user k in time domain could be represented by

$$\tilde{\mathbf{H}}_k(t) = \sqrt{\frac{N_t N_r}{N_{cl} N_{ray}}} \sum_{i=1}^{N_{cl}} \sum_{p=1}^{N_{ray}} \alpha_{ipk} \mathbf{a}_{\text{R}}(\theta_{ipk}) \times \mathbf{a}_{\text{T}}^H(\phi_{ipk}) \delta(t - \tau_{ik}), \quad (2)$$

where N_{cl} and N_{ray} denote the number of clusters and the number of rays in a cluster, α_{ipk}, θ_{ipk}, ϕ_{ipk} are the complex gain, AoD and AoA of the p-th ray in the i-th cluster for user k respectively, τ_{ik} is the time delay for cluster i. In channel model [11], all the rays in one cluster share the same time delay. $\mathbf{a}_{\text{T}}(\theta)$ and $\mathbf{a}_{\text{R}}(\phi)$ are the normalized antenna array response vector corresponding to θ and ϕ respectively, which can be written as

$$\mathbf{a}_{\text{T}}(\theta) = \sqrt{\frac{1}{N_t}} [1, e^{j2\pi \frac{d}{\lambda} \sin(\theta)}, ..., e^{j2\pi (N_t - 1) \frac{d}{\lambda} \sin(\theta)}]^{\text{T}}, \quad (3)$$

and

$$\mathbf{a}_{\text{R}}(\phi) = \sqrt{\frac{1}{N_r}} [1, e^{j2\pi \frac{d}{\lambda} \sin(\phi)}, ..., e^{j2\pi (N_r - 1) \frac{d}{\lambda} \sin(\phi)}]^{\text{T}}, \quad (4)$$

with λ being the wavelength and d being the distance between antennas. Typically, we assume $d = \lambda/2$.

After Fourier transformation, we get the frequency domain channel matrix for user k in subcarrier l, which could be represented as

$$\mathbf{H}_k(l) = \sqrt{\frac{N_t N_r}{N_{cl} N_{ray}}} \sum_{i=1}^{N_{cl}} \sum_{p=1}^{N_{ray}} \alpha_{ipk} \mathbf{a}_{\text{R}}(\theta_{ipk}) \times \mathbf{a}_{\text{T}}^H(\phi_{ipk}) e^{-j\tau_{ik}\omega_l}, \quad (5)$$

with ω_l being the angular frequency of subcarrier l.

III. CONSTRUCTION OF AUTOCORRELATION CHANNEL MATRIX BY ANGULAR INFORMATION

Our objective is to find analog and digital precoding method to maximize the sum data rate of all users. Simply, we assume that receivers decode data without inter-carrier interference (ICI). The data rate of user k is the average data rate of all subcarriers.

Generally, in single carrier MU-MIMO system, analog precoding is used to maximize the effective channel gain for each user and digital precoding aims to suppress inter-user interference [6] [8]. But in MU-MIMO-OFDM system, the problem is that we use the same analog precoding matrix to deal with different channel matrix in all subcarriers. In our work, we try to find the common feature of the channels in all subcarriers and search for a new precoding method to maximize the sum data rate.

With the knowledge of angular information and complex gain in each ray, we demonstrate that all subcarriers has nearly the same autocorrelation channel matrix and construct it by AoD and AoA. It is obvious that the difference of the CSI in each subcarrier appears by the time delay factor $e^{-j\tau_{ik}\omega_l}$. With the representation form of frequency domain channel in (5), autocorrelation matrix in receivers for user k in subcarrier l could be calculated by (6). This autocorrelation matrix is constructed by real time channel information instead of statistical information. From (6), autocorrelation channel matrix is the superposition of cross-correlation matrix between the channel response for all rays. For any two rays in one cluster, they have much stronger correlation than two rays in two clusters, because they have similar AoDs and AoAs, also the delay time in one cluster are the same, so the latter could be ignored and we use (7) to simplify (6).

$$\begin{aligned}\mathbf{R}_{\mathrm{r},k} \approx \frac{N_t N_r}{N_{cl} N_{ray}} \sum_{i=1}^{N_{cl}} \sum_{p=1}^{N_{ray}} \sum_{m=1}^{N_{ray}} \alpha_{ipk} \alpha_{imk}^* \mathbf{a}_{\mathrm{R}}(\phi_{ipk}) \\ \times \mathbf{a}_{\mathrm{T}}^H(\theta_{ipk}) \mathbf{a}_{\mathrm{T}}(\theta_{imk}) \mathbf{a}_{\mathrm{R}}^H(\phi_{imk}),\end{aligned} \tag{7}$$

where α_{imk}^* is the conjugate of α_{imk}. In (7), we only need to calculate the cross-correlation matrix between rays in the same cluster. By this approximation, we could reduce the complexity to calculate $\mathbf{R}_{\mathrm{r},k}$. Besides, the effect of time delay is eliminated because we avoid the cross-correlation between rays with various time delay. So all subcarriers could share the same approximate autocorrelation channel matrix calculated by (7).

In the BS, after analog combining in receivers, the effective channel in transmitter for user k is a MISO channel, which could be represented by (8), with the effective complex gain $\alpha_{\mathrm{t},ipk} = \alpha_{ipk} \mathbf{w}_k^H \mathbf{a}_{\mathrm{R}}(\theta_{ipk})$ for each ray in MISO channel. In the same way, we ignore the cross-correlation between rays in 2 clusters and the approximate autocorrelation channel matrix in transmitter for all subcarriers is calculated by (9)

$$\mathbf{R}_{\mathrm{t},k} \approx \sum_{i=1}^{N_{cl}} \sum_{p=1}^{N_{ray}} \sum_{m=1}^{N_{ray}} \alpha_{\mathrm{t},ipk}^* \alpha_{\mathrm{t},imk} \mathbf{a}_{\mathrm{T}}(\theta_{ipk}) \mathbf{a}_{\mathrm{T}}^H(\theta_{imk}). \tag{9}$$

IV. Hybrid Precoding Method

From section III, we get the approximate autocorrelation channel matrix in transmitter and receivers. By $\mathbf{R}_{\mathrm{t},k}$ and $\mathbf{R}_{\mathrm{r},k}$, our problem is the same as single carrier MU-MIMO precoding.

A. Analog Combiner in UE

In the first stage, users will decide its own analog combiners by $\mathbf{R}_{\mathrm{r},k}$. For users with only one RF chain, there is no enough degree of freedom for interference coordination, so the objective is to maximize the received signal power as

$$\begin{aligned}&\max_{\mathbf{w}_k} \ \mathbf{w}_k^H \mathbf{R}_{\mathrm{r},k} \mathbf{w}_k \\ &s.t. \quad |[\mathbf{w}_k]_{(m)}| = \frac{1}{\sqrt{N_r}} \quad (m = 1, 2, \cdots, N_r)\end{aligned} \tag{10}$$

In digital precoding, the optimal solution is the eigenvector corresponding to the largest eigenvalue. But in constant modulus constraints, there is no closed form solution for analog precoding, so we may find the sub-optimal solution by iterative method.

With the constant modulus constraint in (10), $\mathbf{x} = \sqrt{N_r}\mathbf{w}_k$ forms a complex circle manifold $\mathcal{M}^{N_r} = \left\{\mathbf{x} \in \mathbb{C}^{N_{\mathrm{r}}} : |\mathbf{x}_1| = |\mathbf{x}_2| = \cdots = |\mathbf{x}_{N_r}| = 1\right\}$, which is a Riemannian submanifold of $\mathbb{C}^{N_{\mathrm{r}}}$ [12].

All the tangent vectors on manifold $\mathcal{M}^{N_r}$ construct a tangent space, which could be represented as

$$T_{\mathbf{x}}\mathcal{M}^{N_r} = \left\{\mathbf{y} \in \mathbb{C}^{N_{\mathrm{r}}} : \Re\left\{\mathbf{y} \circ \mathbf{x}^*\right\} = \mathbf{0}_{N_r}\right\}, \tag{11}$$

where $\Re[\cdot]$ denotes the real part of a complex matrix and $\circ$ is Hadamard products.

As we know, the objective function increases fastest in the direction of gradient in Euclidean space. This conclusion is also applicable in manifold space. The projection formula of the Euclidean gradient $\nabla f(\mathbf{x})$ onto the tangent space $T_{\mathbf{x}}\mathcal{M}^{N_r}$ is given in [12] as (12)

$$\begin{aligned}\mathbf{d} &= \mathrm{grad} f(\mathbf{x}) = \mathrm{Proj}_{\mathbf{x}} \nabla f(\mathbf{x}) \\ &= \nabla f(\mathbf{x}) - \Re\left\{\mathrm{diag}[\nabla f(\mathbf{x}) \circ \mathbf{x}^*]\right\} \mathbf{x}.\end{aligned} \tag{12}$$

The gradient of our objective function is easy to obtain

$$\nabla f(\mathbf{x}) = 2\mathbf{R}_{\mathrm{r},k}\mathbf{x}. \tag{13}$$

After adding a step in tangent space, the new point is not included in manifold. That's to say $|(\mathbf{x} + \alpha\mathbf{d})_i| \neq 1$. To stay on the manifold, retraction mapping from $T_{\mathbf{x}}\mathcal{M}^{N_r}$ to $\mathcal{M}^{N_r}$ is used to ensure $\mathbf{x} \in \mathcal{M}^{N_r}$.

$$\begin{aligned}&\mathrm{Retr}_{\mathbf{x}} : T_{\mathbf{x}}\mathcal{M}^{N_r} \to \mathcal{M}^{N_r} : \\ &\qquad \alpha\mathbf{d} \mapsto \mathrm{Retr}_{\mathbf{x}}(\alpha\mathbf{d}) = \mathrm{vec}\left[\frac{(\mathbf{x} + \alpha\mathbf{d})_i}{|(\mathbf{x} + \alpha\mathbf{d})_i|}\right].\end{aligned} \tag{14}$$

Based on the above formula, steps for steepest descend algorithm is shown in Table I. The idea is the same as the steepest descend algorithm in Euclidean space, the search direction is the gradient direction in each iteration, so the result could certainly converge to a local optimal solution but not necessarily a global optimal solution. Besides, our proposed method has much lower complexity than [5] because the gradient function in Euclidean space is easy to calculate.

The effect of iterative step is the same as that in steepest descend algorithm. We consider a larger step in the range of $0 < \alpha < 2/\lambda_{\max}$ for a faster convergence speed. In this paper, we choose $\alpha = 1/\lambda_{\max}$ and $N_I = 20$, where $\lambda_{\max}$ is the biggest eigenvalue of $\mathbf{R}_{\mathrm{r},k}$. In this step value, the objective function will converge in 10 to 20 iterations. With this step value, a much larger step may cause divergence

B. Analog Precoder in BS

In the second stage, users feedback $\mathbf{w}_k$ to the BS and then the BS will determine the autocorrelation channel matrix $\mathbf{R}_{\mathrm{t},k}$. In hybrid precoding, although all users share the whole analog precoding matrix, we could regard each column as the precoding vector for each user. Generally, analog precoder is used to improve the signal power and digital precoder aims

$$\mathbf{R}_{\text{r},k}(l) = \mathbf{H}_k(l)\mathbf{H}_k^H(l) = \frac{N_t N_r}{N_{cl}N_{ray}}(\sum_{i=1}^{N_{cl}}\sum_{p=1}^{N_{ray}} \alpha_{ipk}\mathbf{a}_{\text{R}}(\phi_{ipk})\mathbf{a}_{\text{T}}^H(\theta_{ipk})e^{-j\tau_{ik}\omega_l})(\sum_{i=1}^{N_{cl}}\sum_{p=1}^{N_{ray}} \alpha_{ipk}^*\mathbf{a}_{\text{T}}(\theta_{ipk})\mathbf{a}_{\text{R}}^H(\phi_{ipk})e^{j\tau_{ik}\omega_l})$$
$$= \frac{N_t N_r}{N_{cl}N_{ray}}\sum_{i=1}^{N_{cl}}\sum_{p=1}^{N_{cl}}\sum_{m=1}^{N_{ray}}\sum_{n=1}^{N_{ray}} \alpha_{imk}\mathbf{a}_{\text{R}}(\phi_{imk})\mathbf{a}_{\text{T}}^H(\theta_{imk})e^{-j\tau_{ik}\omega_l}\alpha_{pnk}^*\mathbf{a}_{\text{T}}(\theta_{pnk})\mathbf{a}_{\text{R}}^H(\phi_{pnk})e^{j\tau_{pk}\omega_l}, \tag{6}$$

$$\mathbf{h}_{\text{t},k}(l) = \sqrt{\frac{N_t N_r}{N_{cl}N_{ray}}}\sum_{i=1}^{N_{cl}}\sum_{p=1}^{N_{ray}} \alpha_{ipk}\mathbf{w}_k^H\mathbf{a}_{\text{R}}(\theta_{ipk})\mathbf{a}_{\text{T}}^H(\phi_{ipk})e^{-j\tau_{ik}\omega_l} = \sum_{i=1}^{N_{cl}}\sum_{p=1}^{N_{ray}} \alpha_{\text{t},ipk}\mathbf{a}_{\text{T}}^H(\phi_{ipk})e^{-j\tau_{ik}\omega_l}, \tag{8}$$

TABLE I
STEEPEST DESCEND ALGORITHM FOR ANALOG COMBINER DESIGN BASED ON MANIFOLD OPTIMIZATION

Initialize $\mathbf{x}_0 = \mathbf{1}$, select iterative times N_I and step α.
for $i = 1 : N_I$
    Determine Euclidean gradient $\nabla f(\mathbf{x}_{i-1})$.
    Calculate Riemannian gradient $\mathbf{d}_{i-1} = \text{grad}f(\mathbf{x}_{i-1})$.
    Search new point on manifold $\mathbf{x}_i = \text{Retr}_{\mathbf{x}_{i-1}}(\alpha\mathbf{d}_{i-1})$.
End
$\mathbf{w}_k = \mathbf{x}_{N_I}/\sqrt{N_r}$.

to mitigate the interference between users. So the objective function for analog precoder is the same as analog combiners.

$$\begin{aligned} &\max_{\mathbf{f}_{\text{RF},k}} \mathbf{f}_{\text{RF},k}^H\mathbf{R}_{\text{t},k}\mathbf{f}_{\text{RF},k} \\ &\mathbf{F}_{\text{RF}} = [\mathbf{f}_{\text{RF},1}, \mathbf{f}_{\text{RF},2}, \cdots, \mathbf{f}_{\text{RF},K}] \\ &s.t. \quad |[\mathbf{f}_{\text{RF},k}]_{(m)}| = \frac{1}{\sqrt{N_t}} \qquad (m = 1, 2, \cdots, N_t) \end{aligned} \tag{15}$$

Using steepest descend algorithm in Table I, we could also acquire the solution for the analog precoder.

C. Digital Precoder in BS

Finally, as the analog precoder and analog combiners are determined, the BS could obtain the baseband effective instantaneous CSI for all subcarriers, which are easy to estimate because the size of baseband effective channel will be largely reduced after analog precoding. So we assume that baseband channel matrices for all subcarriers are known to the BS. The effective baseband channel matrix for all users in subcarrier l could be represented as (16)

$$\mathbf{H}_{\text{BB}}(l) = \begin{pmatrix} \mathbf{w}_1^H\mathbf{H}_1(l)\mathbf{f}_{\text{RF},1} & \mathbf{w}_1^H\mathbf{H}_1(l)\mathbf{f}_{\text{RF},2} & \cdots & \mathbf{w}_1^H\mathbf{H}_1(l)\mathbf{f}_{\text{RF},K} \\ \mathbf{w}_2^H\mathbf{H}_2(l)\mathbf{f}_{\text{RF},1} & \mathbf{w}_2^H\mathbf{H}_2(l)\mathbf{f}_{\text{RF},2} & \cdots & \mathbf{w}_2^H\mathbf{H}_2(l)\mathbf{f}_{\text{RF},K} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{w}_K^H\mathbf{H}_K(l)\mathbf{f}_{\text{RF},1} & \mathbf{w}_K^H\mathbf{H}_K(l)\mathbf{f}_{\text{RF},2} & \cdots & \mathbf{w}_K^H\mathbf{H}_K(l)\mathbf{f}_{\text{RF},K} \end{pmatrix}. \tag{16}$$

In this diagonally dominant matrix, diagonal elements denote the power of desired signal and the other elements represent the inter-user interferences. Based on this channel matrix, we use ZF precoding for digital precoder, which is difined as

$$\mathbf{F}_{\text{BB}}(l) = \mathbf{H}_{\text{BB}}^H(l)(\mathbf{H}_{\text{BB}}(l)\mathbf{H}_{\text{BB}}^H(l))^{-1}\mathbf{\Lambda}, \tag{17}$$

where $\mathbf{\Lambda}$ is a diagonal matrix to satisfy the power constraint. Without loss of generality, we consider equal power allocation $\|\mathbf{F}_{\text{RF}}\mathbf{f}_{\text{BB},k}\|_{\text{F}} = 1$.

V. SIMULATION RESULTS

In this section, we prove the performance advantages of our proposed method comparing with SCM based hybrid precoding in [9] and ZF precoding in full-digital architecture.

In our scenario, we consider a single cell massive MIMO system where the BS is equipped with 128 antenna ULA and each user with 16 ULA. Consider the operation time of of simulation programs, the BS transmits only 16 subcarriers, but our method is not limited to the number of subcarriers. The center frequency is 28GHz. Channel parameters are given according to 3GPP TR 38.900 [11] in UMa/LOS scene. Setting the position of BS as the coordinate origin, users are distributed in 2D space. In simulation 1, we observe the sum data rate of all users with the change of the received SNR. We have 8 users with the same received SNR and various azimuth angle. Different number of RF chains is considered, $N_{\text{RF}} = K = 8$ and $N_{\text{RF}} = 4K = 32$. In the SCM based precoding scheme, analog combiners in UEs are the same with our proposed method because only single-antenna user is considered in [9]. In full-digital precoding scheme, the analog combining vector in UE is the optimal solution for (10) without constant modulus constrant and ZF precoding is used in the BS by the effective MISO channel in transmitter. In simulation 2, users distribute randomly in the cell. The received signal power is determined by BS transmitting power and large scale fading factor. The noise power is the same for all users, which is defined by the received SNR of 10dB for cell-edge users. The transimitting power is 46dBm and the radius is 200m. With the increasing of users, we also consider $N_{\text{RF}} = K$ for our proposed method, $N_{\text{RF}} = K$ and $N_{\text{RF}} = 4K$ for the SCM based precoding scheme. Results are shown in Fig. 2 and Fig. 3.

From the result, our proposed method has a close performance to digital ZF precoding, the SNR loss is about 2dB. In condition of high SNR, we can achieve more than 95% performance of digital ZF precoding. This point illustrates that the autocorrelation matrix constructed by angular information is close to the real autocorrelation matrix in each subcarrier. With the increasing of user number, sum data rate of our preposed method appears approximate linear growth. The

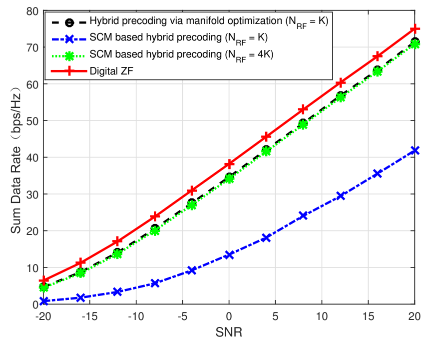

Fig. 2. Sum data rate versus SNR.

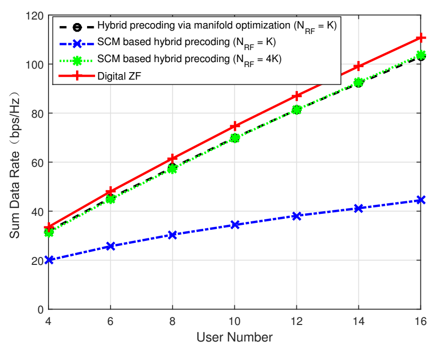

Fig. 3. Sum data rate versus user number.

SCM based precoding scheme has bad performance when $N_{\text{RF}} = K$. With $4K$ RF chains, the sum date rate of SCM based precoding improves greatly but our proposed method has the same performance even if with only K RF chains.

Finally, we observe the effect of angle estimation error to our proposed method. AoD and AoA estimation error is modeled as zero-mean Gaussian distribution and the standard deviation is set to be $\sigma = 0.5^\circ$, $\sigma = 1^\circ$ and $\sigma = 1.5^\circ$. These values are set according to the beamwidth of the BS, which is $180^\circ/N_t \approx 1.4^\circ$. Other simulation parameters are the same with simulation 1.

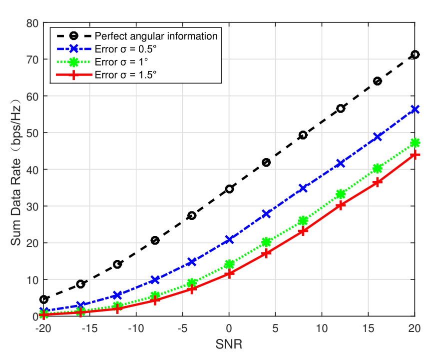

Fig. 4. Sum data rate versus angle estimation error.

As is shown in Fig. 4, with $\sigma = 0.5^\circ$ estimation error, comparing with perfect angular information, the performance degradation is about 8dB. When $\sigma = 1^\circ$ and $\sigma = 1.5^\circ$, the loss is 13dB and 15dB respectively.

VI. Conclusion

In this paper, we propose a hybrid precoding method based on channel angular information in multi-carrier MU-MIMO system. Although the channel information for each subcarrier are different, they share the same AoD and AoA. We demonstrate that all subcarriers have nearly the same autocorrelation channel matrix, which could be constructed by angular information. With these autocorrelation channel matrices in transmitter and receivers, we propose the deepest descend algorithm using manifold optimization for analog precoder and combiners to acquire higher desired signal power. For digital precoder, we choose ZF precoding to mitigate the interference in baseband effective channel. Simulation results show that our proposed method is superior to SCM based precoding and has a close performance to digital ZF with only 2dB SNR loss. Besides, we analyze the effect of angle estimation error and give quantitative results.

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