Aiming Perfectly in the Dark-Blind Interference Alignment Through Staggered Antenna Switching

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Abstract—We propose a blind interference alignment scheme for the vector broadcast channel where the transmitter is equipped with $M$ antennas and there are $K$ receivers, each equipped with a reconfigurable antenna capable of switching among $M$ preset modes. Without any knowledge of the channel coefficient values at the transmitters and with only mild assumptions on the channel coherence structure we show that $\frac{MK}{M+K-1}$ degrees of freedom are achievable. The key to the blind interference alignment scheme is the ability of the receivers to switch between reconfigurable antenna modes to create short term channel fluctuation patterns that are exploited by the transmitter. The achievable scheme does not require cooperation between transmit antennas and is therefore applicable to the $M \times K \times X$ network as well. Only finite symbol extensions are used, and no channel knowledge at the receivers is required to null the interference.

Index Terms—Blind interference alignment, broadcast channel, capacity, degrees of freedom, multiple-input single-output (MISO), multicast.

I. INTRODUCTION

BANDWIDTH is a precious resource for wireless networks. How to share this limited resource among multiple users is the primary challenge. A new signal multiplexing approach, called interference alignment, has shown recently that the bandwidth available to each user can be greatly improved [1], [2]. However, there remain very substantial hurdles in translating existing interference alignment schemes to practice. The most significant of these is the issue of channel knowledge. With few exceptions, the interference alignment schemes proposed so far require perfect, and often global, channel knowledge. In a network of distributed nodes, with time-varying channel coefficient values, this is very difficult if not altogether impossible. With this challenge as our motivation, in this work we pursue the goal of blind interference alignment, i.e., we seek to align interference with no knowledge of channel conditions at either transmitters or receivers.

We consider a multiple-input single-output (MISO) broadcast channel with multiple multicasts, or the compound MISO BC, which refers to the setting where the base station transmitter is equipped with $M$ transmit antennas and each receiver is equipped with a single receive antenna. There are $K$ independent messages intended for $K$ groups of receivers, each group consisting of $J$ distinct receivers who want the same message. The study of the degrees of freedom for the compound MISO BC was pioneered by Weingarten, Shamai and Kramer in [3]. In this work, it was shown that the multiplexing gain in this setting cannot be more than $\frac{MK}{M+K-1}$ under the assumption that the channel states of all users are globally and perfectly known to the transmitters and receivers. It was also conjectured that as the number of users $J$ in each group increases, the multiplexing gain will collapse to unity. Recent work in [4] and [5] disproved this conjecture and showed that interference alignment is still feasible in the compound vector broadcast channel so that a total multiplexing gain of $\frac{MK}{M+K-1}$ is achievable regardless of the number users $J$ in each group, provided $J$ is finite. The surprising result is based on an interference alignment scheme that exploits the separation properties of lattices scaled by rationally independent scalars.

With no channel state information at the transmitter, interference alignment may appear impossible. Yet, it has been shown in [6] that in certain scenarios blind interference alignment is not only possible, but also it can be quite simple. For instance, in the context of the MISO BC with $M = 2$ antennas at the base station and $K = 2$ users, even when the transmitter has no CSIT and the channel coefficients may be drawn from a continuum, [6] showed that if the coherence blocks of the two users are suitably staggered then the outer bound value of $\frac{4}{3}$ DoF is achievable. Moreover, the achievable scheme is a form of repetition coding over a supersymbol consisting of three channel uses.

The supersymbol structure used for blind interference alignment in [6] arises naturally in certain practical settings. However, in general, e.g., if the channel coherence does not display any special properties, the problem of blind interference alignment remains open. In this work we address this setting and consider a bolder approach, summarized in the following question: Can we artificially manipulate the channel itself to create the opportunities that facilitate blind interference alignment?

The goal of manipulating the channel naturally leads us to reconfigurable antennas. A reconfigurable antenna is an antenna that can change its characteristics by dynamically changing its geometry. Specifically, the current distribution over the volume of the antenna is changed by switching on and off various geometrical metallic segments (pixels, strips etc.) that constitute the reconfigurable antenna. Each distinct geometrical configuration

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corresponds to a different mode of operation. Various technologies, such as microelectromechanical switches (MEMS), nanoelectromechanical switches (NEMS), or solid state switches are used to perform the switching operation and offer various trade-offs in terms of the switching speed, insertion loss, monolithic integration capability, size and reliability [7]–[9]. A reconfigurable antenna offers a choice to switch among several pre-set modes. It is more flexible than a single conventional antenna in its ability to switch its radiation pattern among a fixed number of preset modes, and yet it is less flexible than multiple conventional antennas which can be used together with arbitrary beamforming weights to construct logical beams over a continuum of possibilities. For many applications, a single reconfigurable antenna is more desirable than multiple conventional antennas because 1) it needs only one RF chain, and 2) the integrated design of a reconfigurable antenna makes it smaller than multiple conventional antennas. However, the popularity of reconfigurable antennas has been limited by the assumption of fixed preset modes which do not allow continuous adaptability needed for beamforming techniques such as zero forcing, that are typically used to multiplex signals. As a consequence, reconfigurable antennas have heretofore been explored for diversity benefits, but not for multiplexing gain. The focus of research in reconfigurable antenna fabrication has been to create desired radiation patterns. However, in this work, we use antenna switching not to direct the beam in a specific direction, but rather to introduce channel fluctuations at pre-determined time instants. More important, we will use blind antenna switching, i.e., the antenna switching will not be based on the channel state information at the receivers (CSIR). This relaxed requirement could also simplify the design of the reconfigurable antenna. For our purpose in this work, we ignore the specific hardware aspects and focus instead on the conceptual model of a reconfigurable antenna. At a conceptual level, we model a reconfigurable antenna as capable of switching among independent dumb (isotropic) modes, shown in Fig. 2. Note that the conceptual model is identical to antenna selection.

The traditional approach for antenna switching is for each receiver to choose the mode that allows the greatest signal strength for that particular receiver [10], [11]. We call this “selfish” antenna switching. If all the channels are statistically equivalent and the transmitter has no instantaneous CSIT, it is easily seen that the conventional (selfish) antenna selection approach does not allow any DoF advantage.

The following insight is distilled from [6] and forms the basis for this work:

“Instead of each receiver (selfishly) selecting the best reconfigurable antenna mode, if the receivers blindly switch antennas according to a pre-determined signature pattern, then they can artificially create the staggered coherence block structures needed for blind interference alignment.”

Consider the $M - K = 2$ multicast setting. Suppose each receiver blindly switches from one preset mode to another every two symbols. However the switching instants are staggered. While all receivers in group 1 switch on odd time slots, the receivers in group 2 switch on even time slots. This creates the supersymbol structure shown in Fig. 3 which is the key to the blind interference alignment scheme in [6]. Each supersymbol consists of three symbols. The channel state of every user in one group (labeled as group 1 in the figure) changes after the first symbol and remains fixed for the last two symbols, while the channel state of every user in the other group (group 2 in the figure) is fixed for the first two symbols and changes in the last symbol. Once this supersymbol structure is achieved, the staggered coherence block coding (SCBC) scheme proposed in [6] is directly applied to achieve $\frac{3}{2}$ DoF, regardless of the number of receivers in each group. Further, note that this is accomplished without any special assumptions on the channel coherence block structures, except that the coherence time for each user is long enough to span the coding block.

The $M = 2$, $K = 2$ MISO BC example suggests that staggered antenna switching (SAS) with reconfigurable antennas can be used to translate the staggered coherence block coding schemes of [6] into practice. However, there are important distinctions in terms of what kinds of supersymbol structures can result in the setting of [6] versus our setting in this paper. For
example, consider the $K$ user interference channel for which an SCBC scheme is proposed in [6] that is capable of achieving $\frac{K}{2}$ DoF. This SCBC scheme is contingent on a very specific temporal correlation model—the cross channels follow the same temporal correlation structure which is unlike the direct channels. However, note that antenna switching at a receiver cannot selectively change the channels for some transmitters while keeping the channels from other transmitters to that same receiver unchanged. When we switch a receive antenna, all the channel coefficients associated with that receive antenna will change. Thus, some supersymbol structures possible with the staggered coherent block model of [6] are not possible with staggered antenna switching of reconfigurable antennas.

On the other hand, the converse is also true. Staggered antenna switching, because it can follow any complex pattern, allows some supersymbol structures that cannot result simply from staggered coherent blocks. Thus, the staggered antenna switching (SAS) framework that we explore in this paper has a distinct character from the staggered coherence block coding (SCBC) framework considered in [6] and in general neither can be seen as a special case of the other.

We summarize the salient features of the blind interference alignment scheme proposed in this work as follows.

1. No CSIT is required to align interference. Unlike [3] and [4], the channel uncertainty may be spread over a continuum.

2. No cooperation among transmit antennas is needed. Thus the BC results extend to the X channel setting. This is similar to the result in [4] and [6].

3. The multiplexing gain achieved with no CSIT for the MISO BC with multiple $J$ multicasts is the same as the maximum possible multiplexing gain with full CSIT for large $J$.

4. The multiplexing gain achieved with no CSIT for the X channel, is the same as the maximum possible multiplexing gain with full CSIT. This is similar to the result in [4].

5. Alignment is achieved by coding over only a finite number of symbols. This is in contrast to [12] where infinite symbol extensions are needed to achieve the outer bound for X channel even with perfect CSIT.

6. No CSIR is needed for the antenna switching pattern used by each receiver.

7. No CSIR is required to null the interference (without losing the desired signal dimensions). While we do not consider non-coherent communication in this paper, the proposed scheme can be directly extended to non-coherent settings with e.g., differential coding schemes.

8. Unlike [6], no special assumptions are needed on channel coherence structure. In fact, except for the antenna switching patterns, the receivers may even be statistically equivalent, i.e., indistinguishable to the transmitter.

II. SYSTEM MODEL

Consider, as before, the MISO BC with multiple multicasts, where the transmitter has $M$ antennas while each receiver is equipped with one reconfigurable antenna (and thus only one RF chain) that can switch among $M$ preset modes. Since there is no CSIT, the multiple multicast setting is subsumed under the conventional MISO BC setting where we have a unique receiver for each independent message. We will focus on this latter scenario. The results can be extended easily to the multiple multicast setting by having every receiver in a particular group follow the same strategy as the unique receiver that stands as proxy for this group in the classical MISO BC.

Let $K$ be the number of receivers. Further, let us denote the $1 \times M$ channel vector associated with the $m$th preset mode of user $k$’s reconfigurable receive antenna as $h_t^{k}(m) \in \mathbb{C}^{1 \times M}$ where $k \in \mathcal{K} = \{1, 2, \ldots, K\}$ and $m \in \mathcal{M} = \{1, 2, \ldots, M\}$. We assume that the channel vectors are generic, by which we mean that they are drawn from a continuous distribution (bounded away from zero and infinity to avoid degenerate cases), so that any $M$ of them are linearly independent almost surely.

We make no special assumption on the channel coherence block structures, except that the coherence times are long enough so that the channels stay constant across a supersymbol. The supersymbols will be defined later for each $M, K$.

With the staggered antenna switching scheme, the receivers switch between their antenna modes in a predetermined pattern, so that at time $t$, the mode selected by receiver $k$ is $m_k^{(t)}$, and the resulting channel for user $k$ is represented as $h^{k}(m_k^{(t)}(t))$. Thus, the received signal for the $k$th user, at time $t$, is

$$y^{k}(t) = h^{k}(m_k^{(t)}(t))x(t) + z^{k}(t), \quad k \in \mathcal{K}, m_k^{(t)} \in \mathcal{M}$$

where $x(t) \sim \mathcal{CN}(0, \mathbf{I})$ is the additive white Gaussian noise (AWGN). The channel input is subject to an average power constraint $\mathbb{E}[\|x\|^2] \leq P$. Unless explicitly stated otherwise, we assume no channel state information at the transmitter (CSIT), i.e., the channel coefficient values are not known to the transmitters. While the blind interference alignment scheme does not require channel state information at the receivers either, i.e., we need no CSIR to either align interference or to null it out at the receiver, we will keep the assumption of perfect CSIR primarily in order to be able to define the degrees of freedom (or multiplexing gain) as the capacity pre-log, and thereby provide us a clean metric for gauging the extent of interference alignment. We will point out the no-CSIR requirement feature of our proposed scheme later on in this paper. While the channel coefficient values are not known, we assume that the switching pattern functions $m_k^{(t)}$, because they are pre-determined by design (like the codebooks), are known to everyone.

The transmitter sends an independent message $W^{k}$ with rate $R^{k}$ to receiver $k, \forall k \in \mathcal{K}$. A rate tuple $\mathbf{R} = (R^{[1]}, R^{[2]}, \ldots, R^{[K]})$ is achievable if every receiver is able to decode its message with an error probability that can be made arbitrarily small by coding over sufficient channel uses. The closure of the set of all achievable rate tuples is the capacity region $\mathcal{C}$. The degrees of freedom metric, $d$, is defined as

$$d = \lim_{P \to \infty} \max_{\mathbf{R} \in \mathcal{R}} \frac{R^{[1]} + \cdots + R^{[K]}}{\log P}$$
Fig. 4. Alignment block and the corresponding beamforming matrix of user 1 for K-user M x 1 MISO BC.

III. BLIND INTERFERENCE ALIGNMENT FOR THE K-USER M x 1 MISO BC

We present the main result in the following theorem.

Theorem 1: For the K-user M x 1 MISO BC defined in Section II, a total of \( \frac{M K}{M+K-1} \) DoF are achievable, almost surely.

As mentioned before, the achievable scheme relies on designing an antenna switching pattern for each user and designing a beamforming strategy based on the corresponding temporal correlation structure. The goal is to achieve interference alignment, which refers to the construction of signals in such a manner that they cast overlapping shadows at the receivers where they constitute interference while remain distinct where they are desired. In the following subsection, we first highlight the key to the interference alignment schemes used in this work.

A. The Key to Blind Interference Alignment—The Alignment Block

Consider the K-user M x 1 MISO broadcast channel defined in Section II. For simplicity, consider only the transmission of the message for user 1. Suppose multiple symbols are transmitted for user 1. These transmitted symbols for user 1 will cause interference at all other users. Interference alignment means that we would like to keep these symbols distinct at receiver 1, but consolidate them into a smaller subspace at all other receivers. Most importantly, we must accomplish this without any knowledge of the channel coefficients.

1) Staggered Antenna Switching Pattern: Consider M time slots. During these M time slots, receiver 1 switches his reconfigurable antenna mode each time to go through all M modes. All other receivers do not switch their antenna modes, i.e., they listen to all transmissions through the same channel. The resulting supersymbol structure is shown in Fig. 4.

2) Beamforming: During time slot 1, suppose the transmitter sends independent symbols for user 1, one from each transmit antenna. Now, let the transmitter repeat the same transmission over a total of M times. The transmitted vector can be represented as

\[
X = \begin{bmatrix}
x(1) \\
x(2) \\
\vdots \\
x(M)
\end{bmatrix} = \begin{bmatrix}
I \\
I \\
\vdots \\
I
\end{bmatrix} \begin{bmatrix}
\mathbf{h}^{[1]}(1) \\
\mathbf{h}^{[1]}(2) \\
\vdots \\
\mathbf{h}^{[1]}(M)
\end{bmatrix}
\]  

(2)

where I is the M x M identity matrix. Note that the beamforming vectors do not depend on the channel values.

With this scheme (ignoring AWGN), the signal received at user 1 is

\[
Y = \begin{bmatrix}
y^{[1]}(1) \\
y^{[1]}(2) \\
\vdots \\
y^{[1]}(M)
\end{bmatrix}
\]

\[
= \begin{bmatrix}
0 & 0 & 0 \\
0 & \mathbf{h}^{[1]}(2) & 0 \\
\vdots & \vdots & \vdots \\
0 & 0 & \mathbf{h}^{[1]}(M)
\end{bmatrix} \begin{bmatrix}
\mathbf{u}^{[1]}_1 \\
\mathbf{u}^{[1]}_2 \\
\vdots \\
\mathbf{u}^{[1]}_M
\end{bmatrix}
\]  

\[
= \begin{bmatrix}
\mathbf{h}^{[1]}(1) \\
\mathbf{h}^{[1]}(2) \\
\vdots \\
\mathbf{h}^{[1]}(M)
\end{bmatrix} \begin{bmatrix}
\mathbf{u}^{[1]}_1 \\
\mathbf{u}^{[1]}_2 \\
\vdots \\
\mathbf{u}^{[1]}_M
\end{bmatrix}
\]

(3)

Since all rows of the effective channel matrix in (3) are the same, its rank is equal to 1. Thus, the M symbols intended for user 1 cast only a 1-dimensional shadow on all undesired receivers, i.e., they align into one dimension. Moreover, notice that they align along the M x 1 vector \([1 1 \ldots 1]^T\), regardless of the channel values. Therefore, even if the receiver does not know the channel coefficients, it can project the received signal into the null space of the all-ones vector to zero force the interference. This is the key to the blind interference cancellation at the receiver mentioned earlier.

We summarize the key to blind interference alignment as follows: If the channel of the desired user changes while that of all undesired users remains fixed over M symbols, then the transmitter can send M data streams for the desired receiver without...
Note that with such an alignment scheme, $M$ interference vectors can be aligned into one dimension. Thus, intuitively the achievable $\frac{MK}{M+K-1}$ DoF for $K$-user $M \times 1$ MISO BC can be interpreted as follows—Every receiver demands $M$ DoF for a total of $MK$ DoF; at each receiver, the desired signals occupy $M$ dimensions while $(M-K)$ interference streams are aligned into $K-1$ dimensions, for a total of $M+K-1$ dimensions.

The supersymbol structure shown in Fig. 4 will serve as the building block for designing the supersymbol, i.e., antenna switching patterns, for the general $K$-user $M \times 1$ MISO BC. Thus, we refer to it as the alignment block in the following part of this paper. Our goal is to construct the alignment block for each user. Furthermore, with the alignment block, the design of beamforming vectors becomes straightforward. As shown in Fig. 4, over the $M$ symbols of the alignment block, the beamforming matrix is obtained by stacking the $M \times M$ identity matrix $M$ times. Finally, the $M$ symbols constituting the alignment block may not be necessarily consecutive. In this case, it can be obtained through an interleaving of symbols.

With this insight, we begin with the $K$-user $2 \times 1$ MISO BC, go on to $K$-user $3 \times 1$ case, and at last solve the general $K$-user $M \times 1$ case.

### B. $K$-User $2 \times 1$ MISO Broadcast Channel

We begin with the two-user case. Although the solution for this case, as mentioned in the introduction, follows directly from [6], here we use this simplest setting as an example to illustrate how to use the alignment block mentioned in the last section to construct the supersymbol structure.

For the two-user $2 \times 1$ MISO BC, our goal is to achieve $\frac{2}{3}$ DoF. This can be done by sending two data streams, each carrying one DoF, to each user over three symbol extensions. We design the supersymbol as shown in Fig. 5. The first two symbols constitute an alignment block for user 1. For user 2, we design the third symbol such that through an interleaving of the first and third symbols, they are converted into an alignment block.

Based on the supersymbol, we can design the $6 \times 2$ beamforming matrix for each user as follows. The beamforming matrix is constructed by stacking three $2 \times 2$ matrices, which is equal to the number of symbols in the supersymbol. The $j^{\text{th}}$, $j=1,2,3$, block in the beamforming matrix for each user corresponds to the $j^{\text{th}}$ symbol in the supersymbol. For each user, if the symbol belongs to his alignment block, then the corresponding block in his beamforming matrix is a $2 \times 2$ identity matrix. Otherwise, it is a $2 \times 2$ zero matrix. This is illustrated in Fig. 5.

The transmitted signal is

$$
\mathbf{X} = \begin{bmatrix}
1 & 1 \\
0 & 1 \\
0 & 0
\end{bmatrix} \begin{bmatrix}
\mathbf{u}_{[1]}[1] \\
\mathbf{u}_{[2]}[1]
\end{bmatrix} + \begin{bmatrix}
1 & 0 \\
0 & 1
\end{bmatrix} \begin{bmatrix}
\mathbf{u}_{[1]}[2] \\
\mathbf{u}_{[2]}[2]
\end{bmatrix}
$$

where $\mathbf{I}$ is a $2 \times 2$ identity matrix. $\mathbf{u}_{[i]}[j], i=1,2,$ are two independent encoded data streams intended to user $i$, each carrying one DoF. With this scheme, the received signal at user 1 is

$$
\begin{bmatrix}
\mathbf{y}_{[1]}[1] \\
\mathbf{y}_{[2]}[1] \\
\mathbf{y}_{[3]}[1]
\end{bmatrix} = \begin{bmatrix}
\mathbf{h}_{[1]}[1] \\
\mathbf{h}_{[2]}[1] \\
0
\end{bmatrix} \begin{bmatrix}
\mathbf{u}_{[1]}[1] \\
\mathbf{u}_{[2]}[1]
\end{bmatrix} + \begin{bmatrix}
\mathbf{h}_{[1]}[2] \\
\mathbf{h}_{[2]}[2] \\
0
\end{bmatrix} \begin{bmatrix}
\mathbf{u}_{[1]}[2] \\
\mathbf{u}_{[2]}[2]
\end{bmatrix}
$$

where $\mathbf{0}$ is a $1 \times 2$ zero vector. From (5), it can be easily seen that interference is aligned into one dimension along vector $[1 0 1]^T$ while the desired signals appear through a full rank matrix, and are therefore resolvable. In addition, notice that the third row of the desired signals is zero, while that of the interference vector, $[1 0 1]^T$, is 1. Therefore, any linear combination of the desired signals leads to zero in the third row, ensuring linear independence among them. Similarly, user 2 can achieve 2 DoF as well. Thus, 4 DoF can be achieved over three symbol extensions, so that $\frac{4}{3}$ (normalized) DoF are achieved.

**Remark 1:** From (5) note that user 1 can cancel the interference due to user 2 by simply subtracting the third received symbol from the first. This operation does not require any knowledge of channel coefficient values at the receiver and produces an interference-free signal while leaving the desired signal unaffected, although the noise is doubled over the first symbol. The blind interference cancellation property is common to all the blind interference alignment schemes proposed in this paper.

With the understanding of the two-user case, let us consider the general $K$ user $2 \times 1$ MISO BC. The supersymbol structure is shown in Fig. 6. The supersymbol consists of $K+1$ symbols. Each user’s channel state only comes from one of two channel values corresponding to channels associated with two reconfigurable modes of each receiver’s reconfigurable antenna. For user $k$, $k \in K$, the channel state over the first $k$ symbols maintains the first value, it changes to the other value at the $(k+1)^{\text{th}}$ symbol and changes back to the first value at the $(k+2)^{\text{th}}$ symbol and remains fixed until the end of the supersymbol. With this structure, user $k$ can obtain its alignment block through an interleaving of symbols at the first time slot and the $(k+1)^{\text{th}}$ time slot. Over each alignment block, two data streams, each carrying one DoF, are transmitted to its corresponding user while remaining aligned into one dimension at all
other users. According to the supersymbol structure, the beamforming matrix can be obtained as follows:

\[
\begin{bmatrix}
1 & 1 & \ldots & 1 \\
0 & 0 & \ldots & 0 \\
\vdots & \vdots & \ddots & \vdots \\
0 & 0 & \ldots & 1
\end{bmatrix}_{2K+1 \times 2K}^{2(K+1) \times 2K}
\]

The \(k^{th}\) block column carries two data streams for user \(k\). With this scheme, at each user, the desired signals occupy two dimensions in the \(K + 1\)-dimensional signal space while all interference occupies \(K - 1\) dimensions, one from each interferer. As a result, each user can achieve 2 DoF, for a total of \(2K\) DoF, over \(K\) symbol extensions. Thus, \(\frac{2K}{K+1}\) DoF are achieved.

\(K\)-User 3 \(\times\) 1 MISO BC

When the number of antennas at the transmitter increases to 3, the alignment problem is more challenging. However, the key idea, as before, is to construct alignment blocks for each user. We begin with the two-user 3 \(\times\) 1 case.

1) Two-User 3 \(\times\) 1 MISO BC: We need to show \(2 \times 3\) DoF are achievable. This can be achieved with eight symbol extensions, over which each user achieves 6 DoF, for a total of 12 DoF. Each user’s six beams can be sent over two alignment blocks, three for each alignment block, so that they can be aligned into two dimensions at the other receiver, leaving the remaining six dimensions for the desired signals. The supersymbol consisting of eight symbols is shown in Fig. 7. As we can see, the first six symbols constitute two alignment blocks for user 1. On the other hand, two alignment blocks for user 2 can be obtained by adding two additional symbols and through an interleaving of symbols as shown in Fig. 7. It is important to note that for each user, two alignment blocks do not overlap. Thus, signals sent over two alignment blocks are orthogonal in time. As a result, three beams sent over one alignment block are orthogonal to the other three beams sent over the other alignment block, ensuring six beams can be separated at the desired user.

With the supersymbol structure, we can design the beamforming matrices. Similar to the two-antenna case, the beamforming matrix for each user can be obtained according to the alignment block. Unlike the two-antenna case, where each user has only one alignment block, here each user has two alignment blocks. In this case, each user’s beamforming matrix consists of two block columns, each corresponding to one alignment block. Each block column can be designed following the same mapping used in the two-antenna case as illustrated in Fig. 7.

With this scheme, it can be easily seen that the interference is aligned into two dimensions. For example, at user 1, the interference from user 2 is spanned by two columns \([1 \ 0 \ 0 \ 1 \ 0 \ 0 \ 0 \ 1]^T\) and \([0 \ 1 \ 0 \ 0 \ 1 \ 0 \ 0 \ 1]^T\). Moreover, the six dimensions occupied by the desired signal and two interference dimensions do not overlap. This is due to the orthogonality among signals in the seventh and eighth dimensions. Thus, user 1 is able to achieve 6 DoF. Similarly, user 2 can achieve 6 DoF as well.

So far, we have seen two problems that we need to consider and the ideas used to solve them. One is the alignment problem and the other is the linear independence issues. The alignment problem can be solved by constructing non-overlapping alignment blocks for each user. Linear independence issues include linear independence of the desired signals at the desired receiver, and their independence with the interference. For the desired signals, we already see that the signals transmitted over one alignment block are linearly independent. The linear independence of data streams transmitted across different alignment blocks is guaranteed by orthogonality among alignment blocks in time. For the linear independence between desired signals and the interference, it is ensured by orthogonality of signals in the last symbol of each alignment block. Notice that in the last symbol of each alignment block, only the signal sent over this alignment block is active. Thus, after being aligned into one dimension, the interference vector is linearly independent with all other signals.

To make things systematic, we can change the order of symbols in the supersymbol and their corresponding rows in the beamforming matrix. The goal is to separate the alignment problem and linear independence issues. In particular, we group the last symbols of all alignment blocks into a block which we refer to as Block 2 as illustrated in Fig. 8. The remaining part is called Block 1. Basically, Block 1 ensures alignment, while Block 2 guarantees desired signals do not overlap with interference.

We can argue that once Block 1 is designed, Block 2 can be determined automatically. Note every symbol in Block 2 can be
grouped with two symbols in Block 1 as an alignment block. In addition, over an alignment block, the channel state of the desired user changes while that of all undesired users remains fixed. Thus, for the desired user, the last symbol of the alignment block is set to be the third channel value. For all other undesired users’ channels, they are set to be the same as the channel in previous symbols of that alignment block. This is illustrated in Fig. 8 where three symbols constituting one alignment block for user 1 are circled. Notice that once the two symbols are determined in Block 1, the third one in Block 2 is determined uniquely. As a result, we now only need to design the channel structure in Block 1. Once designed, Block 2 can be determined automatically.

2) Structure of Block 1 and Design of Beamforming Vectors: In this section, we will show how to design Block 1 of the supersymbol for $K$-user $3 \times 1$ MISO BC and how to design the beamforming vectors at the transmitter based on Block 1. To understand the structure of Block 1, we first consider the two-user $3 \times 1$ MISO BC. Block 1 is shown in Fig. 9. First note that in Block 1, since there is no third symbol of the alignment block, only two different channel values for each user are needed. In addition, symbols in Block 1 are periodic. We refer to the block in one period as the building block. As shown in Fig. 9, the building block of user 1 consists of $2^1$ symbols while that of user 2 consists of $2^2$ symbols. Since Block 1 consists of four symbols, user 1 has two building blocks while user 2 has only one building block. To design the beamforming vectors, we can group the time slots with two different channel values within every building block of each user as shown in Fig. 9 to constitute the first two symbols of the alignment block.

In general, if there are $K$ users, then Block 1 consists of $2^K$ symbols. Each user’s channel states are periodic in Block 1 with the building block shown in Fig. 10. Note that for simplicity, we use the same color to denote the channel for different users. However, this does not mean that channel values at different users are the same. In fact, they are different with probability one. As we can see, the building block of user $n$ is comprised of two subblocks, each with length $2^{n-1}$, for a total of $2^n$ symbols. The channel state remains fixed within each subblock while it changes across different subblocks. The channel state in the $i^{th}$, $i = 1, 2$, subblock corresponds to the channel vector associated with the $i^{th}$ preset antenna mode at the user. To obtain the temporal correlation signature for user $n$ in Block 1, we can simply repeat its building block $2^{K-n}$ times. In other words, there are $2^{K-n}$ building blocks for user $n$ in Block 1.

To design the beamforming vectors for user $n$, we can group the $i^{th}$, $i = 1, \ldots, 2^{n-1}$ symbol in one subblock with the one in the other subblock within one building block as shown in Fig. 10. Since there are $2^{K-n}$ building blocks, a total of $2^{K-1}$ groups can be created. It can be verified these groups satisfy the requirement of the alignment block. Detailed explanations are deferred to next section where the general $K$-user $M \times 1$ MISO BC is considered.

With this construction, we can calculate the achievable DoF. First let us consider the number of symbols in Block 2. In Block 1, each user has $2^{K-1}$ groups. Since each group needs one symbol in Block 2 to constitute one alignment block, a total of $K \times 2^{K-1}$ symbols are needed in Block 2. Therefore, the total number of symbols in the supersymbol is $2^K + K \times 2^{K-1}$. On the other hand, three beams can be transmitted over one alignment block for the desired user while aligned into one dimension at all other users. Since each user has $2^{K-1}$ alignment block, at the $2^K + K \times 2^{K-1}$ dimensional receiver’s signal space, $3 \times 2^{K-1}$ dimensions are occupied by the desired signals while the remaining $(K-1) \times 2^{K-1}$ dimensions are occupied by the interference. Since signals are orthogonal in Block 2, desired signals do not overlap with the interference. Therefore, the normalized DoF are $K \times 3 \times 2^{K-1} = K$.

Next, we take the three-user $3 \times 1$ MISO BC as an example to illustrate the points mentioned in this section.

3) Example: Three-User $3 \times 1$ MISO BC: Let us first consider Block 1. For three users, Block 1 consists of $2^3$ symbols. The building blocks for user 1, 2, and 3 are shown in Fig. 10. Therefore, there are 4, 2, and 1 building blocks for user 1, 2, and 3, respectively. Block 1 is illustrated in Fig. 11, in which the grouping for each user is also shown. It can be seen that for each user, the channel in two time slots of each group changes at the desired user while it remains the same at the undesired users, as required by the alignment block. After adding Block 2, the complete supersymbol structure for the three-user $3 \times 1$ case is
shown in Fig. 12. The beamforming matrices for user 1, 2, and 3 can be easily obtained through the supersymbol structure as

\[
\begin{bmatrix}
1 & 0 & 0 & 0 \\
1 & 0 & 0 & 0 \\
0 & 1 & 0 & 0 \\
0 & 1 & 0 & 0 \\
0 & 0 & 1 & 0 \\
0 & 0 & 1 & 0 \\
0 & 0 & 0 & 1 \\
0 & 0 & 0 & 1 \\
0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0
\end{bmatrix}
\]

Fig. 12. The supersymbol structure for three-user $3 \times 1$ MISO BC.

\[
\begin{bmatrix}
1 & 0 & 0 & 0 \\
1 & 0 & 0 & 0 \\
0 & 1 & 0 & 0 \\
0 & 1 & 0 & 0 \\
0 & 0 & 1 & 0 \\
0 & 0 & 1 & 0 \\
0 & 0 & 0 & 1 \\
0 & 0 & 0 & 1 \\
0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0
\end{bmatrix}
\]

\[
\begin{bmatrix}
1 & 0 & 0 & 0 \\
1 & 0 & 0 & 0 \\
0 & 1 & 0 & 0 \\
0 & 1 & 0 & 0 \\
0 & 0 & 1 & 0 \\
0 & 0 & 1 & 0 \\
0 & 0 & 0 & 1 \\
0 & 0 & 0 & 1 \\
0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0
\end{bmatrix}
\]

The user 1, 2, and 3 beamforming matrices can be obtained through the supersymbol structure.

\[D. \textit{K-User } M \times 1 \textit{ MISO Broadcast Channel}\]

In this section, we consider the general $K$-user $M \times 1$ case and show how to systematically design the supersymbol structure and the beamforming matrix. This consists of three steps.

Step 1: Design Block 1: Block 1 consists of a total of $(M - 1)^K$ symbols. Each user’s channel state switching pattern is periodic in Block 1 with the building block shown in Fig. 13. As we can see, the building block of user $n$ is comprised of $M - 1$ subblocks, each with length $(M - 1)^n - 1$, for a total of $(M - 1)^n$ symbols. The channel state remains fixed within each subblock while it changes across different subblocks. The channel state in the $m^{th}$, $m = 1, \ldots, M - 1$, subblock corresponds to the channel vector associated with the $m^{th}$ preset antenna mode at the user. To obtain the temporal correlation signature for user $n$ in Block 1, we can simply repeat its building block $(M - 1)^{K - n}$ times. In other words, there are $(M - 1)^K - n$ building blocks for user $n$ in Block 1.

In general, each receiver’s temporal correlation can be described by a function of time whose values come from all the possible channel values it can take. Let us define the function of time $t \in \mathbb{N}$ for user $n$ as $f_n(t)$ whose value is drawn from the set \{h^{n}(1), h^{n}(2), \ldots, h^{n}(M)\}. Then we can denote the channel of Block 1 for user $n$ as shown in (7), at the bottom of the page, where $t = 1, 2, \ldots, (M - 1)^K$, since the length of Block 1 is $(M - 1)^K$.

Step 2: Design Beamforming Matrices: With Block 1 designed, we can now design the beamforming vectors. As mentioned before, the key is to create non-overlapping alignment blocks for each user. Each alignment block corresponds to one block column in the beamforming matrix, which is obtained by placing the $M \times M$ identity matrix at the rows corresponding to the symbol instants of the alignment block and zero matrix elsewhere.

Note that in Block 1, there are $M - 1$ distinct symbols for each user. Therefore, we will construct the first $M - 1$ symbols, which are referred to as a group, of the alignment block. The last symbol for each alignment block is provided in Block 2. Now consider user $n$. Recall that the channel state of the desired user changes over an alignment block. Therefore, we can group the $i^{th}$ symbol in each of $M - 1$ subblocks within one building block as shown in Fig. 13. Since there are $(M - 1)^n - 1$ symbols in one subblock, a total of $(M - 1)^n$ groups can be created from one building block. Mathematically, within each building block, the $i^{th}$, $i = 1, 2, \ldots, (M - 1)^n - 1$, group consists of following symbols:

\[i, i + (M - 1)^{n-1}, \ldots, i + (M - 2)(M - 1)^{n-1}. \quad (8)\]

Such grouping is repeated within each building block for a total of $(M - 1)^K - n$ building blocks for user $n$. Mathematically, we

\[
f_n(t) = \begin{cases} 
\textbf{h}^{n}(1) & t = 1, 2, \ldots, (M - 1)^{n-1}(\text{mod}(M - 1)^n) \\
\vdots & \\
\textbf{h}^{n}(j) & t = (j - 1)(M - 1)^{n-1} + 1, \ldots, j(M - 1)^{n-1}(\text{mod}(M - 1)^n) \\
\vdots & \\
\textbf{h}^{n}(M - 1) & t = (M - 2)(M - 1)^{n-1} + 1, \ldots, (M - 1)^n - 1, 0(\text{mod}(M - 1)^n). 
\end{cases}
\]
can represent the $i^{th}$ group in the $k^{th}$, $k = 1, \ldots, (M-1)^{K-n}$, building block as

$$(k-1)(M-1)^n + i, (k-1)(M-1)^n + i + (M-1)^{n-1}, \ldots, (k-1)(M-1)^n + i + (M-2)(M-1)^{n-1}.\tag{9}$$

Note that $(M-1)^n$ is the length of one building block.

We need to check that each group satisfies the requirement of the alignment block, i.e., the channel state of the desired user changes while that of all other users remains constant in each group. Note that (9) specifies the time instants of symbols for each group, and (7) describes how channel changes with time for every user. Therefore, we can use (7) to verify how channel state changes at each user during the time instants of each group given in (9). Consider user $n$. Let us first verify that the channel state changes at the desired user, i.e., user $n$. Note that the $i^{th}$ group in the $k^{th}$ building block is given in (9). Now we calculate the channel values in these time slots by (7). It can be easily seen that the remainders of these time slots divided by $(M-1)^n$ are as follows:

$$i \mod (M-1)^n, \ldots, i + (M-2)(M-1)^{n-1} \mod (M-1)^n.$$ 

Therefore, according to (7), the channel values at these time slots are $h^{[n]}(1), h^{[n]}(2), \ldots, h^{[n]}(M-1)$. In other words, the channel state changes as required.

Now let us verify that within each group of user $n$ channels of all other users remain unchanged. First consider channels of user $j = 1, 2, \ldots, n-1$. Notice that the remainders of each group’s time slots divided by $(M-1)^{n}$ are the same, i.e., $i \mod (M-1)^{n}$. From (7), if the remainders are the same, the channel state is fixed. Next consider user $j = n+1, \ldots, K$. Notice that the length of subblocks of user $j$ is $(M-1)^2 - 1$, which is no smaller than $(M-1)^n$, the length of the building block of user $n$. In addition, channel of user $j$ remains fixed within each subblock. Thus, as long as symbols in each group of user $n$ belong to the same subblock of user $j$, then the channel state they see at user $j$ is the same. Recall that each group is created within every building block of user $n$. Therefore, if the boundary of the building block of user $n$ is aligned with that of subblock of user $j$, then each group is within the same subblock at user $j$. This can be easily verified since the length of the subblock of user $j$, $(M-1)^2 - 1$, is an integer multiple of that of the building block of user $n$, $(M-1)^n$. For example, as illustrated in Fig. 14, the building block of user $n$ is of length $(M-1)^n$ which is equal to the length of one subblock of user $n+1$. The channel remains constant within a subblock of user $n+1$. Therefore, all groups within a building block of user $n$ see the same channel value at receiver $n+1$. As a result, all groups satisfy the requirement of the alignment block structure.

**Step 3: Design Block 2:** Once Block 1 is designed, Block 2 can be easily determined. Recall that in Block 1, we create the first $M-1$ symbols of the alignment block. Therefore, in Block 2, each symbol serves as the last symbol for each group in Block 1 to create one alignment block. Since for each user, there are $(M-1)^K$ groups, a total of $K(M-1)^{K-1}$ symbols are needed for $K$ users in Block 2. Now we can determine the channel values in Block 2. We divide Block 2 into $K$ subblocks, one with length $(M-1)^{K-1}$. In subblock $n$, we provide the last symbol for user $n$. Therefore, user $n$’s channels are equal to $h^{[n]}(M)$ in subblock $n$. For all other users, each of $(M-1)^{K-1}$ symbols in subblock $n$ is set to be equal to the value of its corresponding group. As we have shown before, each group’s channel remains fixed at the undesired user. Thus, to determine each symbol in subblock $n$ for user $j = 1, \ldots, n-1, n+1, \ldots, K$, we can find the first time instant of its corresponding group in Block 1, then set it to be equal to the channel value at that time instant. From (9), we can see that the first time instant of the $i^{th}$ group in the $k^{th}$ building block for user $n$ is

$$t_1^{[n]}(k,i) = (k-1)(M-1)^n + i.$$  

Therefore, the channels of user $j = 1, \ldots, n-1, n+1, \ldots, K$ in the $n^{th}$ subblock in Block 2 are equal to $f_j(h^{[n]}(k,i))$.

With this scheme, each user achieves $M(M-1)^{K-1}$ DoF over $(M-1)^K + K(M-1)^{K-1}$ symbol extensions. Therefore, the normalized total DoF is equal to

$$\frac{KM(M-1)^{K-1}}{(M-1)^{2K} + K(M-1)^{K-1}} = \frac{M+K}{M+K-1}.$$ 

**Remark 2:** Note that the beamforming strategy does not require cooperation among antennas. Thus, the same achievable schemes derived for $K$-user $M \times 1$ MISO BC can be directly applied to the $M \times K X$ channel where there are $M$ transmitters and $K$ receivers. Each transmitter has a message for each receiver for a total of $MK$ messages in the network.

IV. ACHIEVABLE RATES FOR THE $K$-USER $M \times 1$ MISO BC WITH ZERO-FORCING INTERFERENCE AT THE RECEIVER

In this section, we derive a closed form expression for the rate achieved using the blind interference alignment scheme described in last section and with zero-forcing interference at the receiver. The result is presented in the following theorem.

**Theorem 2:** For the $K$-user $M \times 1$ MISO BC defined in Section II, the achievable sum rate with equal power allocation to each data stream and zero-forcing interference at each user, is

$$R = \sum_{k=1}^{K} \frac{1}{M+K-1} \times \log \det \left( I + \frac{(K + M - 1)P}{M^2 K} H[k] H[k] \right)$$

where $H[k] = [\frac{1}{\sqrt{K}} h^{[k]}(1) \cdots \frac{1}{\sqrt{K}} h^{[k]}(M-1)]^T$.

**Proof:** First, we illustrate how to cancel interference due to one user at all other users. Due to symmetry, let us consider user 1. Recall that the same $M$ symbols are transmitted to the desired user, user 1, over $M$ time slots in one alignment block of
that user. In addition, the channel states remain fixed at all other undesired users over these $M$ time slots. Therefore, the interference caused by user 1 over these time slots is the same at each undesired user. Moreover, in the last time slot of that alignment block, the transmitter only transmits signals to user 1. As a result, the received signal in that time slot at each undesired user is just the interference of user 1. Thus, to cancel interference caused by user 1, all other users can subtract the signal received in the last time slot of one alignment block from the signal received in the previous time slots of that alignment block. With this understanding, let us consider how users decode their symbols. For each user, symbols received across different alignment blocks are decoded separately. In other words, to decode $M$ symbols over one alignment block of one user, we do not use the signals received in the time slots corresponding to all other alignment blocks of that user. Now suppose we want to decode $M$ symbols over one alignment block of user $k$. Notice that in each of first $M - 1$ symbols (time slots) of the alignment block, the desired signal is interfered by all other $K - 1$ users. The signal received in the last symbol, corresponding to the one in Block 2, is interference free. This is because only user $k$’s data streams are transmitted in that time slot. To cancel the interference in each of $M - 1$ time slots, user $k$ needs to do $K - 1$ subtractions, each corresponding to cancel one of $K - 1$ interferers. As a result, the noise $K$ times stronger in these time slots. Normalizing the noise results in a scaling of a factor of $\frac{1}{\sqrt{K}}$ of the channels. Therefore, after eliminating interference, the received signal in one alignment block of user $k$ is

$$y^{k} = \begin{bmatrix} \frac{1}{\sqrt{K}} h^{k}[1] \\ \frac{1}{\sqrt{K}} h^{k}[2] \\ \vdots \\ \frac{1}{\sqrt{K}} h^{k}[M - 1] \\ h^{k}[M] \end{bmatrix} + \tilde{x}$$

(11)

where $\tilde{x} \sim CN(0, I)$. With equal power allocation to each data stream, the rate achieved for $M$ data streams in one alignment block is

$$R^{k} = E \left[ \log \det \left( I + \frac{(K + M - 1)P}{M^2 K} H^{k} H^{k\dagger} \right) \right].$$

(12)

Since there are a total of $(M - 1)^K$ alignment blocks, and a total of $(M - 1)^K + K(M - 1)^{K - 1}$ time slots, the normalized rate for user $k$ is

$$R^{k} = \frac{(M - 1)^{K - 1}}{(M - 1)^K + K(M - 1)^{K - 1}} \cdot E \left[ \log \det \left( I + \frac{(K + M - 1)P}{M^2 K} H^{k} H^{k\dagger} \right) \right]$$

$$= \frac{1}{M + K - 1} \cdot E \left[ \log \det \left( I + \frac{(K + M - 1)P}{M^2 K} H^{k} H^{k\dagger} \right) \right].$$

V. CONCLUSION

Recent work has shown that channel correlations can be exploited to achieve interference alignment even when the transmitter has no information about the exact channel values [6]. This work shows that channel temporal correlations required in [6] can be created, and thus interference alignment can be achieved in practice. The idea is to manipulate channels through antenna selection. By switching antennas during transmission, different temporal correlations can be created at different users. With these new insights, we provide a systematic way to achieve blind interference alignment for the $K$-user $M \times 1$ MISO BC, possibly with multicast traffic so that the outer bound of $\frac{MK}{M + K - 1}$ DoF is achieved. The coding scheme is essentially a simple repetition code over a finite number of symbols. The simplicity of the coding scheme allows us to write closed form achievable rate expressions at any SNR.

In this paper, we focus only on blind interference alignment schemes that exploit receive antenna selection. This does not imply that transmit antenna selection is never useful. Indeed, the two-user MIMO interference channel blind interference alignment example presented in [6] is easily seen to be possible through antenna switching only at one transmitter. Exploiting SCBC schemes for other wireless networks is also an interesting future work.

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