Can the use of reconfigurable antennas overcome the CSI bottleneck for FDD Massive MIMO?

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***Abstract*—Massive MIMO has been proposed as a means of achieving huge spectral efficiency in cellular networks. However, its implementation is usually based on linear precoding schemes based on Channel State Information (CSI). For this reason, the use of the Time Division Duplex (TDD) mode is typically assumed since it removes the dependence on the number of antennas at the transmitter to acquire CSI. Unfortunately, most of the cellular network operate in the Frequency Division Duplex (FDD) mode in which the acquisition of CSI depends proportionally on the number of transmit antennas. For this reason, the implementation of Massive MIMO is handicapped to few transmit antennas for the current cellular networks. In this paper we study the use of reconfigurable antennas that can switch among a set of preset modes, i.e., radiation patterns, through Blind Interference Alignment schemes. We show that it is a solution to exploit a very large number of antennas at the transmitter side, overcoming the bottleneck that today constitutes the acquisition of CSI in Massive MIMO systems operating in FDD mode.**

I. Introduction

Massive Multiple-Input Multiple-Output (MIMO) systems was proposed as a means of increasing both the spectral and the energy efficiency [1]. Basically, Massive MIMO considers the use of Base Stations (BSs) equipped with a large number of transmit antennas to serve a significantly smaller number of users. To take advantage of this exceed of antennas close-loop beamforming techniques such as Liner Zero Forcing Beamforming (LZFB) or Conjugate Beamforming (CB) are usually considered since they achieve the optimal Degrees of Freedom (DoF), also known as multiplexing gain, and array gain [2]. In particular, considering a BS equipped with M antennas serving K single-antenna users, the achievable sum-DoF equals $\min(M,K)$, in addition, under the conditions of favorable propagation an array gain of M would be attainable. However, these linear precoding schemes are usually based on Channel State Information (CSI).

Massive MIMO usually goes hand in hand with the use of the Time Division Duplex (TDD) mode where both uplink and downlink occurs over the same bandwidth at different time slots. For the TDD mode, the uplink CSI at the receiver (CSIR) can be estimated by sending K orthogonal pilots from the users to the BS. By taking advantage of the reciprocity between downlink and uplink, the BS can also obtain downlink CSI at the transmitter (CSIT) after a calibration to compensate the hardware non-symmetry. However, most of the contemporary networks operate in the Frequency Division Duplex (FDD) mode, i.e., uplink and downlink allocated in different frequency bands. For the FDD mode, achieving downlink CSIT requieres the transmission of M orthogonal pilots to estimate the CSIR at each user plus M additional pilots to feed this information back to the transmitter. Moreover, K uplink pilots are also required to obtain CSIR of the users at the BS. Since this training sequence plus data transmission must occur within a channel coherence block, the number of antennas at the BS is restricted to allow the acquisition of CSI, which depends proportionally on M. This restriction is usually referred to as *CSI bottleneck for FDD in Massive MIMO systems* [2].

Parallel to the development of Massive MIMO systems, the design of reconfigurable antennas has been proposed as a means of maximizing the network connectivity [3]. Basically, a reconfigurable antenna is an antenna that can change its characteristics by dynamically modifying its geometry, e.g., micro/nano electromechanical switches (MEMS/NEMS) or solid state switches [4], [5]. Particularly, the use of reconfigurable antennas allow the implementation of Blind Interference Alignment (BIA) schemes that obtain a growth of DoF regarding the number of users in absence of CSIT. BIA is based on exploiting the channel correlations over a specific heterogeneous block fading channel usually denoted as supersymbol. To generate these channel variations, the use of reconfigurable antennas is proposed in [6]. There exist several BIA schemes for cellular networks to align the transmission among clusters of BSs [7], maximize the sum-DoF in homogenous networks [8] or provide diversity [9]. However, due to the large number of transmit antennas considered for Massive MIMO systems, their implementation usually requires a extremely large coherence block as well as users equipped with reconfigurable antennas that can switch among a large number of preset modes. As a result, the use of reconfigurable antennas through BIA schemes has not yet been considered for Massive MIMO.

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In this work we study the implementation of reconfigurable antennas at each user with the aim of reducing the costs of providing CSI for Massive MIMO systems operating in FDD. We devised a novel BIA approach based on antenna grouping that eliminates the dependence of the number of pilots on the number of antennas in the BS when providing CSI in the FDD mode. Moreover, unlike the existing BIA schemes, its implementation can yield spatial diversity gain at the same time as handling limitations on the coherence time of the channel and/or the number of preset modes of the reconfigurable antennas. Simulation results illustrate the benefits of the proposed scheme considering finite coherence blocks regarding CSIT-based schemes such as LZFB and CB.

II. System model and baseline

We consider a downlink Broadcast Channel (BC) where a BS equipped with M antennas transmits to K active users where $M >> K$. Each user k, $k \in \{1,\dots,K\}$, is equipped with a reconfigurable antenna that can switch among N_k preset modes. At a conceptual level, the antenna of each user is capable of switching among N_k independent preset modes providing distinct radiation patterns each. The signal transmitted by the BS can be written in a vector form as $\mathbf{x} = \left[x_1,\dots,x_M\right]^T \in \mathbb{C}^{M\times 1}$. Thus, the signal received by the user k is given by

$$y^{[k]}[t] = \mathbf{h}^{[k]}\left(l^{[k]}[t]\right)^T \mathbf{x}[t] + z^{[k]}[t], \tag{1}$$

where $\mathbf{h}^{[k]}\left(l^{[k]}[t]\right) \in \mathbb{C}^{M\times 1}$ is the channel vector between the M antennas of the BS and user k when it selects the preset mode $l^{[k]}[t]$ at time t and $z^{[k]}[t] \sim \mathcal{CN}(0,1)$ is additive white Gaussian noise. The channel between the i-th antenna of the BS and the user k at present mode l is denoted as $h_i^{[k]}\left(l^{[k]}[t]\right)$, $i \in \{1,\dots,M\}$, so that

$$\mathbf{h}^{[k]}\left(l^{[k]}[t]\right) = \left[h_1^{[k]}\left(l^{[k]}[t]\right) \quad \dots \quad h_M^{[k]}\left(l^{[k]}[t]\right)\right]^T. \tag{2}$$

The channel coefficients are i.i.d. complex Gaussian random variables of zero mean. Moreover, the transmitted signal is subject to an average power constraint $E\left\{\|\mathbf{x}(t)\|^2\right\} \le P$. We assume that the preset mode switching pattern functions, $l^{[k]}[t]$, are known beforehand and that the BS does not have any CSI other than the coherence block, i.e., the time/frequency where the channel response is approximately static. The dimensions of the coherence block are given by the coherence bandwidth B_c and the coherence time T_c, which fit $\tau = B_c T_c$ transmission symbols. For simplicity, we focus on the temporal dimension without loss of generality. Therefore, each symbol extension corresponds to a time slot. Nevertheless, all results can be easily applied to the frequency domain.

A. Massive MIMO baseline

Considering downlink transmission under the conditions of favorable propagation, i.e., the channel vectors between the users and the BSs are pairwisely orthogonal, the sum-capacity is given by

$$C_{\Sigma_{\text{csi}}} = \tau_{\text{csi}}\,\text{DoF}\,\log\left(1+\rho_{\text{down}}c_{\text{array}}\right) \tag{3}$$

	1	2	3
k=1	$\mathbf{h}^{[1]}(1)$	$\mathbf{h}^{[1]}(2)$	$\mathbf{h}^{[1]}(1)$
k=2	$\mathbf{h}^{[2]}(1)$	$\mathbf{h}^{[2]}(1)$	$\mathbf{h}^{[2]}(2)$

Fig. 1. Supersymbol structure for sBIA scheme. $M = 2$, $K = 2$.

where $\text{DoF} = \min(M,K) = K$ corresponds to the multiplexing gain, $c_{\text{array}} = M$ represents the array gain and ρ_{down} is the Signal-to-Noise Ratio (SNR) at the receiver. Besides, in (3), τ_{csi} denotes the ratio between data transmission and the length of the full frame, i.e., including the training sequence required to provide CSI. Thus, it would be possible to obtain a huge spectral efficiency when M and K are large. However, its achievability through linear processing based on CSIT is subject to the CSI bottleneck when operating in the FDD mode. That is, the number of transmit antennas must be lower enough to allow the required training for providing CSI plus data transmission within a coherence block. In the following we study if this performance is achievable by using reconfigurable antennas at the receiver through BIA-based schemes.

III. Blind Interference Alignment for BC

In order to introduce some useful notation and provide a starting point on the use of reconfigurable antennas through BIA-based schemes, we start with a brief review of the BIA scheme derived in [6], which is referred to as standard BIA (sBIA). For illustrative purposes, consider first the 2-user 2×1 MISO BC. Thus, the supersymbol is as shown in Fig. 1 and the transmitted signal is

$$\mathbf{X} = \begin{bmatrix}\mathbf{x}[1]\\ \mathbf{x}[2]\\ \mathbf{x}[3]\end{bmatrix} = \begin{bmatrix}\mathbf{I}_2\\ \mathbf{I}_2\\ \mathbf{0}_2\end{bmatrix}\mathbf{u}_\ell^{[1]} + \begin{bmatrix}\mathbf{I}_2\\ \mathbf{0}_2\\ \mathbf{I}_2\end{bmatrix}\mathbf{u}_\ell^{[2]}, \tag{4}$$

where $\mathbf{x}[t] \in \mathbb{C}^{2\times 1}$ is the signal transmitted during the symbol extension t, $\mathbf{u}_\ell^{[k]} = \left[u_{1,\ell}^{[k]}, u_{2,\ell}^{[k]}\right] \in \mathbb{C}^{2\times 1}$ is the symbol sent to user k during its ℓ-th alignment block, with $u_{i,\ell}^{[k]}$ equal to the symbol transmitted by the i-th antenna of the BS. Moreover, in (4), $\mathbf{I}_2$ and $\mathbf{0}_2$ denote the 2×2 identity and zero matrix, respectively. For a coherence block large enough the channel coefficients at any preset mode do not depend on the temporal index. Thus, focusing on the user 1, the received signal can be written as

$$\begin{bmatrix}y^{[1]}[1]\\ y^{[1]}[2]\\ y^{[1]}[3]\end{bmatrix} = \underbrace{\begin{bmatrix}\mathbf{h}^{[1]}(1)^T\\ \mathbf{h}^{[1]}(2)^T\\ \mathbf{0}\end{bmatrix}}_{\text{rank=2}}\mathbf{u}_1^{[1]} + \underbrace{\begin{bmatrix}\mathbf{h}^{[1]}(1)^T\\ \mathbf{0}\\ \mathbf{h}^{[1]}(1)^T\end{bmatrix}}_{\text{rank=1}}\mathbf{u}_1^{[2]} + \begin{bmatrix}z^{[1]}[1]\\ z^{[1]}[2]\\ z^{[1]}[3]\end{bmatrix}. \tag{5}$$

Notice that the desired symbol $\mathbf{u}_1^{[1]}$ is contained in a rank-2 matrix while the interference is aligned in a rank-1 matrix. As a consequence, the user 1 can employ the third symbol extension to measure and cancel the interference caused by the transmission of $\mathbf{u}_1^{[2]}$ during the first symbol extension. Furthermore, the signal received by user 1 is free of interference during the second symbol extension. After subtracting the interference the received signal for user 1 is

$$\begin{bmatrix}y^{[1]}[1]-y^{[1]}[3]\\ y^{[1]}[2]\end{bmatrix} = \begin{bmatrix}\mathbf{h}^{[1]}(1)^T\\ \mathbf{h}^{[1]}(2)^T\end{bmatrix}\mathbf{u}_1^{[1]} + \begin{bmatrix}z^{[1]}(1)-z^{[1]}(3)\\ z^{[1]}(2)\end{bmatrix}. \tag{6}$$

Since the interference is completely removed, 2 DoF are decodable by solving the previous linear problem. Similarly, the user 2 obtains the 2 DoF of $\mathbf{u}_1^{[2]}$ over the symbol extensions $\{1,3\}$ and measuring the interference due to the transmission of $\mathbf{u}_1^{[1]}$ in the symbol extension $\{2\}$. Therefore, $\frac{4}{3}$ DoF are achievable by using sBIA. It is worth to remark that although sBIA does not require CSIT, the coherence detection of the corresponding symbol $\mathbf{u}_\ell^{[k]}$ is based on CSIR knowledge. However, considering a FDD cellular network providing this information results less demanding than CSIT.

For the K-user $M\times 1$ MISO BC, the sBIA supersymbol comprises $(M-1)^{K-1}$ alignment blocks, $\ell\in\{1,\ldots,(M-1)^{K-1}\}$, per user. Each of these alignment blocks is formed by M symbol extensions over which a BS transmits M symbols to a specific user. During an alignment block of user k, its antenna switches among M preset modes while the preset mode of all other users remains constant. In order to generate the antenna switching pattern of channel modes $\left\{\mathbf{h}^{[k]}(l)\right\}_{l=1}^{M}$ that satisfies the criterion of decodability and alignment, each user k will employ the temporal correlation function $f_{K,M}^{[k]}(t)$. Moreover, the transmission of the M-symbol vector $\mathbf{u}_\ell^{[k]}$ is undertaken according to a beamforming matrix $\mathbf{B}_{K,M}^{[k]}$ composed of $\{0,1\}$ values, i.e., it is determined without the need for CSIT. In order to ensure that the users can measure and remove the interference, the beamforming matrices are designed such that the supersymbol can be divided in two blocks. During Block 1 simultaneous transmission to all users is employed while the BS transmits to the users in an orthogonal fashion during Block 2. These blocks comprise $(M-1)^K$ and $K(M-1)^{K-1}$ symbol extensions, respectively. Therefore, the length of the sBIA supersymbol is

$$\mathcal{L}_{\text{SS}}=(M-1)^K+K(M-1)^{K-1}. \tag{7}$$

Assuming that the channel remains constant over the whole supersymbol, the aforementioned conditions for decodability and alignment let each user attain M DoF in each alignment blok. Therefore, the normalized sum-DoF is equal to

$$\text{DoF}_{\text{sBIA}}=\frac{KM(M-1)^{K-1}}{(M-1)^K+K(M-1)^{K-1}}=\frac{MK}{M+K-1}, \tag{8}$$

which corresponds to the optimal sum-DoF without CSIT [10]. Focused on Massive MIMO, notice that $\text{DoF}_{\text{sBIA}}\approx K$ when $M>>K$, which corresponds with the DoF achieved by linear precoding schemes based on CSIT.

IV. Blind Interference Alignment for Massive MIMO

Although the sBIA scheme is DoF-optimal in the absence of CSIT, there exist several practical issues for the implementation of BIA in Massive MIMO systems. First, to ensure that the decodability and alignment conditions are satisfied, the channel needs to be constant along the supersymbol, whose length grows exponentially with respect to the number of users and with base equal to the number of transmit antennas (see (7)). For instance, the physical channel must remain constant during 10^4 symbol extensions for $M=100$ transmit antennas and only $K=2$ users. Second, the achievability of (8) is subject to having M distinct preset modes in the reconfigurable antenna of each user, which can be highly challenging for large values of M. Moreover, when operating in the FDD mode a preamble of M pilots needs to be transmitted in an orthogonal fashion in order to provide CSIR and to let each user decode the symbols transmitted over each of its alignment blocks. In this section we present a BIA scheme for Massive MIMO (mBIA) based on the use of reconfigurable antennas that provides a trade-off among the number of required pilots, preset modes, diversity and achievable DoF.

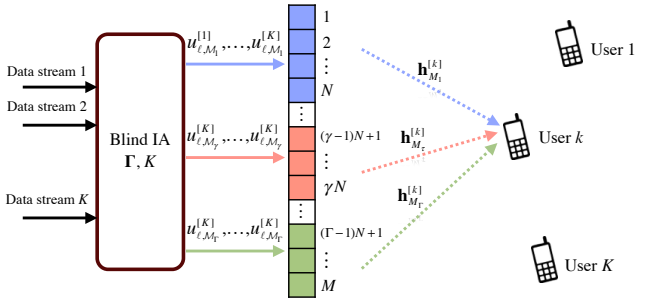


Fig. 2. Scheme of the antenna grouping for the proposed mBIA scheme.

A. Toy example

We first consider a toy example where the BS is equipped with $M=100$ antennas and $K=2$ users. Instead of using sBIA, which would consider 100 preset modes per reconfigurable antennas and close to 10^4 symbol extensions, we propose to divide the set of transmit antennas in 2 groups composed of $N=50$ antennas each referred to as $\mathcal{M}_1$ and $\mathcal{M}_2$, respectively. All the antennas belonging to the same group $\mathcal{M}_\gamma$, $\gamma\in\{1,2\}$, transmit the same signal. Thus, let us consider the same switching pattern as in the 2-user 2×1 MISO BC (see Fig. 1). During the 3 symbol extensions that comprise the proposed supersymbol, the transmitted signals are given by

$$\mathbf{X}=\left[\begin{bmatrix}\mathbf{I}_2\\ \mathbf{I}_2\\ \mathbf{0}_2\end{bmatrix}\otimes\mathbf{1}_{50}\right]\bar{\mathbf{u}}_1^{[1]}+\left[\begin{bmatrix}\mathbf{I}_2\\ \mathbf{0}_2\\ \mathbf{I}_2\end{bmatrix}\otimes\mathbf{1}_{50}\right]\bar{\mathbf{u}}_1^{[2]}, \tag{9}$$

where $\bar{\mathbf{u}}_\ell^{[k]}=[u_{\mathcal{M}_1,\ell}^{[k]}\,u_{\mathcal{M}_2,\ell}^{[k]}]\in\mathbb{C}^{2\times 1}$ and $u_{\mathcal{M}_\gamma,\ell}^{[k]}$ is the symbol transmitted to user k by the set of antennas $\mathcal{M}_\gamma$ during the ℓ-th alignment block, with $\mathbf{1}_n$ being a $n\times 1$ vector of ones and $\otimes$ denoting the Kronecker product. Following the same procedure as described in Section III, the interference can be removed by simply measuring and subtracting it from the corresponding symbol extensions. Thus, omitting the noise, the signal of user 1 after interference subtraction is

$$\begin{bmatrix}y^{[1]}[1]-y^{[1]}[3]\\ y^{[1]}[2]\end{bmatrix}=\begin{bmatrix}\mathbf{h}_{\mathcal{M}_1}^{[1]}(1) & \mathbf{h}_{\mathcal{M}_2}^{[1]}(1)\\ \mathbf{h}_{\mathcal{M}_1}^{[1]}(2) & \mathbf{h}_{\mathcal{M}_2}^{[1]}(2)\end{bmatrix}\begin{bmatrix}\mathbf{1}_{50}\,u_{\mathcal{M}_1,1}^{[1]}\\ \mathbf{1}_{50}\,u_{\mathcal{M}_2,1}^{[1]}\end{bmatrix}, \tag{10}$$

where $\mathbf{h}_{\mathcal{M}_\gamma}^{[k]}(l)\in\mathbb{C}^{N\times 1}$ is the channel between the group of antennas $\mathcal{M}_\gamma$ and the user k at preset mode l. In consequence, the received signal given by (10) can be written as

$$\begin{bmatrix}y^{[1]}[1]-y^{[1]}[3]\\ y^{[1]}[2]\end{bmatrix}=\mathbf{H}_\Sigma^{[k]}\bar{\mathbf{u}}_1^{[1]}=\begin{bmatrix}\mathbf{h}_\Sigma^{[1]}(1)^T\\ \mathbf{h}_\Sigma^{[1]}(2)^T\end{bmatrix}\bar{\mathbf{u}}_1^{[1]}, \tag{11}$$

where $\mathbf{H}_{\Sigma}^{[k]} \in \mathbb{C}^{2\times 2}$ is a full rank matrix and $\mathbf{h}_{\Sigma}^{[k]}(l) = \left[h_{\Sigma,1}^{[k]}(l)\; h_{\Sigma,2}^{[k]}(l)\right]^T \in \mathbb{C}^{2\times 1}$ is the vector given by the sum of the elements of $\mathbf{h}_{\mathcal{M}_\gamma}^{[k]}(l)$ for each group γ denoted as $h_{\Sigma,\gamma}^{[k]}(l) = \sum_{i=(\gamma-1)N+1}^{\gamma N} h_i^{[k]}(l)$. Hence, the user 1 can decode $\bar{\mathbf{u}}_1^{[1]}$ free of interference. Following a similar procedure user 2 can remove the interference and decode $\bar{\mathbf{u}}_1^{[2]}$.

The proposed mBIA scheme achieves $\frac{4}{3}$ DoF in the considered setting where the BS is equipped with $M = 100$ antennas. Although mBIA introduces a penalty in the achievable DoF as compared to sBIA, the proposed scheme offers several advantages that are of great value for Massive MIMO systems. First, notice that it can provide spatial diversity gain since each symbol is received from N paths. Secondly, when operating in FDD, the number of pilots required for providing CSIT able to estimate $\mathbf{H}_{\Sigma}^{[k]}$ (see (11)) has been reduced to 2 pilots instead of $M = 100$ pilots, as would be required by other transmission schemes such as sBIA, CB, or LZFB.

B. General case

For the general case the mBIA scheme divides the M transmit antennas of the BS in Γ sets $\{\mathcal{M}_1,\ldots,\mathcal{M}_\Gamma\}$ of $N = \lfloor \frac{M}{\Gamma} \rfloor$ antennas each as is shown in Fig. 3(c)[1]. Thus, the group M_γ, $\gamma = \{1,\ldots,\Gamma\}$, comprises the transmit antennas indexed as $\{(\gamma-1)N+1,\ldots,\gamma N\}$. The antennas of a group $\mathcal{M}_\gamma$ always transmit the same signal. As a result, the supersymbol structure for mBIA is given by the sBIA scheme for a K-user $\Gamma\times 1$ MISO BC, which is defined by $f_{K,\Gamma}^{[k]}(t)$ [6]. Thus, the length of the supersymbol is given by (7) with M replaced by Γ.

According to the proposed antenna grouping, the $\Gamma\times 1$ symbol vector $\mathbf{u}_\ell^{[k]}$, $\ell \in \{1,\ldots,(\Gamma-1)^{K-1}\}$, is transmitted in the ℓ-th alignment block of the considered supersymbol. However, each group M_γ, $\gamma \in \{1,\ldots,\Gamma\}$, contains N antennas, which provides distinct channel paths. Consequently, the beamforming matrix $\mathbf{B}_{K,\Gamma}^{[k]}$ for the sBIA scheme must be extended N times. The signal transmitted by the BS is then

$$\mathbf{X} = \sum_{k=1}^{K} \left[\mathbf{B}_{K,\Gamma}^{[k]} \otimes \mathbf{1}_N\right] \bar{\mathbf{u}}^{[k]}, \tag{12}$$

where $\bar{\mathbf{u}}^{[k]} = \left[\bar{\mathbf{u}}_1^{[k]} \ldots \bar{\mathbf{u}}_{(\Gamma-1)^{K-1}}^{[k]}\right]$.

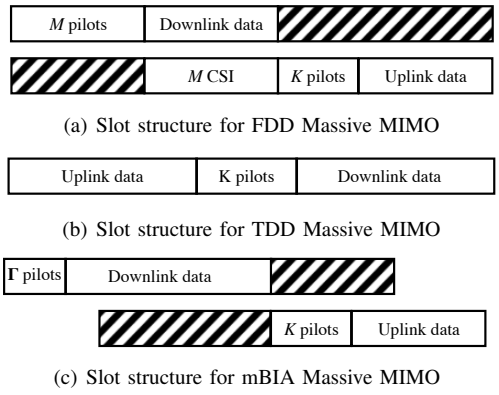
The achievable sum-DoF is given by the K-user $\Gamma\times 1$ MISO BC, which is given by

$$\mathrm{DoF}_{\mathrm{mBIA}} = \frac{\Gamma K}{\Gamma + K - 1}. \tag{13}$$

Notice that the mBIA approach involves a penalty regarding the optimal DoF, which is given by (8). Specifically, the penalty in terms of DoF because of the antenna grouping is

$$\Delta\mathrm{DoF} = \frac{K(K-1)(M-\Gamma)}{(M+K-1)(\Gamma+K-1)}. \tag{14}$$

[1]In the case that $\frac{M}{\Gamma}$ is not an integer, $N = \lfloor \frac{M}{\Gamma} \rfloor$ is the number of antennas for groups $\mathcal{M}_\gamma$, with $\gamma \in \{1,\ldots,\Gamma-1\}$, while the group $\mathcal{M}_\Gamma$ is formed by the last $M-(\Gamma-1)N$ antennas.



(a) Slot structure for FDD Massive MIMO

(b) Slot structure for TDD Massive MIMO

(c) Slot structure for mBIA Massive MIMO

Fig. 3. Transmission structure for FDD, TDD and mBIA using renconfigurable antennas for Massive MIMO.

C. The costs of Channel State Information

The use of CSIT-based schemes operating in the FDD mode involves transmission in close-loop given by the steps for CSI acquisition and data transmission as it is shown in Fig. 3(a). In this case, at least $2M+K$ pilots are required for the implementation of CSIT-based schemes. To avoid the dependence on the number of antennas at the BS, the TDD mode, which only requires K pilots by using the reciprocity between downlink and uplink for providing CSI, has been typically proposed for Massive MIMO systems [1], [2]. Hence, the slot structure for TDD transmission is as shown in Fig. 3(b). Curiously, the use of reconfigurable antennas allow to break the aforementioned close-loop for FDD while providing multiplexing gain. For comparison purposes, a generic slot structure for mBIA is shown in Fig. 3(c). Specifically, by using the proposed mBIA scheme in the FDD mode, only Γ pilots are required to estimate CSIR[2] while CSIT is avoided. However, the channel must remain constant during the transmission of the mBIA supersymbol, which comprises $(\Gamma-1)^{K-1}(\Gamma+K-1)$ symbol extensions. Similarly, K pilots are transmitted via uplink.

For FDD, TDD and mBIA, assuming that half of the communication resources are devoted to data transmission Fig. 4 shows the system dimensions, i.e., the number of service antennas and active users, that verify the constraints imposed by the coherence block of the channel τ on the time available for providing CSI. First, notice the overwhelming advantage of TDD over FDD since the amount of users served by the BS is not constrained by the number of antennas, which is typically large in Massive MIMO. Considering this fact, it can be seen that the proposed mBIA scheme also eliminates the dependence between the number of antennas at the BS and the users. By varying Γ within the set of values that satisfy the coherence block requirements, notice that the proposed scheme can provide a trade-off between the number of served users, spatial diversity gain and achievable sum-DoF. In particular, by decreasing Γ, we can obtain more spatial diversity gain and serve more users subject to some limited coherence block. On the contrary, notice that smaller values of Γ introduce higher penalties in the achievable DoF. For $\Gamma > 2$ the supersymbol length can be a hurdle for the implementation

[2]It may be achieved by the single RF chain, over-sampling pilots with factor N_t [7] or using schemes such as [11] with a N_t linear dimensional cost.

of mBIA. However, beyond the mere use of reconfigurable antennas for Massive MIMO, there exist several BIA schemes, e.g. [12] or [13], to manage the supersymbol length that can be combined with mBIA for Massive MIMO.

D. The use of Alamouti codes for Massive BIA

The mBIA scheme allows to obtain diversity gain since each antenna of the same group transmits exactly the same signals. However, transmission schemes able to exploit this potential diversity gain would be required. In this sense, each entry of the equivalent channel $h^{[k]}_{\Sigma,\gamma}(l)$ corresponds to the sum of N i.i.d Gaussian random variables of zero mean and unit variance (see (11)). In particular, $h^{[k]}_{\Sigma,\gamma}(l) \sim \mathcal{CN}(0, N\sigma_h^2)$ with σ_h^2 equal to the variance of an individual channel coefficient $h^{[k]}_i(l)$ and $h^{[k]}_{\Sigma,\gamma}(l)$ denoting the γ-th entry of $\mathbf{h}^{[k]}_{\Sigma}(l)$. Considering the probability density of the Gaussian distributions notice that the channel coefficients are mostly concentrated around the zero mean. In other words, the mBIA scheme increases the variance of the equivalent channel coefficients, i.e., regarding the simple K-user $\Gamma \times 1$ MISO BC, while the mean remains equal to zero.

With the aim of taking advantage of the increased variance of the channel coefficients achieved by mBIA, we consider the use of Maximum-Ratio Combining methods. Specifically, we focused on the use of mBIA combined with the Alamouti codes. Following the lines of [9], Alamouti coding can be employed by creating $\Gamma' = 2\Gamma$ groups of antennas and repeating twice the supersymbol defined by $f^{[k]}_{K,\Gamma}$. By combining mBIA and Alamouti coding, each desired symbol $u^{[k]}_{\ell,\mathcal{M}_\gamma}$ is transmitted through an equivalent channel given by

$$\widetilde{h}^{[k]}_{\Sigma} = \sum_{\gamma=1}^{\Gamma'} \left| \sum_{i=(\gamma-1)N+1}^{\gamma N} h^{[k]}_i(\lceil \tfrac{\gamma}{2} \rceil) \right|^2, \tag{15}$$

which corresponds to the sum of 2Γ random variables, each with a Chi-square probability distribution of mean $\frac{M}{2\Gamma}$ and variance $\frac{M}{\Gamma}$. Since the addition of independent Chi-square variables is also Chi-square distributed, each symbol $u^{[k]}_{M_\gamma,\ell}$ is decoded through an equivalent channel Chi-square distributed with mean M. Thus, the channel distribution has improved considerably. Moreover, since Alamouti codes are rate 1, this approach does not involve any penalty in the achievable DoF or data rate as compared to the use of mBIA for M antennas divided in Γ groups serving K users. This additional diversity improves considerably the decoding process of the transmitted symbols in the mBIA scheme.

V. Achievable Rates for mBIA

After interference subtraction, the signal of the user k can be written as $\tilde{\mathbf{y}}^{[k]} = \mathbf{H}^{[k]}_{\Sigma}\bar{\mathbf{u}}^{[k]}_{\ell} + \tilde{\mathbf{z}}^{[k]}$, where

$$\mathbf{H}^{[k]}_{\Sigma} = \left[\text{col}\left\{ \sum_{i=1}^{N} h^{[k]}_i(l) \right\}_{l=1}^{\Gamma}, \ldots, \text{col}\left\{ \sum_{i=M-N+1}^{M} h^{[k]}_i(l) \right\}_{l=1}^{\Gamma} \right] \tag{16}$$

and $\tilde{\mathbf{z}}^{[k]}$ is the resulting noise after the increase due to the the interference subtraction step. Assuming constant power allocation [7], the achievable sum-rate is

$$R^{[k]} = \frac{1}{\Gamma + K - 1}\mathbb{E}\left[\log\det\left(\mathbf{I} + \frac{P}{\Gamma}\mathbf{H}^{[k]}_{\Sigma}{\mathbf{H}^{[k]}_{\Sigma}}^{H}\mathbf{R}_z^{-1} \right) \right], \tag{17}$$

where $\mathbf{R}_z = \begin{bmatrix} (2K-1)\mathbf{I}_{\Gamma-1} & \mathbf{0} \\ \mathbf{0} & 1 \end{bmatrix}$.

In comparison with the ideal sum-capacity for Massive MIMO (see (3)), the sum-rate achievable by using mBIA can be reformulated as

$$R_{\Sigma_{\text{mBIA}}} \approx \tau_{mBIA}\,\text{DoF}_{\text{mBIA}} \log\left(1 + \rho_{\text{mBIA}} c_{\text{reconf}}\right), \tag{18}$$

where $\text{DoF}_{\text{mBIA}} = \frac{\Gamma K}{\Gamma + K - 1}$ is the multiplexing gain achieved by use of reconfigurable antennas through the proposed mBIA scheme considering Γ groups of antennas, ρ_{mBIA} is the average SNR taking into consideration the noise increase due to the interference substraction and c_{reconf} represents the antenna gain achieved by the design and use of the reconfigurable antenna. In this sense, notice that it corresponds to the array gain in (3). However, the complexity of achieving c_{reconf} corresponds to the design of the reconfigurable antenna at the receiver side instead of linear processing schemes based on CSIT. Moreover, in (18), τ_{mBIA} denotes the ratio between the symbol extensions for data transmission and the length of the full transmitted frame considering the pilots required to provide CSIR under the proposed mBIA scheme.

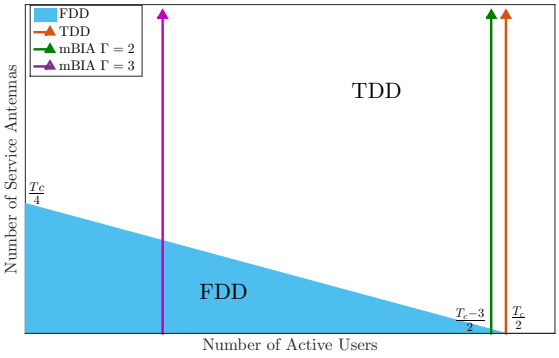


Fig. 4. For Massive MIMO in FDD, TDD, and mBIA, dependence between the number of transmit antennas and the number of users that verifies the constraint imposed by the coherence block when providing CSI.

VI. Simulation Results

The sum-rate achieved by the proposed mBIA scheme regarding the number of antennas at the BS is depicted in Fig. 5 and Fig. 6 for a FDD Massive MIMO system with $K = 6$ users subject to transmission within a coherence block of length τ. Moreover, for comparison purposes the sum-rate achieved by CSIT-based schemes such as LZFB and CB are also shown for the same simulation parameters. In Fig. 5 a coherence block of $\tau = 256$ symbols and a SNR = 6dB is assumed. First, it is interesting to remark that mBIA for $\Gamma = 2$ groups achieves a non-decreasing rate regarding the number of antennas at the BS. On the other hand, there exists a point where the rate achieved by CB or LZFB decreases because of the extremely large training period required to provide CSIT. Notice that zero-rate is achieved by LZFB and CB when the training sequence exceeds the length of the coherence block.

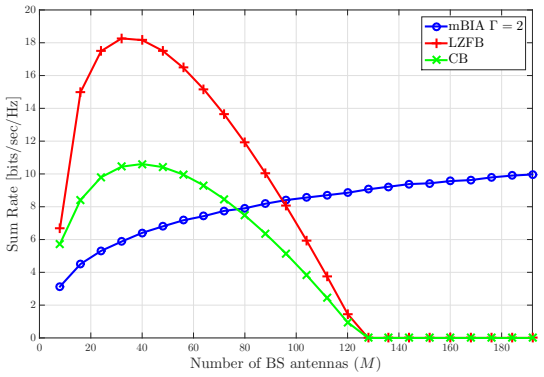


Fig. 5. Achievable sum-rate for mBIA in comparison with LZFB and CB for Massive MIMO regarding the number of BS antennas. $K = 6$, SNR = 6 dB, $\tau = 256$ symbols.

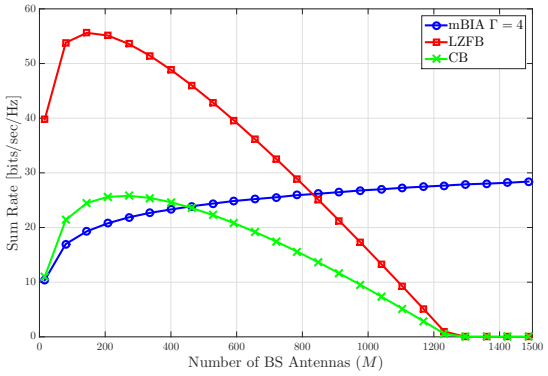


Fig. 6. Achievable sum-rate for mBIA in comparison with LZFB and CB for Massive MIMO regarding the number of BS antennas. $K = 6$, SNR = 18 dB, $\tau = 2500$ symbols.

In this particular case, mBIA outperforms CB and LZFB for a number of transmit antennas at the BS equals 78 and 95, respectively. For a larger coherence block of $\tau = 2500$ symbols and a SNR = 18 dB, it can be seen that more antennas can be employed at the BS for CSIT-based schemes. However, the sum-rate follows a similar shape as the previous case regarding the increase of the number of antennas at the BS. It can be seen that the costs of providing CSIT counterbalances the benefits of the CSIT-based schemes. As a consequence, use of reconfigurable antennas jointly with mBIA allows to use higher number of antennas at the BS.

Fig. 7 shows the symbol error rate (SER) for mBIA assuming $K = 6$ users and QPSK transmission. First, it can be seen that the use of Massive MIMO combined with mBIA attains a satisfactory performance at low SNR values. For instance, a SNR operating point below 6 dB provides a SER below 10^{-2} in most of the cases. Moreover, it is possible to check the benefits of employing Alamouti coding in the proposed mBIA scheme. This approach provides a coding gain of 8 dB and 12 dB for 30 and 300 transmit antennas in the BS, respectively.

VII. Conclusions

In this paper we consider the use of reconfigurable antennas with the aim of overcoming the CSI bottleneck in FDD systems through BIA. Based on antenna grouping, we present

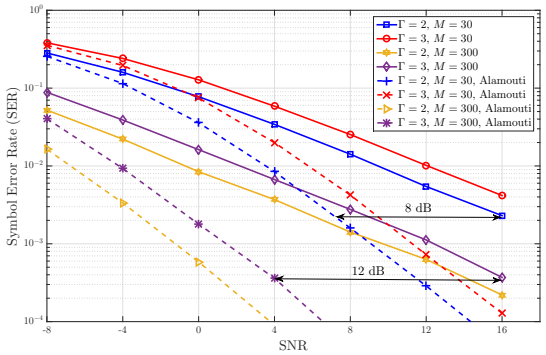


Fig. 7. Symbol Error Rate for the mBIA scheme with $K = 6$ users.

a BIA scheme for FDD Massive MIMO systems that removes the dependence on the number of antennas at the BS to satisfy the coherence block requirements. As a consequence, it results suitable for systems operating in the FDD mode where CSIT-based schemes requiere a training sequence whose length is proportional to the number of antennas at the BS. Moreover, the proposed schemes obtains a trade-off among multiplexing gain, spatial diversity gain and costs of providing CSI. In this sense, future research lines consider both the design of smart reconfigurable antennas and more sophisticated algorithms able to handle the coherence block requirements.

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