An Interference Alignment Scheme for 60 GHz Millimeter-wave Communication System

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***Abstract*-60GHz millimeter wave communication systems usually use directional beamforming to compensate high propagation loss in this band. To fully exploit its potential of spatial reuse owing to directivity, an interference alignment scheme based on subarray partition is proposed in this paper. In this scheme, the whole antenna array is partitioned into *M* subarrays, where beamforming is performed within each subarray, and interference alignment is implemented among *M* subarray pairs. Simulation results have shown that the proposed scheme can further mitigate the interference on the basis of inner-subarray beamforming, and significantly improve the sum-capacity of the network in high transmit SNR region.**

***Keywords*-60 GHz, beamforming, interference alignment, subarray partition**

I. INTRODUCTION

Owing to its 5~7GHz wide available unlicensed bandwidth, 60GHz millimeter wave communication has attracted a growing interest from industry and academic fields in the recent years, and a number of standards bodies have begun the related standardization process[1, 2, 3]. The wavelength as short as 5mm in this band makes it possible to fit many antenna elements onto a quite small area to create an antenna array with high directional gain. Therefore, beamforming is usually used to compensate the high path loss and the high oxygen absorption at 60GHz [4]. This directivity feature brings another benefit for the system, i.e. the ability of spatial reuse. Signal power is focused on a particular direction means that receivers at undesired directions may receive less interference, which makes concurrent transmissions in the same neighborhood possible.

In most countries, there are only three non-overlapping channels at 60GHz band, each of them is around 2GHz wide, so that many of the commonly envisioned 60GHz usage scenarios demand heavy exploitation of spatial reuse. For example, a number of use cases defined by IEEE 802.11 Very High Throughput Study Group (VHT SG) involving enterprise and other public places, whose network topology is a dense one, and several links would have to operate in the same channel. However, interference analysis in the typical scenarios has shown that although directivity inherently mitigates interference to some extent, there are still a large number of cases where co-channel interference is non-negligible. In this situation, one or more links cannot operate simultaneously with other links if no effective interference mitigation technique is used. Therefore, interference mitigation has become one of the most important problems in 60GHz system design, and several approaches are proposed, such as null-forming in PHY layer and time scheduling in MAC layer [5].

In this paper, another approach to mitigate interference, i.e. Interference Alignment (IA) will be introduced into 60GHz directional communication system. Interference alignment has attracted much research interest in recent years [6, 8, 9, 10]. Its main principle is to minimize the interference every transmitter causes to unintended receivers instead of maximizing each receiver's own signal-to-interference-plus- noise-ratio (SINR) independently, thus to achieve a better performance on the sum-Capacity of the network. With the IA method, multiple interference signals are consolidated into a small subspace at each receiver so that the desired signal can be transmitted in the interference-free subspace, and every user in the network is able to achieve nearly one half of the capacity that he could achieve in the absence of all interference [6]. The closed form expression for interference alignment needs global channel knowledge which could be an overwhelming overhead in practice [6, 7]. Fortunately, several algorithms with a good performance requiring only local channel knowledge had been found in recent researches [7, 8, 11]. However, the application of IA technique in a 60GHz system with beamforming mechanism has not been investigated so far. What we are wondering is if the same performance improvement can be obtained in such a directional environment where interference has been reduced to some extent, and how to implement IA when the antenna array is needed to ensure adequate beamforming gain as well as perform interference alignment process.

To optimize the sum-capacity of the 60GHz millimeter wave network, an interference alignment scheme based on subarray partition is proposed in this paper. The whole antenna array equipped with N antenna elements is partitioned into M subarrays, where beamforming is performed within each subarray to ensure the beamforming gain need by 60GHz system, and interference alignment is implemented among these M subarray pairs to mitigate the interference. On the other hand, the complexity of the antenna structure could be much reduced with less Analog-Front-End (AFE) required since M is usually much smaller than N. Simulation results have verified the performance advantage of the proposed IA scheme over the pure beamforming scheme in high transmit SNR region.

In what follows, Section II describes the system model and the proposed subarray partition method. The beaforming

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scheme used within each subarray and IA algorithm used among subarray pairs are illustrated in Section III and Section IV respectively. Section V presents the simulation results, and the conclusion is given in Section VI.

II. SYSTEM MODEL

Consider a 60GHz dense office environment: a room size $X \times Y \times Z$ with K transmitter-receiver pairs randomly placed in it, each transmitter equipped with MN_t antennas and each receiver equipped with MN_r antennas for beamforming.

The desired signal at the receiver is a combination of the strong Line-of-Sight (LOS) component and the scattered non-LOS component [12, 13]. In this paper, the Rician channel matrix is expressed as follows

$$\boldsymbol{H} = \sqrt{\frac{k}{k+1}}\boldsymbol{H}_{Los} + \sqrt{\frac{1}{k+1}}\boldsymbol{H}_{nLos}, \tag{1}$$

where $\boldsymbol{H}_{Los}$ is the Los component, $\boldsymbol{H}_{nLos}$ is the non-Los component, k is the Rician factor defined as the power ratio of $\boldsymbol{H}_{Los}$ and $\boldsymbol{H}_{nLos}$. Practical measurement shows that k could range from 6 to 15dB for a typical 60GHz wireless channel [14].

The LOS channel response component $\boldsymbol{H}_{Los}$ is determined by the distance between the transmitter antennas and the receiver antennas, and can be expressed as follows [13, 15]:

$$\boldsymbol{H}_{Los} = \begin{bmatrix} \frac{1}{d_{1,1}} e^{-jkd_{1,1}} & \cdots & \frac{1}{d_{1,MN_t}} e^{-jkd_{1,MN_t}} \\ \vdots & \ddots & \vdots \\ \frac{1}{d_{MN_r,1}} e^{-jkd_{MN_r,1}} & \cdots & \frac{1}{d_{MN_r,MN_t}} e^{-jkd_{MN_r,MN_t}} \end{bmatrix}, \tag{2}$$

where $d_{i,j}$ is the distance between the j-th Tx antenna and the i-th Rx antenna, and $k = 2\pi/\lambda$ is the wave number corresponding to the carrier wavelength λ.

The non-LOS channel response component $\boldsymbol{H}_{nLos}$ can be modeled as an $MN_r * MN_t$ matrix with each element $h_{m,n}$ being an i.i.d. complex Gaussian random variable, i.e.$h_{m,n} \sim C_N(0, \sigma_H^2)$, where σ_H^2 is the variance of the scattered channel response, and is set to 1 here.

With beamforming method which would be introduced in the next section, the multiple-input multiple-output (MIMO) channel $\boldsymbol{H}$ between any transmitter and receiver could be replaced by an equivalent single-input single-output (SISO) channel expressed as

$$\hat{h} = v^H \boldsymbol{H} \mu, \tag{3}$$

In order to utilize IA method for interference suppression which is based on MIMO transmission, and at the same time, keep the benefit of the high directional antenna gain brought by beamforming between the antenna arrays, here we propose an Interference Alignment Scheme in which the antennas at both the transmitter and the receiver are divided into M subarray pairs with an inter-subarray spacing s_t. Beamforming are performed between each subarray pair to create an equivalent $M * M$ MIMO channel between each of the K transmitter and receiver pairs [13]. Take a rank-2 antenna subarray pair for example, the equivalent channel matrix between a transmitter and a receiver can be written as [13]

$$\widehat{\boldsymbol{H}} = \begin{bmatrix} \hat{h}_{11} & \hat{h}_{12} \\ \hat{h}_{21} & \hat{h}_{22} \end{bmatrix} = \begin{bmatrix} v_1^H \boldsymbol{H}_{11} u_1 & v_1^H \boldsymbol{H}_{12} u_2 \\ v_2^H \boldsymbol{H}_{21} u_1 & v_2^H \boldsymbol{H}_{22} u_2 \end{bmatrix}, \tag{4}$$

where $\boldsymbol{H}_{ij}$ is the channel response between the j-th transmitter subarray and the i-th receiver subarray.

III. DAS BEAMFORMING

Beamforming is usually used in antenna array systems to form a particular beam pattern to achieve a directional gain on the desired direction and put nulls on undesired directions by combining the signals from different antenna elements with appropriate weighting.

For a better understanding of the beamforming method, consider a simple N_r-element receiving antenna array, the antennas of which are arranged in a square shape with the adjacent elements separated by half of the wavelength, as illustrated in Figure.1, the situation at the transmitter would be similar. Let's take *ant0* at position (x_0, y_0, z_0) as the reference element. A narrow band plane wave arrives from direction $\omega_r(\theta_r, \gamma_r)$ would have a propagation difference

$$\Delta d_i = (x_i - x_0)\cos\theta\cos\gamma + (y_i - y_0)\sin\theta\cos\gamma + (z_i - z_0)\sin\gamma, \tag{5}$$

and a time difference $\tau_i = \Delta d_i / c$ on its arrival at antI which is placed at position (x_i, y_i, z_i) compared to *ant0*, where θ is the angle between ω_r and x+, and γ is the angle between ω_r and the x-y plane, *c* is the speed of the propagation of the plane wave front.

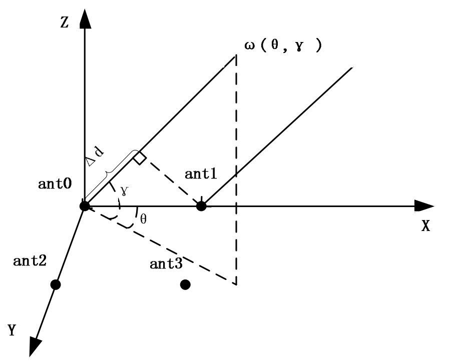

Figure. 1 Desired direction and propagation difference of the antenna array.

The narrow band signal induced on the reference signal could be expressed in complex notation as

$$s_0(t) = m(t)e^{j2\pi f_0 t}, \tag{6}$$

where $m(t)$ is the baseband signal stays almost constant during the time difference due to the narrow band assumption,

f_0 is the carrier frequency. The time difference τ_i results in a phase shifting of φ_i between the signals induced on *ant1* and *ant0*:

$$s_i(t) = m(t)e^{j2\pi f_0(t-\tau_i)} = m(t)e^{j2\pi f_0 t}\, e^{-j2\pi f_0 \tau_i}$$

$$= s_0(t)e^{j\varphi_i}. \quad (7)$$

The receiving signal vector $x(t)$ could be expressed as:

$$\boldsymbol{x}(t) = \boldsymbol{s}(t) + \boldsymbol{n}(t) = s_0(t)\boldsymbol{a} + \boldsymbol{Z}_n(t), \quad (8)$$

where $\boldsymbol{Z}_n(t)$ is the white Gaussian noise vector, $\boldsymbol{a} = [e^{j\varphi_1} \quad e^{j\varphi_2} \quad \ldots \quad e^{j\varphi_{N_r}}]^T$ is the direction vector .

A simple beamforming method called Delay-and-Sum (DAS), which is also the beamforming method used in this paper, is to compensate the phase difference between different antennas by phase shifting or time delay. If we have the signal received by the i-th antenna delayed by $t_i = -\tau_i$ or phase shifted by $\varepsilon_i = -\varphi_i$ [12], the signals received by the antenna array from direction $\omega_r(\theta_r, \gamma_r)$ would be added up completely in phase. In other words, the weighting of the receiver antenna array on direction $\omega_r(\theta_r, \gamma_r)$ could be expressed as a vector $\mu = a_r^H$. The output signal of the antenna array could be expressed as:

$$y(t) = \mu\boldsymbol{x}(t) = N_r s_0(t) + \mu\boldsymbol{Z}_n(t), \quad (9)$$

At the same time, the signals sent by the transmitter antenna array would be added up completely in phase on a particular direction $\omega_t(\theta_t, \gamma_t)$ by a weighting vector $\boldsymbol{v} = a_t^H$. The equivalent channel response can be expressed as follows:

$$\hat{h} = v^H\boldsymbol{H}\mu, \quad (10)$$

The more antennas used in DAS beamforming, the more concentrated to the desired direction the beam pattern is. A receiver antenna array with $\mathrm{N_r}$ antennas could achieve a peak gain of $10\log(N_r)$ dB. Similarly, the transmiter antenna array could get a peak gain of $10\log(N_t)$ dB by beamforming towards the receiver.

IV. INTERFERENCE ALIGNMENT

Consider a K user MIMO interference channel where each user equipped with M_t antennas at the transmitter and M_r antennas at the receiver, the signal received by the n-th receiver can be defined as:

$$\boldsymbol{Y}_n = \textstyle\sum_{m=1}^{K}\boldsymbol{H}_{nm}\boldsymbol{X}_m + \boldsymbol{Z}_n, \quad (11)$$

where $\boldsymbol{Y}_n$ is the $N_r \times 1$ received signal vector at the n-th receiver, $\boldsymbol{H}_{nm}$ is the $M_r \times M_t$ channel response matrix between the m-th transmitter and the n-th receiver. $\boldsymbol{X}_m$ is the $M_t \times 1$ signal vector transmitted by the m-th transmitter, and is under constraint of $E\|\boldsymbol{X}_m\| = P_m$ where P_m is the transmitter power. $\boldsymbol{Z}_n$ is the additive Gaussian noise vector at the n-th receiver with zero mean and unit variance. For the n-th receiver, both the signal and the interference received by the receivers lie in a $\min(M_t, M_r)$ dimensional space.

An interference alignment method is to find transmit pre-coding matrixes $\boldsymbol{V}_m$ and receive interference suppressing matrixes $\boldsymbol{U}_n$ satisfy the following conditions [8]:

$$\boldsymbol{V}_m : N_r \times d, \qquad \boldsymbol{V}_m^H\boldsymbol{V}_m = \boldsymbol{I}_{d_m}, \quad (12)$$

$$\boldsymbol{U}_n : N_r \times d, \qquad \boldsymbol{U}_n^H\boldsymbol{U}_n = \boldsymbol{I}_n, \quad (13)$$

$$\boldsymbol{U}_n^H\boldsymbol{H}_{nm}\boldsymbol{V}_m = 0, \qquad \forall m \neq n, \quad (14)$$

$$rank(\boldsymbol{U}_n^H\boldsymbol{H}_{nm}\boldsymbol{V}_m) = d, \qquad \forall m = n. \quad (15)$$

Substituting (14) and (15) into (11), we can get:

$$\boldsymbol{Y}_n = \boldsymbol{U}_n\boldsymbol{H}_{nn}\boldsymbol{V}_n x_n + \textstyle\sum_{m\neq n}\boldsymbol{U}_n\boldsymbol{H}_{nm}\boldsymbol{V}_m x_m + \boldsymbol{U}_n\boldsymbol{Z}_n, \quad (16)$$

where x_n is a $d \times 1$ vector represents the d data streams the n-th transmitter sends. In other words, interference alignment method is that each transmitter chooses a direction that the interference it causes to undesired receivers lie outside of the d dimensions where the desired signals lie in, and each receiver chooses a direction that all interference can be completely eliminated. An example of 3-user two antennas case is given in Figure.2.

For symmetric systems where every user has the same number of antennas at either the transmitter or the receiver, and every user transmits the same number of data streams, the feasibility condition of IA to achieve a degrees of freedom (DOF) of d is [16]:

$$M_t + M_r \geq (K+1)d. \quad (17)$$

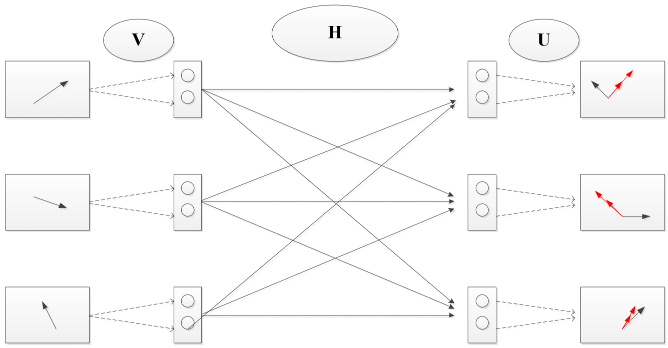

Figure.2 Interference solution for the three user 2 antenna case, the red arrow indicate the direction of the interference

A closed form solution for a 3-user scenario where $M_t = M_r = 2, d = 1$ has been given in [6, 7]. However, this closed form solution requires global channel knowledge which could be an overwhelming overhead in practice. An iterative algorithm needing only local channel knowledge was proposed in [8], which makes IA method much more practicable. The simulation result shows that with the iterative algorithm given in [8], although the feasibility condition of IA is no longer satisfied when the number of users increases, the interference in the network could be mitigated a lot to achieve a better performance of the sum-Capacity.

Closed form solution

For a 3-user MIMO interference network, the pre-coding matrix $\boldsymbol{V}_m$ at the transmitter can be decided by the following constraints to ensure that the interference signals at the receiver are in the same sub-space [7]:

$$\begin{cases} span(\boldsymbol{H}_{12}\boldsymbol{V}_2) = span(\boldsymbol{H}_{13}\boldsymbol{V}_3), \\ \boldsymbol{H}_{21}\boldsymbol{V}_1 = \boldsymbol{H}_{23}\boldsymbol{V}_3, \\ \boldsymbol{V}_1 = \boldsymbol{H}_{32}, \end{cases} \tag{18}$$

which can be expressed equivalently as:

$$\begin{cases} \mathrm{span}(\boldsymbol{V}_1) = \mathrm{span}(\boldsymbol{P}\boldsymbol{V}_1), \\ \boldsymbol{V}_2 = \boldsymbol{H}_{32}^{-1}\boldsymbol{H}_{31}\boldsymbol{V}_1, \\ \boldsymbol{V}_3 = \boldsymbol{H}_{23}^{-1}\boldsymbol{H}_{21}\boldsymbol{V}_1. \end{cases} \tag{19}$$

where $\boldsymbol{P} = \boldsymbol{H}_{31}^{-1}\boldsymbol{H}_{32}\boldsymbol{H}_{12}^{-1}\boldsymbol{H}_{13}\boldsymbol{H}_{23}^{-1}\boldsymbol{H}_{21}$. Then we can set $\boldsymbol{V}_1$ to be $\boldsymbol{V}_1 = [e_1 \quad e_2 \quad \cdots \quad e_d]$ where $e_1, e_2, \dots e_d$ are the d eigenvectors of $\boldsymbol{P}$. $\boldsymbol{V}_2$ and $\boldsymbol{V}_3$ are decided by (19).

At the receiver, the d rows of the interference suppressing matrix $\boldsymbol{U}_n$ are set to be the d eigenvectors corresponding to the smallest d eigenvalues of $\boldsymbol{Q} = \sum_{m\neq n}\boldsymbol{H}_{nm}\boldsymbol{V}_m\boldsymbol{V}_m^H\boldsymbol{H}_{nm}^H$ to achieve a maximum signal-to-interference-ratio (SIR).

Iterative algorithm

Due to (14), (15), on the reciprocal channel $\overleftarrow{\boldsymbol{H}}_{mn}^H = \boldsymbol{H}_{nm}^H$, $\boldsymbol{U}_n$ and $\boldsymbol{V}_m$ satisfy the following conditions [8]:

$$\boldsymbol{V}_m^H\overleftarrow{\boldsymbol{H}}_{mn}^H\boldsymbol{U}_n = 0, \qquad \forall m \neq n, \tag{20}$$

$$rank(\boldsymbol{V}_m^H\overleftarrow{\boldsymbol{H}}_{mn}^H\boldsymbol{U}_n) = d, \qquad \forall m = n. \tag{21}$$

It means that for the reciprocal channel $\boldsymbol{H}_{mn}^H$, $\boldsymbol{U}_n$ turns to be pre-coding matrix at the transmitter and $\boldsymbol{V}_m$ turns to be interference suppression matrix at the receiver, which leads to the iterate algorithm introduced by [8]:

- Transmit pre-coding matrix $\boldsymbol{V}_m$ at each transmitter is randomly chosen under constraint of (15).
- Compute the receive interference suppressing matrix $\boldsymbol{U}_n$ at each receiver. The d rows of $\boldsymbol{U}_n$ is defined as the d eigenvectors corresponding to the smallest d eigenvalues of $\boldsymbol{Q} = \sum_{m\neq n}\boldsymbol{H}_{nm}\boldsymbol{V}_m\boldsymbol{V}_m^H\boldsymbol{H}_{nm}^H$.
- Update the transmit pre-coding matrix $\boldsymbol{V}_m$ at each receiver. The d rows of $\boldsymbol{V}_m$ is defined as the d unit eigenvectors corresponding to the smallest d eigenvalues of $\overleftarrow{\boldsymbol{Q}} = \sum_{m\neq n}\overleftarrow{\boldsymbol{H}}_{mn}\boldsymbol{U}_n\boldsymbol{U}_n^H\overleftarrow{\boldsymbol{H}}_{mn}^H$.
- Repeat 2) to 3) tile $\boldsymbol{I} = \sum_{m=1}^{K}\sum_{n=1,n\neq m}^{K} Tr\,(\boldsymbol{U}_n^H\boldsymbol{Q}\boldsymbol{U}_n)$ converges, where $\boldsymbol{I}$ is defined as the "leakage interference", and it converges as shown in [8].

V. SIMULATION RESULTS

To verify the effectiveness of the proposed IA scheme, we compared its sum-capacity performance with that of pure DAS scheme for 60GHz system through Monte Carlo simulation.

Consider a $5\mathrm{m} \times 5\mathrm{m} \times 3\mathrm{m}$ room with K transmitter-receiver pairs randomly placed in it. At both the transmitter and the receiver, each user is equipped with 2**N* antennas. In the DAS scheme, the antennas are arranged in a square shape, DAS beamforming are performed between the transmitter antenna array and the receiver antenna array as illustrated in Section III. In the proposed IA scheme, the antennas at the transmitter and the receiver are divided into 2 subarrays with a 10cm subarray spacing to create a rank-2 equivalent MIMO channel by DAS beamforming between the corresponding antenna subarray pairs [13], with which each user sends 1 data stream. Inside each subarray, the antennas are arranged in a square shape with an inner antenna element spacing $\frac{\lambda}{2}$. The Rician *k*-factor is set to 6 dB.

Figure.3 (a) and Figure.3 (b) compares the sum-capacity performances of IA and DAS schemes in 3-user case with *N*=16 and 32 respectively. Both figures show that the IA scheme performs better while the sum-capacity of the pure DAS scheme hardly increases in the high transmit SNR region. This is because the interference among users grows as the transmit power increases, so in high SNR region where interference power is higher than the noise power, the sum-capacity could hardly increase if no more interference suppression method is used. As the feasibility condition of IA is satisfied in these scenarios, the interference are almost completely eliminated, and the sum-Capacity in the IA scheme grows linearly with the transmit power does.

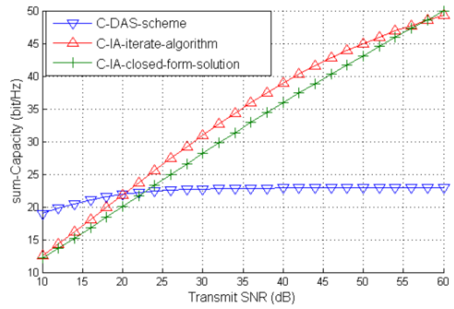

Figure.3 (a) Average sum-capacity of *K*=3 *N*=16 case

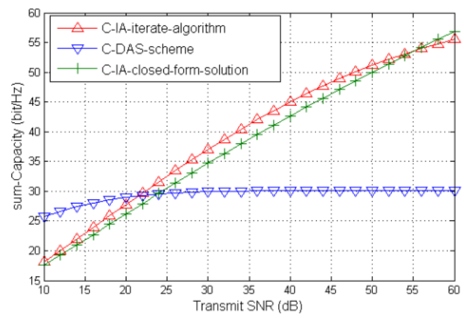

Figure.3 (b) Average sum-capacity of *K*=3 *N*=32 case

Figure.4 shows a 5-user case that each transmitter or receiver equipped with 16 antennas. In this case, the feasibility condition of IA is not satisfied [16], thus a "floor" effect appears in high transmit SNR region since the non-negligible remaining interference prevents the sum-capacity of the network keeping increasing. However, the interference is still reduced greatly by each step of the iterative algorithm, and a much higher sum-capacity than that of the DAS scheme is achieved. Figure.5 illustrates the relation of sum-capacity and user numbers under 30dB transmit SNR and N=16. It can be seen that the performance improvement of IA scheme reduces slightly with the increase of user numbers. That is because an additional user would bring extra interference to the network, while the performance of the IA scheme is constrained when the feasibility condition of IA is no longer satisfied for $K > 3$.

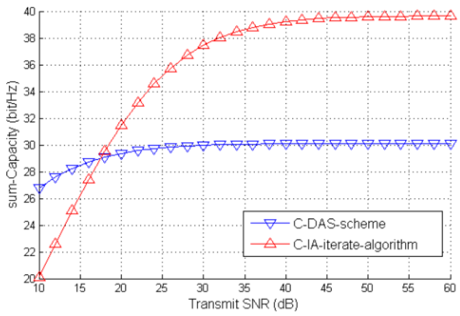

Figure.4 Average sum-capacity of K=5 N=16 case

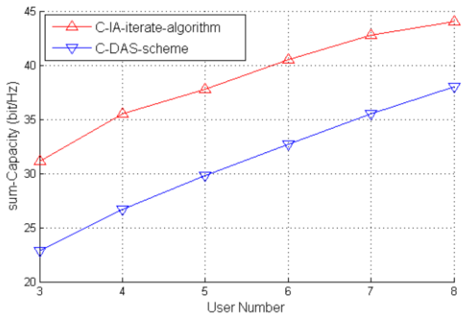

Figure.5 Average sum-capacity of N=16 case under 30dB transmit SNR with different user numbers

VI. CONCLUSION

In this paper, an interference alignment scheme based on subarray partition is proposed for 60GHz millimeter wave network. The whole antenna array is partitioned into M subarrays, where beamforming is performed within each subarray. The beamformed M subarray pairs at both sides of the transmitter and the receiver form an equivalent rank-M MIMO channel, so the interference alignment algorithm can be used to further mitigate the interference and optimize the sum-Capacity of the network. Simulation results show that the pure directional beamforming is not enough for fully exploit the potential of spatial reuse in 60GHz network, while the proposed IA scheme can effectively mitigate the interference and significantly improve the sum-capacity in high transmit SNR region.

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