A New Approach of Interference Alignment through Asymmetric Complex Signaling and Multiuser Diversity

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***Abstract*—In this letter, we consider a new interference alignment (IA) strategy for single-input single-output interference broadcast channels with constant channel coefficients. First, we show that 1.5 degrees of freedom (DOF) is achievable for 3-cell case by utilizing asymmetric complex signaling (AC) and multiuser diversity without symbol extension. It is also investigated that the achievable DOF varies with the user scaling condition and $\omega\left(\sqrt{SNR}\right)$ is required for guaranteeing the DOF of 1.5. To improve the sum-rate performance, user scheduling algorithms combined with the beamforming techniques are suggested which outperform the conventional IA schemes. After introducing an exhaustive scheduling algorithm which shows optimal sum-rate, a simplified scheduling method is also proposed which reduces both scheduling metric computations and the search size.**

***Index Terms*—Interference alignment, multiuser diversity, asymmetric complex signaling.**

I. Introduction

FOR interference channels (IFCs), many researchers have studied interference management algorithms and have analyzed degrees-of-freedom (DOF). The interference alignment (IA) algorithm introduced in [1] [2] nullifies the interference signals and obtains the optimal DOF by simple zero-forcing (ZF) receivers. This scheme requires redundant dimensions in order to align interference to approximately half of the received signal space. Hence, in the single-input single-output (SISO) case, we need a large number of symbol extension on frequency/time varying channels to achieve the optimal DOF. In the constant channel case, the authors in [3] proposed a novel precoding designing method based on asymmetric complex (AC) signaling. However, this scheme still demands the symbol extension to obtain the DOF of 1.2. Also, for uplink environments, an opportunistic strategy to mitigate the interference was suggested in [4].

In this letter, we propose a new approach of IA based on the AC signaling and multiuser diversity (MUD) for 3-cell SISO interference broadcast channels (IFBCs) where each BS covers K users per cell. For channels with constant coefficients, we first verify that the DOF of 1.5 is achievable through MUD and AC signaling without requiring the symbol extension. By utilizing extra spatial dimension introduced through the AC signaling, the beamforming vector can be designed to guarantee at least one DOF. Then, we prove that an additional DOF of 0.5 is achieved by employing the MUD. It is also shown that the DOF ranging from 1 to 1.5 can be obtained depending on the relation between the number of users and SNR.

The algorithms which only optimize DOF usually show poor performance in terms of sum-rate. To overcome this, reference [5] considers beamforming techniques for maximizing weighted sum-rate in multi-input single-output IFCs. In this work, we propose advanced scheduling algorithms combined with the beamforming techniques which outperform the conventional IA schemes in the practical SNR range. First, we introduce a sum-rate maximizing algorithm which fully utilizes the MUD by exhaustively searching the optimal user set over all possible user combinations. As, the computational complexity in this scheme may become too high with growing K, we suggest a simplified method which reduces both scheduling metric computations and the search size. In the simulation section, we will show that our proposed schemes outperform the conventional IA methods in the practical SNR region even with a small number of users.

Note that there is a limitation when generalizing our AC signaling and MUD based schemes to N cells. For more than 3 cells, the interference signals cannot be aligned in the half of the received signal space while being independent of the desired signal without using symbol extension. In such cases, we need to find an intelligent way for combining phase alignment, MUD and symbol extension, and this would be an interesting future topic.

Throughout the letter, the following symbols are employed: i) $f(x) = o\left(g(x)\right)$ states $\lim_{x\to\infty}\frac{f(x)}{g(x)} = 0$. ii) $f(x) = \Theta\left(g(x)\right)$ means that there exist positive constants c_1, c_2 and k such that $0 \leq c_1 g(x) \leq f(x) \leq c_2 g(x)$ for all $x \geq k$. iii) $f(x) = \omega\left(g(x)\right)$ indicates $\lim_{x\to\infty}\frac{g(x)}{f(x)} = 0$ [6].

II. System Model

We consider 3-cell SISO IFBCs where one BS and K users exist in each cell. In this model, each BS transmits independent data symbols to an intended user in the corresponding cell. Suppose that the schedulers select user $(1, k_1)$, $(2, k_2)$ and $(3, k_3)$ where user (i, j) represents the j^{th} user in cell i. Then the channel model can be regarded as 3-cell IFCs, and the received signal of user (i, k_i) is given as

$$y_{k_i}^{(i)} = h_{k_i}^{(ii)} x_{k_i}^{(i)} + \sum_{j=1, j\neq i}^{3} h_{k_i}^{(ij)} x_{k_i}^{(j)} + n_{k_i}^{(i)}$$

Manuscript received December 28, 2010; revised June 20 and September 30, 2011; accepted November 7, 2011. The associate editor coordinating the review of this letter and approving it for publication was H. Shin.

This work was supported by the National Research Foundation of Korea (NRF) grant funded by the Korea government (MEST) (No. 2010-0017909). The material in this paper was presented in part at IEEE Globecom, Dec. 2010.

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Digital Object Identifier 10.1109/TWC.2012.010312.102034

where $k_i \in \{1, \cdots, K\}$ denotes the selected user index for cell i, $x_{k_i}^{(j)}$ indicates the transmitted data from BS j to user (i, k_i), $h_{k_i}^{(ij)}$ represents the complex channel coefficient from BS j to user (i, k_i) and $n_{k_i}^{(i)} \sim \mathcal{CN}(0, \sigma_n^2)$ stands for the additive white Gaussian noise at user (i, k_i).

To secure the spatial space for IA, we employ the AC signaling and the channel input-output equation is written as

$$\mathbf{y}_{k_i}^{(i)} = \mathbf{H}_{k_i}^{(ii)}\mathbf{x}_{k_i}^{(i)} + \sum_{j=1, j\neq i}^{3} \mathbf{H}_{k_i}^{(ij)}\mathbf{x}_{k_i}^{(j)} + \mathbf{n}_{k_i}^{(i)}$$

where $\mathbf{y}_{k_i}^{(i)} = [\text{Re}\{y_{k_i}^{(i)}\} \ \ \text{Im}\{y_{k_i}^{(i)}\}]^T$, $\mathbf{x}_{k_i}^{(i)} = [\text{Re}\{x_{k_i}^{(i)}\} \ \ \text{Im}\{x_{k_i}^{(i)}\}]^T$, $\mathbf{n}_{k_i}^{(i)} = [\text{Re}\{n_{k_i}^{(i)}\} \ \ \text{Im}\{n_{k_i}^{(i)}\}]^T$ and

$$\mathbf{H}_{k_i}^{(ij)} = \left|h_{k_i}^{(ij)}\right| \mathbf{J}(\theta_{k_i}^{(ij)}).$$

Here, $\text{Re}\{\cdot\}$ and $\text{Im}\{\cdot\}$ indicates the real part and imaginary part, respectively, and $\theta_{k_i}^{(ij)}$ equals $\angle h_{k_i}^{(ij)}$. Also $\mathbf{J}(\theta)$ is defined as

$$\mathbf{J}(\theta) = \begin{bmatrix} \cos\theta & -\sin\theta \\ \sin\theta & \cos\theta \end{bmatrix}.$$

Now, the channel model can be considered as virtual multiple antenna systems where both BS and users have two antennas, which means that we can utilize the redundant antenna space for IA. For this, we design the signal from transmitter i as $\mathbf{x}_{k_i}^{(i)} = \mathbf{v}^{(i)}u^{(i)}$ where $\mathbf{v}^{(i)} \in \mathbb{R}^{2\times 1}$ is the beamforming vector with the per-base power constraint $\|\mathbf{v}^{(i)}\|^2 \leq P$ and $u^{(i)}$ represents the real scalar data.

III. Achievable Degrees of Freedom with AC Signaling and MUD

In this section, we prove that the DOF of 1.5 is achievable in 3-cell IFBCs without the symbol extension method and the time/frequency varying channel coefficients. In the cellular network system, the achievable DOF for i^{th} cell is derived as

$$d^{(i)} = \lim_{\rho\to\infty} \frac{R^{(i)}(\rho)}{\log_2(\rho)}$$

where $R^{(i)}$ indicates the achievable rate for cell i at SNR ρ in bps/Hz.

First, we employ the AC signaling to guarantee the achievable DOF of 1, and then utilize MUD to obtain an additional DOF. The relationship between the achievable DOF performance and the number of users will be also investigated.

A. IA with AC signaling

For achieving the DOF of 1.5, each beamforming vector needs to be designed to satisfy the following IA conditions as [2]

$$\text{span}\left(\mathbf{H}_{k_1}^{(12)}\mathbf{v}^{(2)}\right) = \text{span}\left(\mathbf{H}_{k_1}^{(13)}\mathbf{v}^{(3)}\right) \quad (1)$$

$$\text{span}\left(\mathbf{H}_{k_2}^{(21)}\mathbf{v}^{(1)}\right) = \text{span}\left(\mathbf{H}_{k_2}^{(23)}\mathbf{v}^{(3)}\right) \quad (2)$$

$$\text{span}\left(\mathbf{H}_{k_3}^{(31)}\mathbf{v}^{(1)}\right) = \text{span}\left(\mathbf{H}_{k_3}^{(32)}\mathbf{v}^{(2)}\right). \quad (3)$$

Both (1) and (2) can be achieved if the beamforming vectors is determined as [3][7]

$$\mathbf{v}^{(1)} = \sqrt{P}\,\mathbf{j}_c(\theta_{k_2}^{(23)} - \theta_{k_2}^{(21)} + \theta_0), \quad (4)$$

$$\mathbf{v}^{(2)} = \sqrt{P}\,\mathbf{j}_c(\theta_{k_1}^{(13)} - \theta_{k_1}^{(12)} + \theta_0), \quad (5)$$

$$\mathbf{v}^{(3)} = \sqrt{P}\,\mathbf{j}_c(\theta_0) \quad (6)$$

where $\mathbf{j}_c(\theta) = \begin{bmatrix} \cos\theta & \sin\theta \end{bmatrix}^T$. Here, θ_0, which represents the phase of $\mathbf{v}^{(3)}$, can be set as any value because it has no effect on the sum-rate performance. Thus, we set θ_0 to 0 for notational conveniences.

The beamforming vectors in (4)$-$(6) manipulate the channel phases and align intereference for users $(1, k_1)$ and $(2, k_2)$. Hence, the achievable DOF of 1 is obtained by applying ZF receivers at each user. However, the condition (3) cannot be satisfied by the beamformers in (4)$-$(6) due to random channel realizations. Instead, in the following, by exploiting the abundant channel characteristics of users in the 3^{rd} cell, we obtain an additional achievable DOF.

B. Analysis for the achievable DOF

After aligning the interference of user $(1, k_1)$ and $(2, k_2)$, the received signal at user $(3, k_3)$ is represented by

$$\begin{aligned}\mathbf{y}_{k_3}^{(3)} &= \mathbf{H}_{k_3}^{(33)}\mathbf{v}^{(3)}u^{(3)} + \mathbf{H}_{k_3}^{(31)}\mathbf{v}^{(1)}u^{(1)} + \mathbf{H}_{k_3}^{(32)}\mathbf{v}^{(2)}u^{(2)} + \mathbf{n}_{k_3}^{(3)} \\ &= \sqrt{P}\left|h_{k_3}^{(33)}\right|\mathbf{j}_c\left(\theta_{k_3}^{(33)}\right)u^{(3)} + \sqrt{P}\left|h_{k_3}^{(31)}\right|\mathbf{j}_c\left(\psi_{1,k_3}\right)u^{(1)} \\ &+ \sqrt{P}\left|h_{k_3}^{(32)}\right|\mathbf{j}_c\left(\psi_{2,k_3}\right)u^{(2)} + \mathbf{n}_{k_3}^{(3)}\end{aligned}$$

where $\psi_{1,k_3} = \theta_{k_3}^{(31)} + \theta_{k_2}^{(23)} - \theta_{k_2}^{(21)}$ and $\psi_{2,k_3} = \theta_{k_3}^{(32)} + \theta_{k_1}^{(13)} - \theta_{k_1}^{(12)}$. Different from the received signal of user $(1, k_1)$ and $(2, k_2)$, the interference signal of user $(3, k_3)$ cannot be eliminated by a ZF receiver. Nonetheless, we apply the ZF receiver $\mathbf{j}_r\left(\frac{\pi}{2} - \psi_{1,k_3}\right)$ where $\mathbf{j}_r(\theta) = \begin{bmatrix} \cos\theta & -\sin\theta \end{bmatrix}$ to nullify one of the interference term. Then the achievable individual rate for user $(3, k_3)$ is calculated as

$$R^{(3)}(\rho) = \frac{1}{2}\log_2\left(1 + \frac{\left|h_{k_3}^{(33)}\right|^2 \rho\cos^2\left(\frac{\pi}{2} - \psi_{1,k_3} + \theta_{k_3}^{(33)}\right)}{\left|h_{k_3}^{(32)}\right|^2 \rho\cos^2\left(\frac{\pi}{2} - \psi_{1,k_3} + \psi_{2,k_3}\right) + 1}\right) \quad (8)$$

where $\frac{1}{2}$ is introduced because a real data is transmitted in our channel model. From (8), we can derive the achievable DOF for 3-cell SISO IFBCs in the following theorem.

Theorem 1: The DOF of 1.5 is achievable in 3-cell SISO IFBCs without the symbol extension if and only if $K = \omega\left(\sqrt{\rho}\right)$.

Proof: The DOF of 0.5 can be obtained from user $(3, k_3)$ if the following condition is satisfied

$$\rho\cos^2\left(\frac{\pi}{2} - \psi_{1,k_3} + \psi_{2,k_3}\right) \leq \epsilon \quad (9)$$

where ϵ is an arbitrary small positive number independent of SNR. Then, we can have the lower bound of the DOF for user $(3, k_3)$ as

$$d^{(3)} \geq \frac{1}{2}P_{IA}$$

where

$$P_{IA} = \lim_{\rho\to\infty} Pr\left\{\bigcup_{k_3=1}^{K} \rho\cos^2\left(\frac{\pi}{2} - \psi_{1,k_3} + \psi_{2,k_3}\right) \leq \epsilon\right\}. \quad (10)$$

TABLE I
ACHIEVABLE DOF BASED ON USER SCALING BOUNDS

User Scaling Bounds	Achievable DOF
$o\left(\sqrt{\rho}\right)$	$d_\Sigma = 1$
$\Theta\left(\sqrt{\rho}\right)$	$1 < d_\Sigma < 1.5$
$\omega\left(\sqrt{\rho}\right)$	$d_\Sigma = 1.5$

$$R_\Sigma(S) = \sum_{m=1}^{3} \frac{1}{2}\log_2\left[\det\left(\mathbf{I}_2 + \left(\sum_{i\neq m}\mathbf{H}_{k_m}^{(mi)}\mathbf{v}^{(i)}\mathbf{v}^{(i)T}\mathbf{H}_{k_m}^{(mi)T} + \frac{\sigma_n^2}{2}\mathbf{I}_2\right)^{-1}\mathbf{H}_{k_m}^{(mm)}\mathbf{v}^{(m)}\mathbf{v}^{(m)T}\mathbf{H}_{k_m}^{(mm)T}\right)\right] \quad (7)$$

Examining (10), we can notice that both ψ_{1,k_3} and ψ_{2,k_3} depend only on the channel characteristic of user $(3, k_3)$, because $\theta_{k_2}^{(23)}$, $\theta_{k_2}^{(31)}$, $\theta_{k_1}^{(13)}$ and $\theta_{k_1}^{(12)}$ have already been determined in the process of obtaining the DOF of 1 in the previous subsection. Hence, we can replace $\frac{\pi}{2} - \psi_{1,k_3} + \psi_{2,k_3}$ by θ_k for notational convenience.

Then P_{IA} is lower-bounded by

$$\begin{aligned} P_{IA} &= \lim_{\rho\to\infty} Pr\left\{\bigcup_{k=1}^{K}\cos^2\theta_k \leq \epsilon\rho^{-1}\right\} \\ &= 1 - \lim_{\rho\to\infty} Pr\left\{\bigcap_{k=1}^{K}\cos^2\theta_k > \epsilon\rho^{-1}\right\} \\ &= 1 - \lim_{\rho\to\infty}\left(1 - Pr\left\{\cos^2\theta \leq \epsilon\rho^{-1}\right\}\right)^K \\ &= 1 - \lim_{\rho\to\infty}\left(1 - F_{\cos^2\theta}\left(\epsilon\rho^{-1}\right)\right)^K \\ &\geq 1 - \lim_{\rho\to\infty}\left(1 - \frac{2\sqrt{\epsilon\rho^{-1}}}{\pi}\right)^K \qquad (11) \\ &= 1 - \lim_{\rho\to\infty}\exp\left(-\frac{2\sqrt{\epsilon}}{\pi}\right)^{K\rho^{-\frac{1}{2}}} \end{aligned}$$

where $F_{cos^2\theta}(\alpha)$ means the cumulative distribution function of the random variable $\cos^2\theta$ given by

$$F_{\cos^2\theta}(\epsilon\rho^{-1}) = \frac{1}{\pi}\left(\cos^{-1}\left(-\sqrt{\epsilon\rho^{-1}}\right) - \cos^{-1}\left(\sqrt{\epsilon\rho^{-1}}\right)\right), \quad (12)$$

and the inequality (11) is derived by the property of the inverse trigonometric function as [8]

$$\cos^{-1} z = \frac{\pi}{2} - \sum_{n=0}^{\infty}\left(\frac{(2n)!}{2^{2n}(n!)^2}\frac{z^{2n+1}}{(2n+1)}\right).$$

It is clear in (11) that the lower-bound of P_{IA} approaches to 1 as the number of users K asymptotically increases much faster than $\sqrt{\rho}$. Finally, the total number of DOF $\sum d^{(i)}$ is lower-bounded $1 + \frac{1}{2}P_{IA}$ which converges to 1.5 if $K = \omega(\sqrt{\rho})$. ■

We can notice that the lower bound of the achievable DOF varies from 1 to 1.5 depending on the relation between the number of users and SNR. Table I presents the achievable DOF with different user growth rates. When the number of users grows much faster than $\sqrt{\rho}$, we surely find the user set whose phases satisfies (9), which means that each user can decode interference free real data.

IV. ADVANCED SCHEDULING ALGORITHMS

In the previous section, we have discussed that the DOF of 1.5 is achievable with the user condition of $\omega(\sqrt{\rho})$. However, this condition may be impractical, and an algorithm only focusing on the DOF tends to show degraded sum-rate performance in low and mid SNR regime. Motivated by this, we propose advanced scheduling algorithms in conjunction with the beamforming techniques which outperform the conventional schemes in the practical SNR regime even with the small number of users.

A. Exhaustive Search Algorithm

The optimal scheduling algorithm searches for the best user set among all possible K^3 user combinations. By employing the beamforming vectors (4)-(6), the sum-rate can be computed as (7) at the top of the this page where $S = (k_1, k_2, k_3) \in \{1, \cdots, K\}^3$ denotes the selected user combination. As a result, the optimal user combination is determined by

$$(k_1^*, k_2^*, k_3^*) = \arg\max_{S\in\{1,\cdots,K\}^3} R_\Sigma(S).$$

This user selection method requires the search size of K^3 and fully utilizes the MUD so that the enhanced sum-rate performance is expected as the number of users in each cell increases.

B. Reduced Search Algorithm

In this subsection, we propose a simplified scheduling algorithm which reduces both sum-rate metric calculations and the search size of the exhaustive search algorithm. In this way, we can save computational complexity at the expense of little performance loss. To begin with, we show how the sum-rate expression in (7) can be simplified. Since the interference signals at user $(1, k_1)$ and $(2, k_2)$ are perfectly aligned, we express their individual rates under the assumption that ZF receivers $\mathbf{j}_r\left(\frac{\pi}{2} - \theta_{k_1}^{(13)}\right)$ and $\mathbf{j}_r\left(\frac{\pi}{2} - \theta_{k_2}^{(23)}\right)$ are applied for user $(1, k_1)$ and $(2, k_2)$, respectively. Then the rates for user $(1, k_1)$ and $(2, k_2)$ are computed as

$$R_{ZF}^{(i)} = \frac{1}{2}\log_2\left(1 + \frac{\left|h_{k_i}^{(ii)}\right|^2 P\cos^2\left(\theta_{k_i}^{(ii)} + \theta_{k_{\bar{i}}}^{(\bar{i}3)} - \theta_{k_{\bar{i}}}^{(\bar{i}i)} + \frac{\pi}{2} - \theta_{k_i}^{(i3)}\right)}{\frac{\sigma_n^2}{2}}\right)$$

where $i = 1, 2$ and we define $\bar{1} = 2$ and $\bar{2} = 1$.

The rate for user $(3, k_3)$, on the other hand, is given without any approximation as

$$R^{(3)}=\frac{1}{2}\log_2\left(1+\frac{\alpha_1\alpha_3\sin^2(\delta_1-\delta_3)+\alpha_2\alpha_3\sin^2(\delta_2-\delta_3)+\frac{\sigma_n^2}{2}\alpha_3}{\alpha_1\alpha_2\sin^2(\delta_1-\delta_2)+\frac{\sigma_n^2}{2}(\alpha_1+\alpha_2)+\frac{\sigma_n^4}{4}}\right) \quad (13)$$

where we have

$$\alpha_1=\left|h_{k_3}^{(31)}\right|^2 P,\ \alpha_2=\left|h_{k_3}^{(32)}\right|^2 P,\ \alpha_3=\left|h_{k_3}^{(33)}\right|^2 P,$$
$$\delta_1=\theta_{k_3}^{(31)}+\theta_{k_2}^{(23)}-\theta_{k_2}^{(21)},\ \delta_2=\theta_{k_3}^{(32)}+\theta_{k_1}^{(13)}-\theta_{k_1}^{(12)},\ \delta_3=\theta_{k_3}^{(33)}.$$

Consequently, the simplified sum-rate for user combination (k_1, k_2, k_3) is written by

$$R_{\Sigma}^{sim}(k_1, k_2, k_3) = R_{ZF}^{(1)} + R_{ZF}^{(2)} + R^{(3)}. \quad (14)$$

Now, we discuss how to reduce the search size. The reduced search algorithm is composed of the following steps. First, user $(3, k_3)$ is determined by the one who has the maximal individual rate. However, we can notice that the rate of user $(3, k_3)$ in (13) is influenced by not only the channel phases of user $(3, k_3)$ but also user $(1, k_1)$ and $(2, k_2)$. Hence, to identify user $(3, k_3)$ only by its channel characteristic, (13) is approximated as

$$\tilde{R}^{(3)} = \frac{1}{2}\log_2\left(1+\frac{\alpha_1\alpha_3+\alpha_2\alpha_3+\sigma_n^2\alpha_3}{\alpha_1\alpha_2+\sigma_n^2(\alpha_1+\alpha_2)+\frac{\sigma_n^4}{2}}\right).$$

Here, we set other cell dependent terms of (13), $\sin^2_{i\neq j}(\delta_i-\delta_j)$, as $\frac{1}{2}$ because its expected value for random channel phases is $\frac{1}{2}$. As a result, we first choose user $(3, k_3)$ as

$$\tilde{k}_3 = \arg\max_{k_3\in\{1,\cdots,K\}} \tilde{R}^{(3)}. \quad (15)$$

Now that the channel magnitude and the phase of user $(3, \tilde{k}_3)$ are available, we are able to utilize the simplified sum-rate expression in (14) for identifying users $(1, k_1)$ and $(2, k_2)$ with the given user $(3, \tilde{k}_3)$. Hence, the user combination (k_1^*, k_2^*) can be chosen by

$$(k_1^*, k_2^*) = \arg\max_{(k_1,k_2)\in\{1,\cdots,K\}^2} R_{\Sigma}^{sim}\left(k_1, k_2, \tilde{k}_3\right). \quad (16)$$

After that, we update the choice of user $(3, k_3)$ to improve the sum-rate performance. Since we already know the channel information of users $(1, k_1^*)$ and $(2, k_2^*)$, the newly updated user $(3, k_3^*)$ is determined with the simplified metric as

$$k_3^* = \arg\max_{k_3\in\{1,\cdots,K\}} R^{(3)} \quad (17)$$

where δ_1 and δ_2 in $R^{(3)}$ are computed by users $(1, k_1^*)$, $(2, k_2^*)$ and $(3, k_3)$.

As for the search complexity, the proposed scheme requires the search size of $K^2 + 2K$. In (16), we can reduce the number of multiplications for computing the metric from 192 to 42 by employing the approximated sum-rate in (14) instead of (7). Also, 26 and 32 multiplications are required at the stage of (15) and (17), respectively. Thus, the total number of multiplications is $42K^2+58K$. For instance, in the case of $K = 10$, the required multiplications of the optimal scheduling algorithm and the reduced search algorithm are equal to 192000 and 4780, respectively. Hence, the reduced search algorithm exhibits much lower complexity than the exhaustive method, and this advantage grows with increasing K.

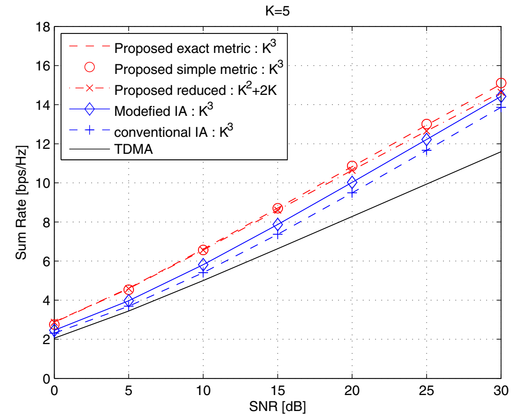

Fig. 1. Sum-rate performance in 3-cell IFBCs with K=5.

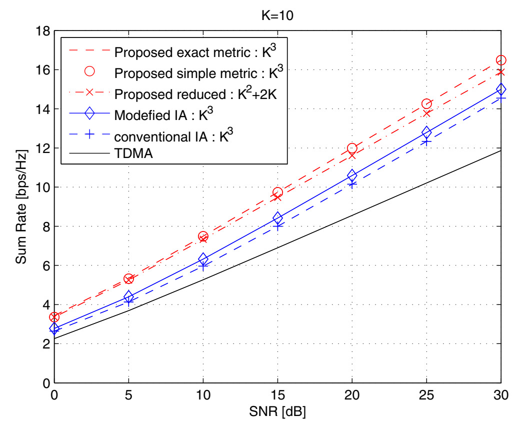

Fig. 2. Sum-rate performance in 3-cell IFBCs with K=10.

V. SIMULATION RESULTS

In this section, we compare the proposed algorithms with other IA based schemes to check the effectiveness of our strategy. Here, we will only consider the fast-fading model for simplicity. The simulation results with a pathloss model will exhibit a similar tendency to the present results. It is assumed that all users are located in the cell edge where is the same distance from all BSs. In simulations, full power P is consumed at the each BS and SNR is represented as $\frac{P}{\sigma_n^2}$. The channel coefficients are generated from an independent and identically distributed complex Gaussian distribution with zero mean and unit variance. The conventional IA refers to the beamforming strategy with symbol extension suggested in [2]. We also consider the modified IA in [9] which improves the sum-rate performance compared to the conventional IA. Here, the TDMA scheme supports a user with the maximum rate among $3K$ users at each time slot. In the simulations, the conventional IA and the modified IA employ the beamforming vectors in [2] and those in [9], respectively and search the best user set which maximizes their sum-rate from K^3 user combinations.

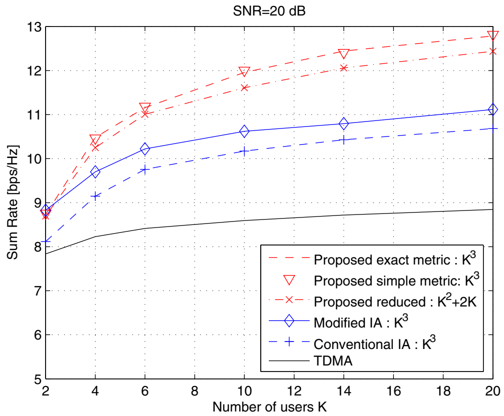

Fig. 3. Sum-rate performance in terms of K at SNR = 20 dB.

In Figures 1 and 2, the sum-rate performance of various algorithms are plotted with 5 and 10 users, respectively. The simplified sum-rate expression in (14) shows the near-optimal performance compared to the exact sum-rate metric. We also notice that our proposed algorithms outperform the existing schemes in the practical SNR range, and the performance gap increases with larger K. It can be found that the sum-rate gap between the exhaustive search and the reduced search is small.

Figure 3 exhibits the sum-rate comparison in terms of the number of users at the SNR of 20 dB. Observing the sum-rate gap, we can see that the gain of the MUD is greater in the proposed algorithm than other conventional methods. The proposed methods show better sum-rate performance than other schemes even in small K. Notice that the proposed methods do not require the symbol extension, and thus substantially reduce the complexity in precoder designs.

VI. Conclusion

In this letter, we have exploited the MUD and AC signaling in interference broadcast channels. We have shown that the DOF of 1.5 is achievable without symbol extension and time/frequency varying environment. Motivated by this, we have proposed an exhaustive search algorithm combined with beamforming vectors which maximizes the sum-rate. To reduce both sum-rate metric computations and the search size, a simplified search algorithm has been suggested. In the simulations, we have confirmed that the proposed schemes outperform the conventional IA schemes in the practical SNR region with much lower complexity. For more than 3-cell cases, the question is how much DOFs are achievable with AC signaling and MUD in the constant channel. To generalize our work to more than 3-cell cases, we need to consider other ways such as symbol extension.

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