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Dynamic Interference Steering in Heterogeneous Cellular Networks

Z. LI^{1,2}, (Member, IEEE), F. GUO³, C. SHU³, KANG G. SHIN⁴, (Life Fellow, IEEE), AND J. LIU⁵, (Member, IEEE)

¹School of Cyber Engineering, Xidian University, Xi'an 710126, China

²Shaanxi Key Laboratory of Information Communication Network and Security, Xi'an University of Posts and Telecommunications, Xi'an 710121, China

³School of Telecommunications, Xidian University, Xi'an 710071, China

⁴Department of Electrical Engineering and Computer Science, University of Michigan, Ann Arbor, MI 48109-2121, USA

⁵Center for Cybersecurity Research and Development, National Institute of Informatics, Tokyo 101-8430, Japan

Corresponding author: Z. Li (zli@xidian.edu.cn)

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ABSTRACT With the development of diverse wireless communication technologies, interference has become a key impediment in network performance, thus making effective interference management (IM) essential to accommodate a rapidly increasing number of subscribers with diverse services. Although there have been numerous IM schemes proposed thus far, none of them are free of some form of cost. It is, therefore, important to balance the benefit brought by and cost of each adopted IM scheme by adapting its operating parameters to various network deployments and dynamic channel conditions. We propose a novel IM scheme, called *dynamic interference steering* (DIS), by recognizing the fact that interference can be not only suppressed or mitigated but also steered in a particular direction. Specifically, DIS exploits both channel state information and the data contained in the interfering signal to generate a signal that modifies the spatial feature of the original interference to partially or fully cancel the interference appearing at the victim receiver. By intelligently determining the strength of the steering signal, DIS can steer the interference in an optimal direction to balance the transmitter's power used for interference steering and the desired signal's transmission. DIS is shown via simulation to be able to make better use of the transmit power, hence enhancing users' spectral efficiency effectively.

INDEX TERMS Interference, costs, adaptive signal processing.

I. INTRODUCTION

Due to the broadcast nature of wireless communications, concurrent transmissions from multiple source nodes may cause interferences to each other, thus degrading subscribers' data rate. Interference management (IM) is, therefore, crucial to meet the ever-increasing demand of diverse users' Quality-of-Service (QoS).

Interference alignment (IA) is a powerful means of controlling interference and has thus been under development in recent years. By preprocessing signals at the transmitter, multiple interfering signals are mapped into a certain signal subspace, i.e., the overall interference space at the receiver is minimized, leaving the remaining subspace interference-free [1], [2]. IA is shown to be able to achieve the information-theoretic maximum DoF (Degree of Freedom) in some interference networks [3], [4]. However, to achieve such

a promising gain, it is required to use either infinite symbol extensions over time/frequency [4] or a large number of antennas at each receiver [5], both which are not realistic. That is, IA emerges as a promising IM scheme, but its applicability is severely limited by the high DoF requirement.

Some researchers have attempted to circumvent the stringent DoF requirement by proposing other IM schemes, such as interference neutralization (IN). IN refers to the distributed zero-forcing of interference when the interfering signal traverses multiple nodes before arriving at the undesired receivers/destinations. The basic idea of IN has been applied to deterministic channels and interference networks [6]–[10], both of which employ relays. IM was studied in the context of a deterministic wireless interaction model [6], [7]. In [6], IN was proposed and an admissible rate region of the Gaussian ZS and ZZ interference-relay networks was obtained,

where ZS and ZZ denote two special configurations of a two-stage interference-relay network in which some of the cross-links are weak. Mohajer *et al.* [7] further translated the exact capacity region obtained in [6] into a universal characterization for the Gaussian network. The key idea used in the above interference networks with relays is to control the precoder at the relay so that the sum of the channel gains of the newly-created signal path via the relay and the direct path to the destination becomes zero. A new scheme called *aligned interference neutralization* was proposed in [8] by combining IA and IN. It provides a way to align interference terms over each hop so as to cancel them over the air at the last hop. However, this conventional relay incurs a processing delay — while the direct path does not — between a source-destination pair, limiting the DoF gain in a wireless interference network. To remedy this problem, an instantaneous relay (or relay-without-delay) was introduced in [9] and [10] to obtain a larger capacity than the conventional relay, and a higher DoF gain was achieved without requiring any memory at the relay. Although the studies based on this instantaneous relay can provide some useful theoretical results, this type of relay is not practical. On the other hand, IN can mitigate interference, but the power overhead of generating neutralization signals also affects the system’s performance. To the best of our knowledge, the power overhead has not been considered in any of the existing studies related to IN — either more recent interference neutralization [8]–[10] or the same ideas known via other names for many years, such as distributed orthogonalization, distributed zero-forcing, multiuser zero-forcing, and orthogonalize-and-forward [11], [12]. In practice, higher transmit power will be used by IN when the interference is strong, thus making less power available for the desired data transmission. Furthermore, IN may not even be available for mobile terminals due to their limited power budget. Recently, a dynamic IN was proposed in [13] by taking into account the power cost for IN. This scheme can optimally make use of the transmit power used for interference neutralization and desired signal’s transmission, so that user’s SE is effectively improved.

By recognizing the fact that interference can be not only neutralized but also steered in a particular direction, one can “steer” interference, which we call *interference steering* (IS) [14]. That is, a steering signal is generated to modify the interference’s spatial feature so as to steer the original interference in the direction orthogonal to that of the desired signal perceived at the victim receiver. Note, however, that IS simply steers the original interference in the direction orthogonal to the desired signal, which we call *Orthogonal-IS* (OIS) in the following discussion, regardless of the underlying channel conditions, as well as the strength and spatial feature of interference(s) and the intended transmission. Therefore, the tradeoff between the benefit of IS (i.e., interference suppression) and its power cost was not considered there. Since the more transmit power is spent on steering interference, the less power for the desired signal’s transmission will be available, one can naturally raise a question: “Is it always

necessary to steer interference in the direction orthogonal to the desired signal?”

To answer the above question, we propose a new IM scheme, called *dynamic interference steering* (DIS). With DIS, the spatial feature of the steered interference at the intended receiver is intelligently determined so as to achieve a balance between the transmit power consumed by IS and the residual interference due to the imperfect interference suppression, thus improving the user’s SE.

The contributions of this paper are two-fold:

- Proposal of a novel IM scheme called *dynamic interference steering* (DIS). By intelligently determining the strength of steering signal, we balance the transmit power used for IS and that for the desired signal’s transmission. DIS can also subsume orthogonal-IS as a special case, making it more general.
- Extension of DIS to general cases where the number of interferences from macro base station (MBS), the number of desired signals from a pico base station (PBS) to its intended pico user equipment (PUE), and the number of PBSs and PUEs are all variable.

The rest of this paper is organized as follows. Section II describes the system model, while Section III details the dynamic interference steering. Section IV presents the generalization of DIS and Section V evaluates its performance and overhead. Finally, Section VI concludes the paper.

Throughout this paper, we use the following notations. The set of complex numbers is denoted as \mathbb{C} , while vectors and matrices are represented by bold lower-case and upper-case letters, respectively. Let \mathbf{X}^T , \mathbf{X}^H and \mathbf{X}^{-1} denote the transpose, Hermitian, and inverse of matrix \mathbf{X} . $\|\cdot\|$ and $|\cdot|$ indicate the Euclidean norm and the absolute value. $\mathbb{E}(\cdot)$ denotes statistical expectation and $\langle \mathbf{a}, \mathbf{b} \rangle$ represents the inner product of two vectors.

II. SYSTEM MODEL

We consider the downlink¹ transmission in heterogeneous cellular networks (HCNs) composed of overlapping macro and pico cells [15]. As shown in Fig. 1, macro and pico base stations (MBSs and PBSs) are equipped with N_{T_1} and N_{T_0} antennas, whereas macro user equipment (MUE) and PUE are equipped with N_{R_1} and $N_{R_0} > 1$ antennas, respectively. Since mobile stations/devices are subject to severer restrictions in cost and hardware, than a base station (BS), the BS is assumed to have no less antennas than a UE, i.e., $N_{T_i} \geq N_{R_i}$ where $i = 0, 1$. The radio range, d , of a picocell is known to be 300m or less, whereas the radius, D , of a macrocell is around 3000m [15]. Let \mathbf{x}_1 and \mathbf{x}_0 denote the transmit data vectors from MBS and PBS to their serving subscribers, respectively. $\mathbb{E}(\|\mathbf{x}_1\|^2) = \mathbb{E}(\|\mathbf{x}_0\|^2) = 1$ holds. For clarity of exposition, our design begins with the assumption of beam-forming (BF), i.e., only one data stream is sent from MBS to

¹Since Tx-side cooperation is always difficult or even unavailable in the uplink scenario, we assume neither IS nor DIS is applicable for the uplink transmission.

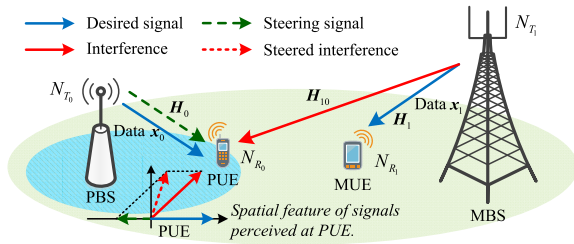


FIGURE 1. System model.

MUE (or from PBS to PUE). Then, x_1 and x_0 become scalars x_1 and x_0 . We will generalize this to multiple data streams sent from MBS and PBS in Section IV. We use P_1 and P_0 to denote the transmit power of MBS and PBS, respectively. Let $\mathbf{H}_0 \in \mathbb{C}^{N_{R_0} \times N_{T_0}}$ and $\mathbf{H}_1 \in \mathbb{C}^{N_{R_1} \times N_{T_1}}$ be the channel matrices from MBS to MUE and from PBS to PUE, respectively, whereas that from MBS to PUE is denoted by $\mathbf{H}_{10} \in \mathbb{C}^{N_{R_0} \times N_{T_1}}$. We adopt a spatially uncorrelated Rayleigh flat fading channel model to model the elements of the above matrices as independent and identically distributed zero-mean unit-variance complex Gaussian random variables. We assume that all users experience block fading, i.e., channel parameters remain constant in a block consisting of several successive time slots and vary randomly between blocks. Each user can accurately estimate CSI w.r.t. its desired and undesired Txs and feed it back to the associated BS via a low-rate error-free link. We assume reliable links for the delivery of CSI and signaling. The delivery delay is negligible relative to the time scale on which the channel state varies.

As mobile data traffic has increased significantly in recent years, network operators have strong preference of open access to offload users' traffic from heavily loaded macrocells to other infrastructures such as picocells [15], [16]. Following this trend, we assume each PBS operates in an open mode, i.e., users in the coverage of a PBS are allowed to access it. The transmission from MBS to MUE will interfere with the intended transmission from PBS at PUE. Nevertheless, due to the limited coverage of a picocell, PBS will not cause too much interference to MUE, and is thus omitted in this paper. As a result, the interference shown in Fig. 1 is asymmetric.

Since picocells are deployed to improve the capacity and coverage of existing cellular systems, each picocell has subordinate features as compared to the macrocell, and hence the macrocell transmission is given priority over the picocell's transmission. Specifically, MBS will not adjust its transmission for pico-users. However, we assume that PBS can acquire the information of x_1 via inter-BS collaboration; this is easy to achieve because PBS and MBS are deployed by the same operator [17]. With such information, DIS can be implemented to adjust the disturbance in a proper direction at PUE. One should note that the system model we adopted is also known as the cognitive Z-interference channel (ZIC) [18]. The ZIC where one transmitter-receiver pair is interference-free, is an interference channel (IC) [19].

Cognitive ZIC is characterized by the fact that the transmitter of the pair that suffers from interference is cognitive [18], i.e., being capable of acquiring the knowledge of interference. Since the transmission from MBS to MUE depends only on \mathbf{H}_1 and is free from interference, we only focus on the pico-users' transmission performance.

Although we take HCN as an example to design our scheme, it should be noticed that other types of network as long as they are featured as 1) collaboration between the interfering Tx and victim Tx is available, and 2) the interference topology is asymmetric, our scheme is applicable.

III. DYNAMIC INTERFERENCE STEERING

As mentioned earlier, by generating a duplicate of the interference and sending it along with the desired signal, the interference could be steered in the direction orthogonal to the desired signal at the intended PUE with orthogonal-IS. However, the tradeoff between the benefit of interference steering and its power cost has not been considered before. That is, under a transmit power constraint, the more power consumed for IS, the less power will be available for the intended signal's transmission. To remedy this deficiency, we propose a novel IM scheme called *dynamic interference steering* (DIS). By intelligently determining the strength of steering signal, the original interference is adjusted in an appropriate direction. DIS balances the transmit power used for generating the steering signal and that for the desired signal's transmission.

A. SIGNAL PROCESSING OF DIS

As mentioned above, since the macrocell receives higher priority than picocells, MBS will not adjust its transmission for pico-users. In what follows, we use $N_{T_i} = N_{R_i} \geq 2$ where $i = 0, 1$ as an example, but our scheme can be easily extended to the case of $N_{T_i} \geq N_{R_i}$.

Due to path loss, the mixed signal received at PUE can be expressed as:

$$\mathbf{r}_0 = \sqrt{P_0 10^{-0.1L_0}} \mathbf{H}_0 \mathbf{p}_0 x_0 + \sqrt{P_1 10^{-0.1L_{10}}} \mathbf{H}_{10} \mathbf{p}_1 x_1 + \mathbf{n}_0 \tag{1}$$

where the column vectors \mathbf{p}_0 and \mathbf{p}_1 represent the precoders for data symbols x_0 and x_1 sent from PBS and MBS, respectively. The first term on the right hand side (RHS) of Eq. (1) is the desired signal, the second term denotes the interference from MBS, and \mathbf{n}_0 represents for the additive white Gaussian noise (AWGN) with zero-mean and variance σ_n^2 . The path loss from MBS and PBS to a PUE is modeled as $L_{10} = 128.1 + 37.6 \log_{10}(\eta_{10}/10^3)$ dB and $L_0 = 38 + 30 \log_{10}(\eta_0)$ dB, respectively [20], where the variable $\eta_{(\cdot)}$, measured in meters (m), is the distance from the transmitter to the receiver.

The estimated signal at PUE after post-processing can be written as $\tilde{r}_0 = \mathbf{f}_0^H \mathbf{r}_0$ where \mathbf{f}_0 denotes the receive filter. Recall that the picocell operates in an open mode, and a MUE in the area covered by PBS will become a PUE and

then be served by the PBS. The interference model shown in Fig. 1 has an asymmetric feature in which only the interference from MBS to PUE is considered. Moreover, since the macrocell is given priority over the picocells, MBS will not adjust its transmission for the PUEs, and hence transmit packets to MUE based only on \mathbf{H}_1 . Here we adopt the singular value decomposition (SVD) based BF transmission, but can also use other types of pre- and post-processing. Applying SVD to \mathbf{H}_i ($i = 0, 1$), we get $\mathbf{H}_i = \mathbf{U}_i \Sigma_i \mathbf{V}_i^H$. We then employ $\mathbf{p}_i = \mathbf{v}_i^{(1)}$ and $\mathbf{f}_i = \mathbf{u}_i^{(1)}$, where $\mathbf{v}_i^{(1)}$ and $\mathbf{u}_i^{(1)}$ are the first column vectors of the right and left singular matrices (\mathbf{V}_i and \mathbf{U}_i), respectively, both of which correspond to the principal eigen-mode of \mathbf{H}_i .

From Fig. 1 one can see that the strengths of desired signal and interference at PUE depend on the network topology, differences of transmit power at PBS and MBS, as well as channel conditions. All of these factors affect the effectiveness of IM. For clarity of presentation, we define $P_0^e = P_0 10^{-0.1L_0}$, $P_1^e = P_1 10^{-0.1L_{10}}$, where P_0^e and P_1^e indicate the transmit power of PBS and MBS incorporated with the path loss perceived by PUE. With this definition, consideration of various network topologies and transmit power differences can be simplified to P_0^e and P_1^e . In what follows, we first present the basic principle of orthogonal-IS (OIS), and then elaborate on the design of dynamic-IS (DIS) where we provide the existence and calculation of the optimal steering signal.

With OIS, PBS acquires interference information, including data and CSI, from MBS via inter-BS collaboration and by PUE's estimation and feedback, respectively. PBS then generates a duplicate of the interference and sends it along with the desired signal. The former is used for interference steering at PUE, whereas the latter carries the payload. The received signal at PUE then becomes:

$$\mathbf{r}_0 = \sqrt{P_0^e - P_{OIS}^e} \mathbf{H}_0 \mathbf{p}_0 x_0 + \sqrt{P_1^e} \mathbf{H}_{10} \mathbf{p}_1 x_1 + \sqrt{P_{OIS}^e} \mathbf{H}_0 \mathbf{p}_{OIS} x_1 + \mathbf{n}_0 \quad (2)$$

where $P_{OIS}^e = P_{OIS} 10^{-0.1L_0}$. P_{OIS} represents the power overhead of OIS, and \mathbf{p}_{OIS} is the precoder for the steering signal. We first define the directions of the desired signal and the original interference combined with the steering signal as $\mathbf{d}_s = \frac{\mathbf{H}_0 \mathbf{p}_0}{\|\mathbf{H}_0 \mathbf{p}_0\|}$ and $\mathbf{d}_{i+s_r} = \frac{\sqrt{P_1^e} \mathbf{H}_{10} \mathbf{p}_1 + \sqrt{P_{OIS}^e} \mathbf{H}_0 \mathbf{p}_{OIS}}{\|\sqrt{P_1^e} \mathbf{H}_{10} \mathbf{p}_1 + \sqrt{P_{OIS}^e} \mathbf{H}_0 \mathbf{p}_{OIS}\|}$, respectively. Then, the original interference should be steered in the direction orthogonal to the desired signal by letting $\langle \mathbf{d}_s, \mathbf{d}_{i+s_r} \rangle = 0$. As shown in Fig. 2(a), both the disturbance (\mathbf{i}) and the steering signal (\mathbf{s}_r), can be decomposed into an in-phase component and a quadrature component, denoted by the superscripts In and Q , respectively, w.r.t. the intended transmission \mathbf{s} , i.e., $\mathbf{i} = \mathbf{i}^{In} + \mathbf{i}^Q$ and $\mathbf{s}_r = \mathbf{s}_r^{In} + \mathbf{s}_r^Q$. When $\mathbf{s}_r^{In} = -\mathbf{i}^{In}$, OIS is realized. Furthermore, since the length of a vector indicates the signal's strength, OIS minimum power overhead is achieved when $\mathbf{s}_r = \mathbf{s}_r^{In}$, i.e., $\mathbf{s}_r^Q = \mathbf{0}$. Hence, in order to reduce power cost, we let $\mathbf{s}_r^Q = \mathbf{0}$. It can be easily seen that $\mathbf{i}^{In} = \sqrt{P_1^e} \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1$ where $\mathbf{P} = \mathbf{d}_s (\mathbf{d}_s^T \mathbf{d}_s)^{-1} \mathbf{d}_s^T$ denotes the projection matrix. To implement OIS, the steering

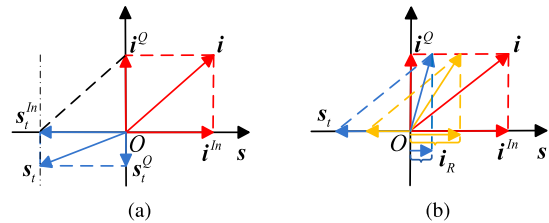


FIGURE 2. Illustrations of OIS and DIS. (a) An illustration of OIS. (b) An illustration of DIS.

signal should satisfy $\sqrt{P_{OIS}^e} \mathbf{H}_0 \mathbf{p}_{OIS} = -\sqrt{P_1^e} \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1$. This equation can be decomposed into $\mathbf{H}_0 \mathbf{p}_{OIS} = -\alpha \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1$ and $P_{OIS}^e = \beta P_1^e$ where $\alpha \sqrt{\beta} = 1$, so that we can get $\mathbf{p}_{OIS} = -\alpha \mathbf{H}_0^{-1} \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1$. Note that $\|\mathbf{p}_{OIS}\| = 1$ is not guaranteed, i.e., \mathbf{p}_{OIS} affects the power cost of OIS.

When $N_{T_i} > N_{R_i}$ ($i = 0, 1$), the inverse of \mathbf{H}_0 should be replaced by its Moore-Penrose pseudo-inverse. The mechanism can then be generalized. In addition, when the interference is too strong, P_0 may not be sufficient for OIS, in such a case, we can simply switch to the non-interference management (non-IM) mode, e.g., matched filtering (MF) at the victim receiver, or other IM schemes with less or no transmit power consumption such as zero-forcing reception.

By adopting $\mathbf{f}_0 = \mathbf{u}_0^{(1)}$ as the receive filter, the SE of PUE with OIS can be computed as:

$$c_0^{OIS} = \log_2 \left\{ 1 + \frac{(P_0^e - P_{OIS}^e) [\lambda_0^{(1)}]^2}{\sigma_n^2} \right\}, \quad (3)$$

where $\lambda_0^{(1)}$ is the largest singular value of \mathbf{H}_0 , indicating the amplitude gain of the principal spatial sub-channel.

From Eq. (3), we can see that although the original interference is steered into the orthogonal direction of the desired signal and the disturbance to the intended transmission is completely eliminated, it accompanies a transmit power loss, P_{OIS}^e , degrading the received desired signal strength. One can then raise a question: "is the orthogonal-IS always necessary/worthwhile?" To answer this question, we may adjust both the direction and strength of the steering signal adaptively to implement dynamic IS. Note, however, that in order to minimize the transmit power overhead, the steering signal should be opposite to the projection of interference on the desired signal. Thus, only the strength of steering signal needs to be adjusted. In what follows, we generalize the OIS to DIS by introducing a coefficient $\rho \in (0, 1]$ called the *steering factor*, representing the portion of in-phase component of the disturbance w.r.t. the desired signal to be mitigated. When $\rho = 1$, orthogonal-IS is realized, while DIS approaches non-IM as $\rho \rightarrow 0$. Without ambiguity, we adopt \mathbf{s}_r to represent the dynamic steering signal. Then, we have $\mathbf{s}_r = -\rho \mathbf{i}^{In} = -\rho \sqrt{P_1^e} \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1$. As illustrated by Fig. 2(b), when $\rho < 1$, interference \mathbf{i} is steered into a direction not orthogonal to \mathbf{d}_s , i.e., \mathbf{i}^{In} is not completely eliminated. Then, the steered interference becomes $\mathbf{i} + \mathbf{s}_r = (1 - \rho) \mathbf{i}^{In} + \mathbf{i}^Q$, whose projection on \mathbf{d}_s is non-zero, i.e., provided that $\rho < 1$, a residual interference, expressed as $\mathbf{i}_R = \mathbf{i}^{In} + \mathbf{s}_r = (1 - \rho) \mathbf{i}^{In}$, exists.

Similarly to the discussion about OIS, to implement DIS, the following equation should hold:

$$\sqrt{P_{DIS}^e} \mathbf{H}_0 \mathbf{p}_{DIS} = -\rho \sqrt{P_1^e} \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1. \quad (4)$$

For clarity of exposition, we normalize the precoder so that the direction and strength requirements for DIS could be decoupled from each other. Then, the expression of DIS implementation is given as:

$$\begin{cases} \mathbf{p}_{DIS} = -\mathbf{H}_0^{-1} \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1 / \|\mathbf{H}_0^{-1} \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1\| \\ P_{DIS}^e = \rho^2 P_1^e \|\mathbf{H}_0^{-1} \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1\|^2, \end{cases} \quad (5)$$

where \mathbf{p}_{DIS} is the precoder for steering signal and P_{DIS}^e denotes the power overhead for DIS at PBS, i.e., P_{DIS} , incorporated with path loss $10^{-0.1L_0}$.

The received signal at PUE with DIS is then

$$\mathbf{r}_0 = \sqrt{P_0^e - P_{DIS}^e} \mathbf{H}_0 \mathbf{p}_0 x_0 + \sqrt{P_1^e} (1 - \rho) \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1 x_1 + \mathbf{n}_0, \quad (6)$$

where the second term on the RHS of Eq. (6) indicates the residual interference, which is the in-phase component of the original interference $\sqrt{P_1^e} \mathbf{H}_{10} \mathbf{p}_1 x_1$ combined with the steering signal $\rho \sqrt{P_{DIS}^e} \mathbf{H}_0 \mathbf{p}_{DIS} x_1$ w.r.t. the desired transmission.

By employing $\mathbf{f}_0 = \mathbf{u}_0^{(1)}$ as the receive filter, the achievable SE of PUE adopting DIS can be calculated as:

$$c_0^{DIS} = \log_2 \left\{ 1 + \frac{(P_0^e - P_{DIS}^e) [\lambda_0^{(1)}]^2}{\sigma_n^2 + I_R} \right\}, \quad (7)$$

where $I_R = P_1^e \|\mathbf{f}_0^H (1 - \rho) \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1\|^2$ denotes the strength of residual interference after post-processing at PUE.

Based on the above discussion, it can be easily seen that $I_R = 0$ when $\rho = 1$, i.e., DIS becomes OIS. So, DIS includes OIS as a special case, making it more general.

B. OPTIMIZATION OF STEERING FACTOR ρ

In what follows, we will discuss the existence of the optimal ρ , denoted by ρ^* , with which PUE's SE can be maximized with limited P_0 . Based on the Shannon's equation, we can instead optimize the signal-to-interference-plus-noise ratio (SINR) of PUE, denoted by φ_0 .

Substituting Eq. (5) into Eq. (7), we can obtain φ_0 as:

$$\begin{aligned} \varphi_0 &= \frac{(P_0^e - \rho^2 P_1^e \|\mathbf{H}_0^{-1} \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1\|^2) [\lambda_0^{(1)}]^2}{(1 - \rho)^2 P_1^e \|\mathbf{f}_0^H \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1\|^2 + \sigma_n^2} \\ &= \frac{P_0^e [\lambda_0^{(1)}]^2 - \rho^2 P_1^e \|\mathbf{g}\|^2 [\lambda_0^{(1)}]^2}{\rho^2 P_1^e |\chi|^2 - (2\rho - 1) P_1^e |\chi|^2 + \sigma_n^2}, \end{aligned} \quad (8)$$

where $\mathbf{g} = \mathbf{H}_0^{-1} \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1$ and $\chi = \mathbf{f}_0^H \mathbf{P} \mathbf{H}_{10} \mathbf{p}_1$.

Eq. (8) can be simplified as:

$$\varphi_0 = \frac{A - \rho^2 B}{C - \rho D + \rho^2 E} \quad (9)$$

where $A = P_0^e [\lambda_0^{(1)}]^2$, $B = P_1^e \|\mathbf{g}\|^2 [\lambda_0^{(1)}]^2$, $C = P_1^e |\chi|^2 + \sigma_n^2$, $D = 2P_1^e |\chi|^2$ and $E = P_1^e |\chi|^2$. Note that all of these coefficients are positive.

Next, we elaborate on the existence of ρ^* under the P_0 constraint, with which φ_0 is maximized. By substituting \mathbf{g} into Eq. (5), we can see that when $P_0^e > P_1^e \|\mathbf{g}\|^2$, PBS has enough power to steer the interference into the direction orthogonal to the desired signal, i.e., OIS is achievable. Otherwise, when $P_0^e \leq P_1^e \|\mathbf{g}\|^2$, the maximum of ρ , denoted by ρ_{max} is limited by the PBS's transmit power. In such a case, $\sqrt{\frac{P_0^e}{P_1^e \|\mathbf{g}\|^2}}$ is the maximum portion of the in-phase component of the original interference w.r.t. the desired signal that can be mitigated with P_0 . Based on the above discussion, we have $\rho^* \in (0, \rho_{max}]$ where $\rho_{max} = \min\left(1, \sqrt{\frac{P_0^e}{P_1^e \|\mathbf{g}\|^2}}\right)$. In what follows, we first prove the solvability of ρ^* , and then show the quality of the resulting solution(s).

By computing the derivative of φ_0 to ρ and setting it to 0, we get:

$$\frac{\frac{BD}{2} \rho^2 - (BC + AE)\rho + \frac{AD}{2}}{\frac{1}{2}(C - D\rho + \rho^2 E)^2} = 0. \quad (10)$$

Since the denominator cannot be 0, we only need to solve

$$\frac{BD}{2} \rho^2 - (BC + AE)\rho + \frac{AD}{2} = 0 \quad (11)$$

which is a quadratic equation with one unknown. Let's define $\Delta = (BC + AE)^2 - ABD^2$. Since $\Delta = (BC + AE + \sqrt{ABD})(BC + AE - \sqrt{ABD})$, and $BC + AE + \sqrt{ABD}$ is positive, we only need to show that $BC + AE - \sqrt{ABD} > 0$. The proof is given below by Eq. (12), as shown at the bottom of this page.

Based on the above discussion, we can obtain two solutions $\rho_{\pm}^* = \frac{(BC+AE) \pm \sqrt{\Delta}}{BD}$. We then need to verify the qualification of the two solutions ρ_{\pm}^* . We first investigate the feasibility of the larger solution $\rho_+^* = \frac{(BC+AE) + \sqrt{\Delta}}{BD}$. The first term of ρ_+^* can be rewritten as:

$$\begin{aligned} \frac{BC + AE}{BD} &> \frac{P_1^e \|\mathbf{g}\|^2 [\lambda_0^{(1)}]^2 P_1^e |\chi|^2 + P_0^e [\lambda_0^{(1)}]^2 P_1^e |\chi|^2}{2P_1^e \|\mathbf{g}\|^2 [\lambda_0^{(1)}]^2 P_1^e |\chi|^2} \\ &= \frac{1}{2} + \frac{P_0^e}{2P_1^e \|\mathbf{g}\|^2}. \end{aligned} \quad (13)$$

$$\begin{aligned} BC + AE - \sqrt{ABD} &= P_1^e \|\mathbf{g}\|^2 [\lambda_0^{(1)}]^2 (\sigma_n^2 + P_1^e |\chi|^2) + P_0^e [\lambda_0^{(1)}]^2 P_1^e |\chi|^2 - 2\sqrt{P_0^e [\lambda_0^{(1)}]^2 P_1^e \|\mathbf{g}\|^2 [\lambda_0^{(1)}]^2 P_1^e |\chi|^2} \\ &> (P_1^e)^2 \|\mathbf{g}\|^2 [\lambda_0^{(1)}]^2 |\chi|^2 + P_0^e P_1^e [\lambda_0^{(1)}]^2 |\chi|^2 - 2(P_0^e)^{\frac{1}{2}} (P_1^e)^{\frac{3}{2}} [\lambda_0^{(1)}]^2 \|\mathbf{g}\| |\chi|^2 \\ &= P_1^e [\lambda_0^{(1)}]^2 |\chi|^2 \left[(P_1^e)^{\frac{1}{2}} \|\mathbf{g}\| - (P_0^e)^{\frac{1}{2}} \right]^2 \geq 0. \end{aligned} \quad (12)$$

The second term of ρ_+^* is:

$$\begin{aligned} \frac{\sqrt{\Delta}}{BD} &= \sqrt{\frac{(BC + AE)^2}{B^2D^2} - \frac{A}{B}} \\ &> \sqrt{\left(\frac{1}{2} + \frac{P_0^e}{2P_1^e\|\mathbf{g}\|^2}\right)^2 - \frac{P_0^e}{P_1^e\|\mathbf{g}\|^2}} \\ &= \left|\frac{1}{2} - \frac{P_0^e}{2P_1^e\|\mathbf{g}\|^2}\right|. \end{aligned} \quad (14)$$

Then, we can get:

$$\begin{aligned} \rho_+^* &> \left(\frac{1}{2} + \frac{P_0^e}{2P_1^e\|\mathbf{g}\|^2}\right) + \left|\frac{1}{2} - \frac{P_0^e}{2P_1^e\|\mathbf{g}\|^2}\right| \\ &= \begin{cases} 1, & \frac{P_0^e}{2P_1^e\|\mathbf{g}\|^2} \leq \frac{1}{2} \\ \frac{P_0^e}{P_1^e\|\mathbf{g}\|^2}, & \frac{P_0^e}{2P_1^e\|\mathbf{g}\|^2} > \frac{1}{2}. \end{cases} \end{aligned} \quad (15)$$

Note that $\rho_+^* \leq 1$, and hence $\frac{P_0^e}{2P_1^e\|\mathbf{g}\|^2}$ should not be less than or equal to $\frac{1}{2}$. However, when $\frac{P_0^e}{2P_1^e\|\mathbf{g}\|^2} > \frac{1}{2}$, $\rho_+^* > \frac{P_0^e}{2P_1^e\|\mathbf{g}\|^2}$ is equivalent to $\rho_+^* > 1$. As a result, $\rho_+^* \notin (0, \rho_{max}]$ where $\rho_{max} = \min\left(1, \sqrt{\frac{P_0^e}{P_1^e\|\mathbf{g}\|^2}}\right)$, i.e., ρ_+^* is not acceptable.

As for $\rho_-^* = \frac{(BC+AE)-\sqrt{\Delta}}{BD}$, since $ABD^2 > 0$, $BC + AE > \sqrt{\Delta}$ holds, thus justifying $\rho_-^* > 0$. Then, we prove $\rho_-^* < \rho_{max}$ as follows. First, we define a function $g(C) = (BC + AE) - \sqrt{\Delta}$. Since the derivative of $g(C)$ to C , $g'(C) = \frac{B\sqrt{\Delta}-B(BC+AE)}{\sqrt{\Delta}} < 0$, $g(C)$ is a monotonically decreasing function of variable C . Let $C' = P_{T_1}^e|\chi|^2$, then $C = \sigma_n^2 + P_1^e|\chi|^2 > C'$, thus leading to $g(C) < g(C')$. Similarly to the derivations of Eqs. (13)–(15), we get:

$$\rho_-^* = \frac{g(C)}{BD} < \frac{g(C')}{BD} = \begin{cases} 1, & \frac{P_0^e}{2P_1^e\|\mathbf{g}\|^2} \geq \frac{1}{2} \\ \frac{P_0^e}{P_1^e\|\mathbf{g}\|^2}, & \frac{P_0^e}{2P_1^e\|\mathbf{g}\|^2} < \frac{1}{2}. \end{cases} \quad (16)$$

Eq. (16) is equivalent to $\rho_-^* < \rho_{max}^2 = \min\left(1, \frac{P_0^e}{P_1^e\|\mathbf{g}\|^2}\right)$. Since $0 < \rho_{max}^2 \leq 1$, $\rho_{max}^2 \leq \rho_{max}$ holds, thus proving $\rho_-^* < \rho_{max}$.

Finally, we prove that ρ_-^* could achieve the maximum φ_0 . Since it can be proved that $h(\rho) = \frac{BD}{2}\rho^2 - (BC + AE)\rho + \frac{AD}{2}$ is a monotonically decreasing function of variable $\rho \in (0, \rho_{max}]$, when $0 < \rho < \rho_-^*$, we get $h(\rho) > h(\rho_-^*) = 0$. Similarly, when $\rho_-^* < \rho < \rho_{max}$, $h(\rho) < 0$ can be derived. Thus, ρ_-^* corresponds to the maximum φ_0 . The optimal steering factor is calculated as $\rho^* = \rho_-^*$.

IV. GENERALIZATION OF DIS

A. GENERALIZED NUMBER OF INTERFERENCES

So far, we have assumed that the MBS sends a single data stream to MUE, i.e., only one interference is imposed on the PUE. When multiple desired signals are sent from a MBS, the proposed DIS can be extended as follows.

Since picocells are deployed within the coverage of a macrocell, interferences from the other MBSs are negligible. For clarity of presentation, Fig. 3 shows a two-interference situation as an example, where \mathbf{i}_1 and \mathbf{i}_2 are the interferences. Only one desired signal is considered.

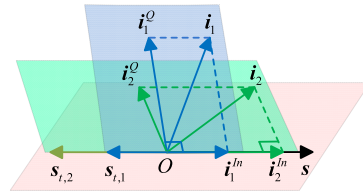


FIGURE 3. Generalization of the number of interferences.

As shown in this figure, each interference can be decomposed into an in-phase component and a quadrature component w.r.t. the desired signal. Then, DIS can be applied to each interference separately, following the processing as described in Section III.

For $N > 1$ interferences, the received signal at the victim PUE with DIS can be expressed as:

$$\begin{aligned} \mathbf{r}_0 &= \sqrt{P_0^e - \sum_{n=1}^N P_{DIS,n}^e} \mathbf{H}_0 \mathbf{p}_0 x_0 + \sum_{n=1}^N \sqrt{P_{1,n}^e} \mathbf{H}_{10} \mathbf{p}_{1,n} x_{1,n} \\ &+ \sum_{n=1}^N \sqrt{P_{DIS,n}^e} \mathbf{H}_0 \mathbf{p}_{DIS,n} x_{1,n} + \mathbf{n}_0, \end{aligned} \quad (17)$$

where $\mathbf{x}_1 = [x_{1,1}, \dots, x_{1,n}, \dots, x_{1,N}]$ is the transmit data vector of MBS. The transmission of $x_{1,n}$ causes the interference term $\sqrt{P_{1,n}^e} \mathbf{H}_{10} \mathbf{p}_{1,n} x_{1,n}$ to the PUE, in which $P_{1,n}^e$ is the transmit power for $x_{1,n}$, i.e., $P_{1,n}$, incorporated with the path loss $10^{-0.1L_{10}}$. $\mathbf{p}_{1,n}$ denotes the precoder for $x_{1,n}$. $\sum_{n=1}^N P_{1,n}^e = P_1^e$ holds. PBS generates a duplicate of this interference with the power overhead $P_{DIS,n}$ where $P_{DIS,n}^e = P_{DIS,n} 10^{-0.1L_0}$ and the precoder $\mathbf{p}_{DIS,n}$ so as to adjust the interference to an appropriate direction at the victim PUE.

Similarly to the derivation of Eq. (5), we can obtain the DIS design for the n -th interfering component as:

$$\begin{cases} \mathbf{p}_{DIS,n} = -\mathbf{H}_0^{-1} \mathbf{P} \mathbf{H}_{10} \mathbf{p}_{1,n} / \|\mathbf{H}_0^{-1} \mathbf{P} \mathbf{H}_{10} \mathbf{p}_{1,n}\| \\ P_{DIS,n}^e = \rho_n^2 P_{1,n}^e \|\mathbf{H}_0^{-1} \mathbf{P} \mathbf{H}_{10} \mathbf{p}_{1,n}\|^2, \end{cases} \quad (18)$$

where \mathbf{P} represents the projection matrix depending only on the spatial feature of the desired transmission, with which we can calculate the in-phase component of the interference caused by the transmission of $x_{1,n}$ w.r.t. the intended signal. ρ_n is the steering factor for the steering signal carrying $x_{1,n}$.

One should note that when there are multiple interferences, it is difficult to determine the optimal steering factors for all the interfering components. However, we can allocate a power budget $P_{0,n}$, satisfying $\sum_{n=1}^N P_{0,n} < P_0$, to each interference, and then by applying DIS to each disturbance under its power budget constraint, a vector of n sub-optimal steering factors is achieved. $P_{0,n}$ can be assigned with the

same value or based on the strength of interferences. The achievable SE of the PUE can then be calculated as:

$$c_0^{DIS} = \log_2 \left\{ 1 + \frac{\left(P_0^e - \sum_{n=1}^N P_{DIS,n}^e \right) [\lambda_0^{(1)}]^2}{\sigma_n^2 + \sum_{n=1}^N I_{R,n}} \right\} \quad (19)$$

where $I_{R,n} = P_1^e \|\mathbf{f}_0^H (1 - \rho_n) \mathbf{P} \mathbf{H}_{10} \mathbf{p}_{1,n}\|^2$ ($n = 1, \dots, N$) is the strength of the n -th residual interference to the intended transmission. $\mathbf{f}_0 = \mathbf{u}_0^{(1)}$ is the receive filter for data x_0 .

To further elaborate on the extension of the proposed scheme, we provide below an algorithm for $N = 2$, with which the optimal ρ_n can be determined, maximizing the system SE. For simplicity, we use the function $f(\rho_1, \rho_2)$ to denote c_0^{DIS} under $N = 2$ in the following description. The above algorithm can be extended to the case of $N > 2$. Due to space limitation, we do not elaborate on this any further in this paper.

Algorithm 1

- 1: Take the derivative of $f(\rho_1, \rho_2)$ to ρ_1 and ρ_2 respectively, to obtain $f'_{\rho_n}(\rho_1, \rho_2) = \frac{\partial f(\rho_1, \rho_2)}{\partial \rho_n}$ where $n = 1, 2$.
- 2: Compute the stationary point $(\tilde{\rho}_1, \tilde{\rho}_2)$ of $f(\rho_1, \rho_2)$ by solving the equation $f'_{\rho_n}(\rho_1, \rho_2) = 0$. We define set Φ consisting of $(\tilde{\rho}_1, \tilde{\rho}_2)$.
- 3: Calculate the second-order derivative of $f(\rho_1, \rho_2)$ at the stationary point $(\tilde{\rho}_1, \tilde{\rho}_2)$, i.e., $f''_{\rho_1, \rho_2}(\tilde{\rho}_1, \tilde{\rho}_2) = \frac{\partial^2 f(\rho_1, \rho_2)}{\partial \rho_1 \partial \rho_2} \Big|_{\rho_1=\tilde{\rho}_1, \rho_2=\tilde{\rho}_2}$. For clarity of exposition, we define $\mathcal{A} = f''_{\rho_1, \rho_2}(\tilde{\rho}_1, \tilde{\rho}_2)$. Similarly, we define variables $\mathcal{B} = f''_{\rho_1, \rho_1}(\tilde{\rho}_1, \tilde{\rho}_2)$ and $\mathcal{C} = f''_{\rho_2, \rho_2}(\tilde{\rho}_1, \tilde{\rho}_2)$.
- 4: Check whether the stationary point is an extreme point or not. If $\mathcal{A}^2 - \mathcal{B}\mathcal{C} < 0$ and $\mathcal{B} < 0$, $(\tilde{\rho}_1, \tilde{\rho}_2)$ is an extreme point; otherwise not. We can obtain the set of extreme points and the value of $f(\rho_1, \rho_2)$ at each extreme point correspondingly. We define the extreme value set as Ω .
- 5: Since both ρ_1 and ρ_2 range from 0 to 1, i.e., $\rho_n \in (0, 1]$, the maximum value of $f(\rho_1, \rho_2)$ may exist at the boundary points. We define set $\mathcal{F} = f(\rho_1, \rho_2) \Big|_{\rho_n \in \{0, 1\}, n=1, 2}$.
- 6: Determine the optimal (ρ_1^*, ρ_2^*) outputting the maximum value of $f(\rho_1, \rho_2) \Big|_{\rho_1=\rho_1^*, \rho_2=\rho_2^*}$ by searching all the elements in sets Ω and \mathcal{F} .

B. GENERALIZED NUMBER OF DESIRED DATA STREAMS

We now generalize the number of desired signals, denoted by M , sent from PBS to its PUE. For clarity of exposition, we take $M = 2$ and the number of interferences $N = 1$ as an example as shown in Fig. 4. However, this discussion can be readily extended to more general parameter settings. As can be seen from the figure, the interference \mathbf{i} forms a plane with each of the desired signals \mathbf{s}_m where $m = 1, 2$. The projection of \mathbf{i} onto \mathbf{s}_m is denoted by \mathbf{i}_m^{in} (in-phase component), whereas the quadrature component is \mathbf{i}_m^O . By applying DIS to each $(\mathbf{i}, \mathbf{s}_m)$ pair, a set of steering signals can be determined.

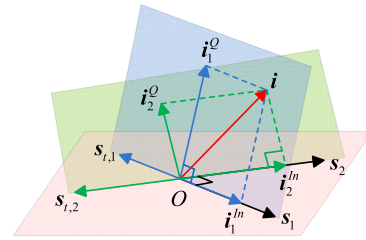


FIGURE 4. Generalization of the number of desired signals.

Since multiple data streams are sent from PBS to PUE via mutually orthogonal eigenmodes/subchannels, and the steering signal is opposite to the spatial feature of the desired transmission it intends to protect, an arbitrary steering signal, say $\mathbf{s}_{t,m}$, is orthogonal to any of the other desired signals $\mathbf{s}_{m'}$ where $m' \neq m$. Hence, no additional interference will be created by the steering signal $\mathbf{s}_{t,m}$.

Based on the above discussion, the number of steering signals is equal to that of the desired data streams M . Thus, the mixed signal received at PUE can be expressed as:

$$\mathbf{r}_0 = \sum_{m=1}^M \sqrt{P_{0,m}^e - P_{DIS,m}^e} \mathbf{H}_0 \mathbf{p}_{0,m} x_{0,m} + \sqrt{P_1^e} \mathbf{H}_{10} \mathbf{p}_1 x_1 + \sum_{m=1}^M \sqrt{P_{DIS,m}^e} \mathbf{H}_0 \mathbf{p}_{DIS,m} x_1 + \mathbf{n}_0 \quad (20)$$

where $\mathbf{x}_0 = [x_{0,1}, \dots, x_{0,m}, \dots, x_{0,M}]$ is the transmit data vector of PBS. $P_{0,m}^e$ denotes the transmit power budget for $x_{0,m}$, i.e., $P_{0,m}$, incorporated with the path loss $10^{-0.1L_0}$. $P_{DIS,m}$ is the power cost for steering the interference away from the m -th desired signal, and $P_{DIS,m}^e = P_{DIS,m} 10^{0.1L_0}$. $\mathbf{p}_{0,m}$ and $\mathbf{p}_{DIS,m}$ represent the precoders for $x_{0,m}$ and its steering signal, respectively. Similarly to the derivation of Eqs. (5) and (18), the DIS solution for multi-desired-signal situation can be readily obtained. In such a case, the achievable SE of PUE can be expressed as:

$$c_0^{DIS} = \sum_{m=1}^M \log_2 \left\{ 1 + \frac{\left(P_{0,m}^e - P_{DIS,m}^e \right) [\lambda_0^{(m)}]^2}{\sigma_n^2 + I_{R,m}} \right\}. \quad (21)$$

$\lambda_0^{(m)}$ denotes the amplitude gain of the m -th desired transmission from PBS to PUE. The residual interference $I_{R,m} = P_1^e \|\mathbf{f}_{0,m}^H (1 - \rho_m) \mathbf{P}_m \mathbf{H}_{10} \mathbf{p}_1\|^2$ where $\mathbf{f}_{0,m} = \mathbf{u}_0^{(m)}$ is the receive filter for data $x_{0,m}$. The projection matrix $\mathbf{P}_m = \mathbf{d}_{s,m} (\mathbf{d}_{s,m}^T \mathbf{d}_{s,m})^{-1} \mathbf{d}_{s,m}^T$ where $\mathbf{d}_{s,m} = \frac{\mathbf{H}_0 \mathbf{p}_{0,m}}{\|\mathbf{H}_0 \mathbf{p}_{0,m}\|}$.

It should be noted that the optimal $P_{DIS,m}^e$, or equivalently ρ_m is dependent on $P_{0,m}^e$ where $\sum_{m=1}^M P_{0,m}^e = P_0^e$ holds. Thus, different power allocations will yield different DIS solutions. For example, P_0 can be equally allocated to the M intended data transmissions, or in terms of the quality of subchannels and/or the strength of interference imposed on each desired signal. Then, suboptimal performance w.r.t. the transmission from PBS to PUE is achieved. How to jointly determine the optimal $P_{0,m}$ and $P_{DIS,m}$ is our future work.

C. GENERALIZED NUMBER OF PBSS AND PUES

We discuss the generalization of the number of PBSSs deployed in the coverage of a macrocell and the number of PUEs served by each PBS. As mentioned before, PBSSs are installed by the network operator. Inter-pico-cell interference could therefore be effectively avoided by the operator’s planned deployment or resource allocation. Even when inter-pico-cell interference exists, our scheme can be directly applied by treating the interfering PBS as the MBS in this paper, and DIS is implemented at the PBS associated with the victim PUE. It should be noted that the proposed DIS is applicable to the scenario with asymmetric interferences. Otherwise, concurrent data transmissions should be scheduled or other schemes should be adopted to address the interference problem, which is beyond the scope of this paper. As for the multi-PUE case, each PUE can be assigned an exclusive channel so as to avoid co-channel interference, which is consistent with various types of wireless communication systems, such as WLANs. In summary, with an appropriate system design, the proposed DIS can be applied to the system with multiple PBSSs and PUEs.

V. EVALUATION

We evaluate the performance of the proposed mechanism using MATLAB. We set $d = 300\text{m}$, $D = 3000\text{m}$, $P_0 = 23\text{dBm}$ and $P_1 = 46\text{dBm}$ [15]. The path loss is set to $L_{10} = 128.1 + 37.6 \log_{10}(\eta_{10}/10^3)$ dB and $L_0 = 38 + 30 \log_{10}(\eta_0)$ dB where $\eta_0 \leq d$ and $\eta_{10} \leq D$. Since L_0 and L_{10} are dependent on the network topology, P_0^e ranges from -89dBm to 23dBm , whereas P_1^e varies between -100dBm to 46dBm . For clarity of presentation, we adopt $\bar{\gamma} = 10 \log_{10}(\gamma)$ where $\gamma = P_1^e/\sigma_n^2$. We also define $\xi = P_0^e/P_1^e$. Then, based on the above parameter settings, $\xi \in [-135, 123]$ dB. Note, however, that we obtained this result for extreme boundary situations, making its range too wide to be useful. Without specifications, the simulation is done under $N_{T_0} = N_{T_1} = N_{R_0} = 2$ antenna configuration. However, same conclusion can be drawn with various parameter settings. In practice, a PBS should not be deployed close to MBS and mobile users may select an access point based on the strength of reference signals from multiple access points. Considering this practice, we set $\xi \in [0.1, 100]$ in our simulation. There are M desired signals and N interferences. In the following simulation, when the power overhead of an IM scheme exceeds P_0 at the victim Tx, we simply switch to non-IM mode, i.e., matched filtering (MF) is employed by letting $\mathbf{f}_{0,m} = \mathbf{u}_0^{(m)}$, while the interference remains unchanged.

Fig. 5 shows two samples of the relationship between ρ and PUE’s achievable SE. The interference shown in Fig. 5(a) is relatively weak, and hence the transmit power of PBS is sufficient for OIS, i.e., ρ_{max} can be as large as 1. In Fig. 5(b), since the interference is strong, and hence, when $\rho > \rho_{max}$ where $\rho_{max} < 1$, there won’t be enough power for PBS to realize OIS. In such a case, we simply switch off IS and adopt non-IM. In both figures, the optimal ρ , denoted by ρ^* , is computed as in Section III, which corresponds

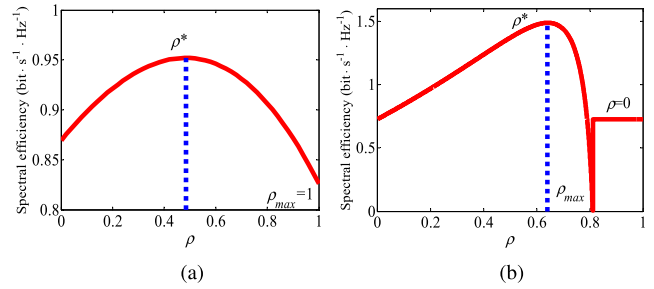


FIGURE 5. SE of PUE vs. ρ under $\bar{\gamma} = 0\text{dB}$, $M = N = 1$, and $\xi = 1$. (a) Weak interference. (b) Strong interference.

to the maximum SE. We can conclude from Fig. 5 that in order to better utilize the transmit power for both IS and data transmission, it is necessary to intelligently determine the appropriate strength of the steering signal, or equivalently, the direction into which the interference is steered.

Fig. 6 plots the PUE’s average SE versus ρ for different ξ . The average ρ^* , marked by pentagram, which corresponds to the PUE’s maximum SE, grows as ξ increases. When ξ gets too high, a large portion of interference is preferred to be mitigated. As shown in the figure, given $\xi = 100$, the average ρ^* is approximately 0.9. In addition, since the strength of the desired signal relative to the interference grows with an increase of ξ , the PUE’s SE performance improves with ρ .

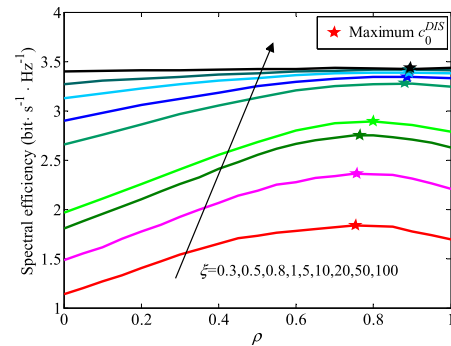


FIGURE 6. SE of PUE vs. ρ under $\bar{\gamma} = 5\text{dB}$, $M = N = 1$, and different ξ .

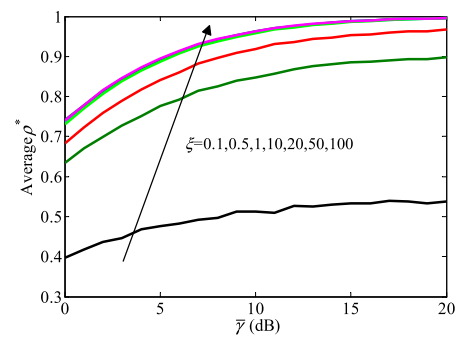


FIGURE 7. Average ρ^* vs. $\bar{\gamma}$ under $M = N = 1$, and different ξ .

Fig. 7 shows the average ρ^* versus $\bar{\gamma}$ under $M = N = 1$ and different ξ . As the figure shows, with fixed ξ , the average ρ^* grows with an increase of $\bar{\gamma}$. This is because the

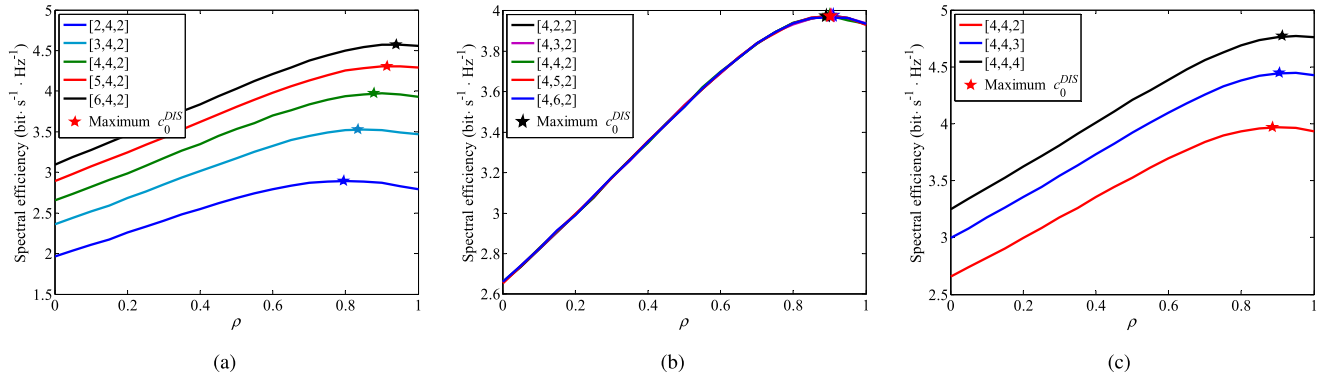


FIGURE 8. SE of PUE vs. ρ under $\bar{\gamma} = 5\text{dB}$, $\xi = 1$, $M = N = 1$, and different antenna settings. (a) N_{T_0} is variable. (b) N_{T_1} is variable. (c) N_{R_0} is variable.

interference gets stronger with increasing $\bar{\gamma}$, and hence, to achieve the maximum SE, ρ^* should increase, i.e., more interference imposing onto the desired signal should be mitigated. Given the same $\bar{\gamma}$, the average ρ^* grows with an increase of ξ , which is consistent with Fig. 6.

Fig. 8 plots PUE’s average SE along with ρ under different antenna settings. We use a general form $[N_{T_0} N_{T_1} N_{R_0}]$ to express the antenna configuration. In Fig. 8(a), N_{T_1} and N_{R_0} are fixed, and N_{T_0} varies from 2 to 6. Since the transmit array gain of the desired signal grows with an increase of N_{T_0} , meaning that the desired signal becomes stronger relative to the interference, and more interference can be eliminated by DIS with the same power overhead as N_{T_0} grows, both the average ρ^* and the achievable SE improve as N_{T_0} increases. In Fig. 8(b), N_{T_0} and N_{R_0} are fixed while N_{T_1} ranges from 2 to 6. Although N_{T_1} varies, MBS causes random interferences to PUE as the PUE adopts \mathbf{f}_0 to decode x_0 regardless of the interference channel \mathbf{H}_{10} . Hence, both ρ^* and the PUE’s average SE under different N_{T_1} remain similar. In Fig. 8(c), N_{T_0} and N_{T_1} are fixed while N_{R_0} ranges from 2 to 4. With such antenna settings, since N_{T_0} and N_{T_1} are fixed, the processing gain with the transmit antenna array doesn’t change for the desired signal or the interference. However, as N_{R_0} increases, the receive gain for the intended signal grows as the filter vector \mathbf{f}_0 , an $N_R \times 1$ vector designed to match \mathbf{H}_0 . As a result, the desired signal, relative to the interference, after the receive filtering becomes stronger with an increase of N_{R_0} , thus enhancing the PUE’s SE.

Fig. 9 shows the probabilities that the power overhead of DIS, OIS and IN are greater than P_0 , i.e., $Prob(P_{\mathcal{M}} > P_0)$ where $P_{\mathcal{M}}$ denotes the power cost at PBS with IM scheme \mathcal{M} . We use a general form $[N \xi \mathcal{M}]$ to denote the parameter settings for different mechanisms, where N is the number of interferences. $\xi = P_0^e/P_1^e$. P_1^e is the total power of the interferer, i.e., P_1 , incorporated with path loss. $Prob(P_{\mathcal{M}} > P_0)$ is shown to increase as ξ decreases since a small ξ results in a strong interference, incurring higher $P_{\mathcal{M}}$. $Prob(P_{IN} > P_0)$ is notably higher than $Prob(P_{OIS} > P_0)$ and $Prob(P_{DIS} > P_0)$, and $Prob(P_{DIS} > P_0)$ is less than $Prob(P_{OIS} > P_0)$. This is because IN consumes more power than DIS and OIS. For DIS, the power overhead increases

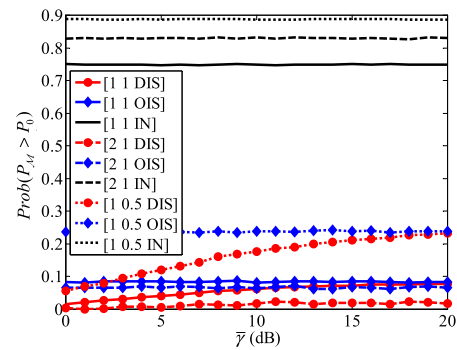


FIGURE 9. $Prob(P_{\mathcal{M}} > P_0)$ vs. $\bar{\gamma}$ under $M = 1$, and different ξ and N .

as $\bar{\gamma}$ grows and approaches OIS when $\bar{\gamma}$ becomes too large. One may note in Fig. 9 that with fixed ξ and P_1^e , $Prob(P_{IN} > P_0)$ with $N = 2$ interferences is higher than that with a single interference. However, as for DIS and OIS, a larger N produces lower $Prob(P_{\mathcal{M}} > P_0)$. This phenomenon can be explained by the results illustrated in Fig. 10.

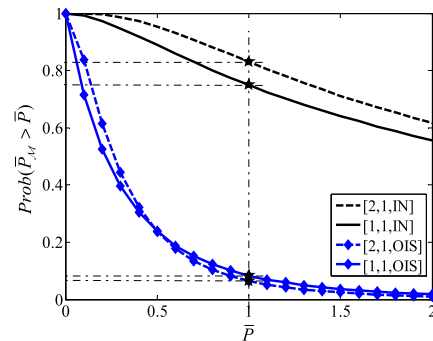


FIGURE 10. $Prob(P_{\mathcal{M}} > \bar{P})$ vs. \bar{P} under $\xi = 1$, $M = 1$, and different N .

Fig. 10 plots the distribution of $\bar{P}_{\mathcal{M}}$ for an arbitrary $\bar{\gamma}$ where $\bar{P}_{\mathcal{M}}$ and \bar{P} represents for $P_{\mathcal{M}}$ and the power value P normalized by P_0 . Since $Prob(\bar{P}_{DIS} > \bar{P})$ varies with $\bar{\gamma}$ whereas $Prob(\bar{P}_{IN} > \bar{P})$ and $Prob(\bar{P}_{OIS} > \bar{P})$ do not, for simplicity, we only study IN and OIS. As shown in the figure, $Prob(\bar{P}_{IN} > \bar{P})$ with $N = 2$ is no less than that with

$N = 1$. As for OIS, when $\bar{P} < 0.5$, $Prob(\bar{P}_{OIS} > \bar{P})$ with $N = 2$ is larger than that with $N = 1$. However, as \bar{P} grows larger than 0.5, 2 interferences incur statistically less power overhead. When $\bar{P} = 1$, $Prob(\bar{P}_{\mathcal{M}} > \bar{P})$ with different schemes shown in Fig. 10 is consistent with the results given in Fig. 9.

Fig. 11 shows the PUE's average SE with different IM schemes. Besides IN, OIS and DIS, zero-forcing (ZF) reception and zero-forcing beamforming (ZFBF), as well as DPC-based schemes are also simulated. As discussed in Section II, the system model we adopted is a cognitive ZIC [18] where PBS can accurately acquire the knowledge of the interference from MBS. Then, by intelligently precoding, e.g., dirty paper coding (DPC) [21], at PBS, interference at the PUE can be mitigated. Thus, an interference-free single-user point-to-point transmission from PBS to PUE can be realized which upper bounds PUE's SE. In [22], a practical application of DPC for the Gaussian cognitive ZIC was presented where the victim transmitter first obtains the interference signal from the interfering transmitter and then uses DPC to improve the performance of the victim link. For fairness comparison, we assume the victim transmitter can causally and accurately estimate the interference and operate in a full duplex manner [22], then the victim's SE can be computed as follows,

$$c_0^{DPC} = (1 - \varepsilon) \log_2 \left\{ 1 + \frac{[\lambda_0^{(1)}]^2 P_0^e}{\sigma_n^2} \right\} \quad (22)$$

where ε indicates the cost of estimating the interference. As can be seen from Eq. (22), the interference at the victim receiver can be completely eliminated by employing DPC. Moreover, one should note that when $\varepsilon = 0$, the SE upper bound is achieved. With ZF reception, a receive filter being orthogonal to the unintended signal is adopted so as to nullify the interference at PUE, but an attenuation w.r.t. the desired signal results. As for ZFBF, we let PBS adjust its beam so that the desired signal is orthogonal to the interference at the intended receiver. It should be noted that for either IN, IS with fixed ρ , OIS or DIS, when the power overhead at PBS exceeds P_0 , i.e., the IM scheme is unavailable, we simply switch

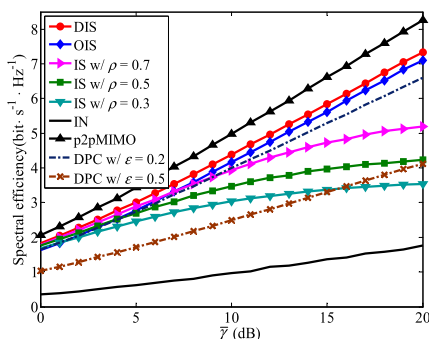


FIGURE 11. SE of PUE vs. $\bar{\gamma}$ with various IM schemes under $M = N = 1$, and $\xi = 1$.

to ZF reception. As shown in the figure, p2pMIMO, i.e., DPC with $\varepsilon = 0$, yields the best SE performance. DIS ranks the second best in SE. When $\bar{\gamma}$ is low, noise is the dominant factor affecting the PUE's SE. Therefore, IS with fixed ρ ($\rho < 1$) yields similar SE to OIS. Moreover, although DIS can achieve the second highest SE, its benefit is limited in the low $\bar{\gamma}$ region. As $\bar{\gamma}$ grows large, ρ^* increases accordingly, and hence IS with large ρ exceeds that with small ρ in SE. In addition, OIS gradually outperforms those IS schemes with fixed ρ as $\bar{\gamma}$ increases. Moreover, by intellectually determining ρ^* , the advantage of DIS becomes more pronounced with an increase of $\bar{\gamma}$. Although IN yields more power overhead than IS with fixed ρ , DIS and OIS, with the help of ZF, IN yields slightly higher SE than ZF reception. As for ZFBF, more desired signal power loss results as compared to ZF, thus yielding inferior SE performance. As for the DPC-based scheme proposed in [22], it can be clearly seen from Fig. 11 that due to the overhead of estimating interference, even when $\varepsilon = 0.2$, it's SE performance is still inferior to our method.

Fig. 12 plots the PUE's average SE with various mechanisms under different numbers of interferences and desired signals. We use a general form $[M \ N \ \mathcal{M}]$ to denote the parameter settings, where M represents the number of desired signals, N is the number of interferences, and \mathcal{M} denotes the IM schemes. When $M > 1$, equal power allocation is adopted, i.e., $P_{0,m}^e$ ($m = 1, \dots, M$) in Eq. (21) is P_0^e/M . As shown in the figure, OIS achieves the highest SE among the three schemes, whereas IN yields the lowest SE. Since ρ^* approaches 1 as $\bar{\gamma}$ increases, with the same M and N , DIS becomes OIS when $\bar{\gamma}$ grows too large. Given fixed ξ , IN yields better SE when there are 2 interfering signals than the single interference case. As for DIS and OIS, SE with 2 interferences is lower than that with one disturbance. This is consistent with the results shown in Figs. 9 and 10.

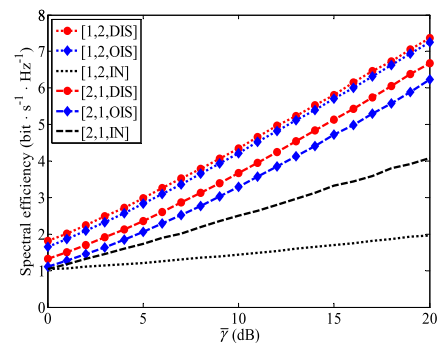


FIGURE 12. SE of PUE vs. $\bar{\gamma}$ with various IM schemes under $\xi = 1$ and different M and N .

VI. CONCLUSION

In this paper, we proposed a new interference management scheme, called *Dynamic Interference Steering* (DIS), for heterogeneous cellular networks. By intelligently determining the strength of the steering signal, the original interference is steered into an appropriate direction. DIS can balance the transmit power used for generating the steering signal and that

for the desired signal's transmission. Our in-depth simulation results show that the proposed scheme makes better use of the transmit power, and enhances users' spectral efficiency.

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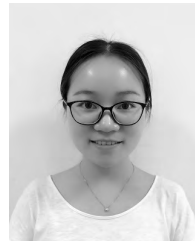
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Z. LI (S'08–M'10) was a Visiting Scholar and a Research Scientist with the University of Michigan from 2013 to 2015. He is currently an Associate Professor with Xidian University, Xi'an, Shaanxi, China. He is also a Research Scientist with the LEO Group, Xuzhou, Jiangsu, China. His research interests include wireless communication, resource allocation, interference management, and network security.



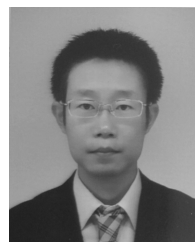
F. GUO is currently pursuing the master's degree with Xidian University. Her research interests include wireless communication and interference management.



C. SHU is currently pursuing the master's degree with Xidian University. Her research interests include wireless communication and interference management.



KANG G. SHIN (LF'12) is currently the Kevin and Nancy O'Connor Professor of computer science with the Department of Electrical Engineering and Computer Science, University of Michigan, Ann Arbor. His current research focuses on QoS sensitive computing and networking, and embedded real-time and cyber-physical systems.



J. LIU (S'10–M'12) is currently an Assistant Professor with the Cyber Security Research Center, National Institute of Informatics, Japan. His research interests include mobile ad hoc networks, 5G communication systems, D2D communications, cyber security, and physical layer security.

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