A Mixed-Integer Programming Approach to Interference Exploitation for Massive-MIMO

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***Abstract*—A novel low-complexity transmission scheme for Massive Multiuser Multi-Input Multi-Output (M-MU-MIMO) is proposed, where Transmit Antenna Selection (TAS) and beamforming are jointly performed to exploit multiuser interference. Two separate solutions to the deriving optimization problem are proposed: a mixed-integer programming approach that can optimally solve the TAS-beamforming problem and a heuristic convex approach, based on the assumption of matched filtering beamforming. Numerical results prove that the proposed multiuser interference exploiting approaches are able to greatly outperform previous state-of-the-art schemes, where TAS and beamforming are disjointedly solved.**

***Index Terms*—Massive MIMO, Multiuser MIMO, Antenna selection, Convex Beamforming, Interference optimization**

I. Introduction

Massive MIMO (M-MIMO) has experienced a growing interest from the research community [1] because of its promising benefits. More specifically, MIMO systems with very large arrays (VLAs) proved to be able to achieve extremely high throughputs and to be affected by very favorable propagation effects that lead linear precoding techniques to be asymptotically optimal [2]. At the same time, M-MIMO concepts can be directly applied to Multi-User (MU) scenarios, where the base station (BS) employs a VLA to perform secure, robust and energy-efficient communications with increased throughputs [3].

However, the practical implementation of M-MIMO comes with particularly challenging hardware and signal processing requirements. In fact, each radiating element needs a corresponding radio-frequency (RF) chain with its amplifier, analog-to-digital converter (ADC) and mixer. As a consequence, small active units [4] are expected to be preferred in order to respect cost and space constraints [5]–[7].

In the past, antenna selection [8] tackled the hardware complexity of multiple antenna systems by using only a subset of the available antennas. The selection can be performed according to several metrics, such as: received signal-to-noise ratio [9] and achievable capacity [10]. Recent works focused on antenna selection algorithms precisely designed for M-MIMO systems, since a direct application to M-MIMO is nearly prohibitive [11]–[13]. More specifically, [14] and [15] proposed the use of convex optimization for M-MIMO AS systems, for a massively distributed antenna system and for channel capacity optimization, respectively.

Recently, concepts of constructive multi-user interference (MUI) have been applied to M-MU-MIMO, showing that the benefits of interference exploitation can be extended to large-scale systems. In fact, previous works [16]–[21] proved that downlink MUI can be manipulated in order to increase received power at the user side. Instead, the authors in [22] have extended interference-based symbol-level precoding concepts to MU-MIMO communications security applications via Directional Modulation [23].

Conventionally, transmit antenna selection (TAS) based schemes approach selection and downlink beamforming as two disjointed optimization problems, by first defining the antenna subset and then proceeding with linear or nonlinear precoding [8]–[12], [24]. In this paper, a novel transmission scheme is proposed, where both TAS and beamforming are jointly performed according to constructive interference exploitation. The deriving optimization is a Mixed-Integer Programming (MIP) problem and can be efficiently solved by commercial optimization solvers. In addition to the optimal MIP solution, a heuristic convex approach is proposed, where Matched Filter beamforming is assumed.

The proposed schemes are specifically designed for PSK modulation transmissions, however, recent works have shown that constructive MUI concepts can also be applied to Quadrature Amplitude Modulation (QAM). More specifically, the authors in [25], [26] have shown that the predictable interference at the BS can constructively superimpose with the desired signal at the receiver side for 16-QAM and [27] has provided constraints and metrics for constructive MUI in QAM and asymmetric phase-shift keying (APSK) modulations.

Notation: The following notation is used throughout the paper. Upper case boldfaced letters identify matrices (i.e. $\mathbf{X}$), lower case boldfaced letters are used for vectors (i.e. $\mathbf{x}$), vector subindices are used to identify the columns of a matrix (i.e. $\mathbf{x}_m$ is the m-th column of $\mathbf{X}$), $diag(\cdot)$ identifies the diagonal of a matrix, superscripts $(\cdot)^T$, $(\cdot)^H$ and $(\cdot)^*$ stand respectively for transpose, Hermitian transpose and complex conjugate.

II. System Model

Consider a downlink transmission in a multiuser scenario, where the BS employs an N-sized VLA to communicate with K single antenna users. The received signal $\mathbf{y}$ is a $\mathbb{C}^{K\times 1}$ vector, which collects the received complex symbols

$$y_k = \mathbf{h}_k^T\mathbf{x} + n_k = \sum_{n=1}^{N} h_{n,k}x_n + n_k, \quad (1)$$

where $\mathbf{h}_k$ is the $\mathbb{C}^{N\times 1}$ channel vector experienced by the k-th user, i.e. the channel matrix $\mathbf{H} = [\mathbf{h}_1, ..., \mathbf{h}_k, ..., \mathbf{h}_K]$, and n_k is the k-th component of the $\mathbb{C}^{K\times 1}$ zero mean additive white Gaussian noise vector $\mathbf{n}$, i.e. $\mathbf{n} \sim \mathcal{CN}(0, N_0)$ with N_0 being the noise variance. The entries of the channel matrix $\mathbf{H}$ represent the complex path gain between the n-th antenna at the BS and the k-th user and are modeled as independent Rayleigh fading, i.e. $h_{n,k}, \forall n \in \{1, ..., N\}, k \in \{1, ..., K\}$ is a zero mean independent and identically distributed complex Gaussian variable [28].

In a TAS-based system, the analytical definition of the k-th user received signal $\tilde{y}_k$ is modified as follows

$$\tilde{y}_k = \sum_{n=1}^{N} h_{n,k}\tilde{x}_n + n_k, \tag{2}$$

where $\tilde{x}_n$ represents the n-th element of the $\mathbb{C}^{N\times 1}$ transmitted signal $\tilde{\mathbf{x}}$ and whose value is null when its index corresponds to one of the deactivated antennas, i.e., $\tilde{x}_n = 0,\ \forall n \notin \mathcal{N}$ with $\mathcal{N}$ being the subset of transmitting antennas with cardinality $card(\mathcal{N}) = N_t$.

A. Benchmark Schemes

The proposed transmission schemes are compared to a cascade of existing state-of-the-art TAS and beamforming techniques. More specifically, two separate TAS algorithms are considered: a simple path gain-based selection [9], which has been thoroughly studied in the literature for its simplicity, and a capacity maximization technique [15], which can be performed through convex optimization and has shown to achieve near optimal performances. With regards to beamforming, both Signal-to-interference-and-noise ratio (SINR) balancing [29] and linear zero-forcing (ZF) are considered.

1) Path Gain (PG) TAS: Path gain selection at the transmitter is easily performed by identifying the antennas with higher path gains [9]. Accordingly, the subset of transmitting antennas $\mathcal{N}$ can be analytically identified as

$$\mathcal{N} = \arg\max_{N_t}\left\{\sum_{n=1}^{N}|h_{n,1}|^2, ..., \sum_{n=1}^{N}|h_{n,k}|^2, ..., \sum_{n=1}^{N}|h_{n,K}|^2\right\} \tag{3}$$

where $\max\limits_{N_t}$ identifies the N_t largest values of the argument.

2) Capacity (Cap) TAS: In [15], TAS is performed with the aim to maximize system sum-capacity under the assumption of dirty-paper coding (DPC) at the transmitter. System sum-capacity with DPC is defined as

$$\mathcal{E} = log_2\left[\det\left(\mathbf{I}_K + \rho\mathbf{H}^H\mathbf{\Delta}\mathbf{H}\right)\right], \tag{4}$$

where $\mathbf{I}_K$ is a K-dimensional identity matrix, ρ is the signal-to-noise ratio and $\mathbf{\Delta}$ is the $N \times N$ dimensional selection matrix. The selection matrix is diagonal with binary entries: null values, i.e., $\Delta_{n,n} = 0$, infer that the corresponding antennas are deactivated and unitary values, i.e., $\Delta_{n,n} = 1$, identify the antennas that have been selected for transmission. Accordingly, the optimization problem for sum-capacity maximization antenna selection can be defined as

$$\begin{aligned}\mathcal{P}_{Cap}:\quad &\max_{\mathbf{\Delta}}\quad log_2\left[\det\left(\mathbf{I}_K + \rho\mathbf{H}^H\mathbf{\Delta}\mathbf{H}\right)\right]\\ &\text{s. t.}\quad \textstyle\sum_{n=1}^{N}\Delta_{n,n} = N_t, \Delta_{n,n} \in \{0,1\},\end{aligned} \tag{5}$$

The binary constraints imposed over the diagonal of the selection matrix cause the optimization problem $\mathcal{P}_{Cap}$ to be clearly non-convex. Nevertheless, such constraints can be relaxed and the deriving optimization problem was proven to be near-optimal when compared to exhaustive search approaches [15]. Accordingly, the convex-relaxation of $\mathcal{P}_{Cap}$ can be defined as

$$\begin{aligned}\mathcal{P}'_{Cap}:\quad &\max_{\mathbf{\Delta}}\quad log_2\left[\det\left(\mathbf{I}_K + \rho\mathbf{H}^H\mathbf{\Delta}\mathbf{H}\right)\right]\\ &\text{s. t.}\quad \textstyle\sum_{n=1}^{N}\Delta_{n,n} = N_t, \Delta_{n,n} \in \{0,1\},\end{aligned} \tag{6}$$

which leads to the TAS subset definition

$$\mathcal{N} = \arg\max_{N_t}\{\Delta_{1,1}, ..., \Delta_{n,n}, ..., \Delta_{N,N}\}. \tag{7}$$

3) SINR Balancing Beamformer: After the TAS is performed and the subset of transmitting antennas is identified, classical approaches from the literature proceed in defining the corresponding beamformer. In [29], a beamforming technique based on the maximization of the minimum received SINR is presented, here called SINR balancing. In fact, k-th user SINR is a direct function of the beamforming vectors $\mathbf{p}_k,\ \forall k$, as

$$\gamma_k = \frac{\left|\mathbf{h}_k^T\mathbf{p}_k\right|^2}{\sum\limits_{j\neq k}\left|\mathbf{h}_k^T\mathbf{p}_j\right|^2 + N_0} = \frac{\left|\sum\limits_{n=1}^{N}h_{n,k}p_{n,k}\right|^2}{\sum\limits_{j\neq k}\left|\sum\limits_{n=1}^{N}h_{n,k}p_{n,j}\right|^2 + N_0}, \tag{8}$$

where $p_{n,k}$ represents the n-th element of the beamforming vector $\mathbf{p}_k$.

Accordingly, given a power constraint P_T and a SINR threshold Γ_k, the SINR balancing optimization problem can be defined as

$$\begin{aligned}\mathcal{P}_{SINR}:\quad &\max_{\mathbf{p}_k}\quad \Gamma_k\\ &\text{s. t.}\quad \gamma_k \geq \Gamma_k, \sum_{k=1}^{K}\|\mathbf{p}_k\|^2 \leq P_T,\end{aligned} \tag{9}$$

which is non-convex and requires an algorithmic approach in order to be solved [29].

After the weight-beamforming vectors of all the users have been identified, the transmitted signal is derived as $\mathbf{x} = \sum\limits_{k=1}^{K}\mathbf{p}_k u_k$, where u_k represents the constellation symbol corresponding to the k-th user.

III. Constructive Interference Antenna Selection and Beamforming

Multi-user interference (MUI) is generally considered as a damaging element for communications, with signal processing aiming to reduce its effects [29]–[31]. However, early works

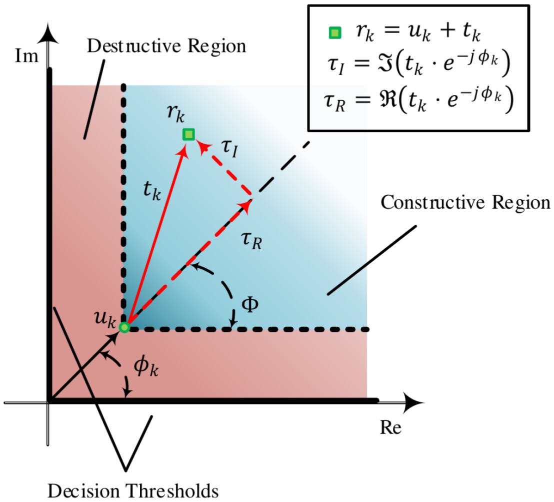

Fig. 1: Constructive and Destructive MUI regions for 4-PSK.

on interference exploitation for PSK signals have shown that both linear precoding [17] and TAS systems in [11] can benefit from MUI. In this section, the theoretical foundations of MUI-exploitation in PSK modulations are first discussed and then applied to the proposed optimization problems for a joint TAS-beamforming transmission.

A. MUI exploitation in Phase-Shift-Keying (PSK) signals

Multi-user interference symbols in PSK modulated transmissions can be deconstructed into the linear combination of a constructive or beneficial component and a destructive or detrimental component. The two components are distinguishable via simple geometrical concepts, described in details in MUI exploiting literature [17], [19], [20]. In general, constructive MUI causes the received signal in a noiseless scenario r_k to be positioned further away from the decision thresholds of the desired constellation symbol u_k, while destructive MUI leads r_k to lie closer to the decision thresholds. The MUI symbol for the k-th user can be explicitly defined in the following

$$t_k = r_k - u_k = \sum_{n=1}^{N} h_{n,k} x_n - u_k. \tag{10}$$

The condition for constructive MUI in a generalized M-ordered PSK modulation [19] is analytically expressed by the following inequality

$$\Re\left(t_k \cdot e^{-j\phi_k}\right) \tan\Phi - \left|\Im\left(t_k \cdot e^{-j\phi_k}\right)\right| \geq 0, \tag{11}$$

where $\Re(\cdot)$ and $\Im(\cdot)$ identify the extraction of the real and imaginary part of the complex argument, respectively, and Φ represents the central angle of the constellation sectors identified by the decision thresholds, whose value is a function of the constellation order M, i.e., $\Phi = \pi/M$. In (11), the MUI symbol t_k is phase-shifted according to the phase of the desired constellation symbol for the k-th user $u_k = e^{j\phi_k}$, as it allows to isolate the effects of interference over phase and amplitude of u_k, as shown in Fig.1.

At the same time, the inequality (11) provides a quantitative evaluation of how constructive or destructive MUI is for the k-th user. In fact, a negative (11) implies that r_k lies in the destructive region past the decision thresholds, the red area in Fig.1, while a positive (11) suggests that MUI is forcing r_k further away from the decision thresholds, the blue region in Fig.1.

B. Joint MIP Constructive Selection and Beamforming (MIP-CSB)

The condition for constructive MUI in (11) can be used to define the following optimization problem, where TAS and beamforming are jointly performed

$$\begin{aligned} \mathcal{P}_{CSB}: \quad & \max_{\mathbf{a},\tilde{\mathbf{x}}} \ \min_{k} \left\{\Re\left(t_k e^{-j\phi_k}\right)\tan\Phi - \left|\Im\left(t_k e^{-j\phi_k}\right)\right|\right\} \\ & \text{s. t.} \quad t_k = \sum_{n=1}^{N} h_{n,k}\tilde{x}_n - u_k, \\ & \qquad |\tilde{x}_n| \leq a_n, \\ & \qquad \sum_{n=1}^{N} a_n = N_t, a_n \in \{0,1\}. \end{aligned} \tag{12}$$

Here, $\mathbf{a}$ represents the selection vector, whose entries are unitary when the corresponding antenna-index is connected to an RF chain for transmission and null when the corresponding antenna-index is deactivated. Binary constraints in (12) cause the optimization problem to be clearly non-convex, however it can be efficiently solved by commercial optimization tools such as MoSek. As we can see, $\mathcal{P}_{CSB}$ is designed in order to jointly perform the selection (i.e., by identifying the vector of transmitting antennas $\mathbf{a}$) and design the transmitted signal $\tilde{\mathbf{x}}$. The joint optimization allows us to fully exploit the beneficial components of MUI, achieving significant transmission benefits and a particularly interesting trade-off between system complexity and performances.

C. Joint Matched Filtering Constructive Selection (MFCS)

In addition to the optimal solution achieved by MIP-CSB for joint TAS-beamforming, a heuristic approach to $\mathcal{P}_{CSB}$, called Matched Filtering Constructive Selection (MFCS), is proposed. In MFCS, the solution to TAS-beamforming is achieved by solving the convex optimization problem that derives under the key assumption that simple matched filtering (MF) beamforming is performed at the transmitter side. Accordingly, the optimization problem for TAS and beamforming can be represented as follows

$$\begin{aligned} \mathcal{P}_{MFCS}: \quad & \max_{\tilde{\mathbf{\Delta}}} \ \min_{k} \left\{\Re\left(c_k e^{-j\phi_k}\right)\tan\Phi - \left|\Im\left(c_k e^{-j\phi_k}\right)\right|\right\} \\ & \text{s. t.} \quad c_k = u_k \sum_{n=1}^{N} h_{n,k} \sum_{i=1}^{N} h_{i,k}^{*}\tilde{\Delta}_{i,n} - u_k, \\ & \qquad \sum_{n=1}^{N} \tilde{\Delta}_{n,n} = N_t, \tilde{\Delta}_{n,n} \in [0,1], \\ & \qquad \tilde{\Delta}_{i,j} = 0, \forall i \neq j. \end{aligned} \tag{13}$$

The optimization problem (13) is convex and can be efficiently solved by standard convex optimization tools. More specifically (13) can be represented as a standard Second-Order Cone Programming (SOCP) problem [32] with a concave objective function [19]. In fact, the cost function of $\mathcal{P}_{_{MFCS}}$ can be deconstructed into the combination of two functions: a linear function $\Re\left(c_k e^{-j\phi_k}\right)$ and a concave function $-\left|\Im\left(c_k e^{-j\phi_k}\right)\right|$, since the extraction of the imaginary and real part of a linear function preserves its linearity [19].

It is important to remark that (13) leads to a selection matrix $\tilde{\mathbf{\Delta}}$ whose entries are not binary, since the binary constraints of $\mathcal{P}_{_{CSB}}$ have been relaxed. Accordingly, before computing the selection vector $\mathbf{a} = diag(\mathbf{\Delta})$, it is necessary to identify the binary counterpart of $\tilde{\mathbf{\Delta}}$ as

$$\Delta_{n,n} = \begin{cases} 1, & \forall \quad n \in \left\{ \arg\max\limits_{N_t} \left[diag(\tilde{\mathbf{\Delta}}) \right] \right\} \\ 0 & \forall \quad n \notin \left\{ \arg\max\limits_{N_t} \left[diag(\tilde{\mathbf{\Delta}}) \right] \right\}, \end{cases} \tag{14}$$

where $\arg\max\limits_{N_t}[\cdot]$ is used to extract the indices of the N_t largest elements of the argument.

After the selection vector $\mathbf{a} = diag(\mathbf{\Delta})$ has been identified, the transmitted signal $\tilde{\mathbf{x}}$ can be computed as

$$\tilde{x}_n = \begin{cases} 1/\xi_n \sum\limits_{k=1}^{K} h_{n,k} u_k, & \text{if} \quad a_n = 1, \\ 0 & \text{if} \quad a_n = 0, \end{cases} \tag{15}$$

where ξ_n is a scaling factor, which guarantees a unitary transmitted power $\sum\limits_{n=1}^{N} |\tilde{x}_n|^2 = 1$.

IV. Results

This section analyzes the proposed transmission schemes in terms of Symbol-Error-Rate (SER) at the receiver side, power efficiency and running times. Results are presented for 4-PSK transmissions; however, the proposed schemes can be directly applied to any PSK modulation order. Figure legends are characterized by the following notation: MIP-CSB identifies the MUI-exploiting scheme based the MIP solution, MFCS is used to classify the heuristic solution to the TAS-beamforming problem based on the MF assumption at the BS and Cap-SINR represents the approach from the literature where TAS and beamforming are disjointly performed according to Cap selection [15] and SINR balancing [29], respectively. In addition, the proposed schemes are compared to two low-complexity approaches from the literature: PG-ZF, where path gain selection (PG) and zero forcing (ZF) linear beamforming are performed in cascade, and CIM-HY from [11], where TAS is performed in order to maximize constructive interference (CIM) and the transmitted signal is derived through hybrid (HY) linear precoding. In the proposed simulations, a single-cell downlink M-MIMO scenario with $K = 5$ single-antenna mobile users is considered, where the BS possesses perfect channel-state information (CSI) and employs a VLA of $N = 100$ antennas.

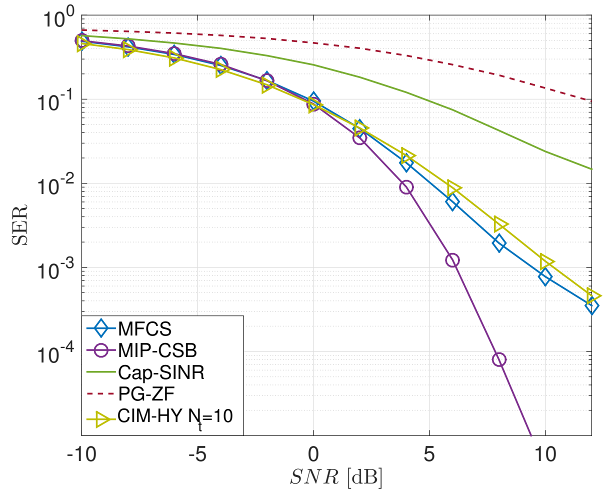

Fig. 2: 4-PSK Symbol Error Rate when $K = 5$, $N = 100$ and $N_t = 5$ with perfect CSI.

A. Symbol Error Rate (SER)

Fig.2 shows that the proposed schemes are able to rapidly achieve very low values of SER, while outperforming all the benchmark techniques, including the Cap-SINR scheme, characterized by a combination of Cap TAS [15] and SINR balancing beamforming [29]. While the MFCS scheme is based on a convex relaxation, it is still able to strongly outperform both the Cap-SINR and the PG-ZF techniques, hence proving to be a valuable approach for low-complexity and low-power scenarios. Additionally, it can be seen that the proposed schemes outperform the previous TAS scheme based on the exploitation of constructive MUI. The performance improvements are supported by two main factors: first, the optimality of the solutions achieved by the proposed MIP-CSB and MFCS schemes is guaranteed, and second, the proposed schemes perform a joint TAS-beamforming, which allows to achieve the maximum benefits from both problems, solved instead in a separate manner by CIM-HY.

B. Power Efficiency

The benefits and trade-offs introduced by the proposed schemes are further described by evaluating the power efficiency over throughput η_T, defined as the ratio

$$\eta_T = \frac{T}{P_{amp} + N_t \cdot P_{RF}} = \frac{(1 - BLER) \cdot m \cdot K}{P_{amp} + N_t \cdot P_{RF}}, \tag{16}$$

where $BLER$ is the block error rate, $m = \log_2(M)$ is the bit information per symbol, K is the number of users in the chosen scenario, P_{amp} $[W]$ represents the power consumption of the amplifier and P_{RF} $[W]$ identifies the power by one RF chain, which is characterized by digital-analog converter, mixer and filter. When computing (16), realistic power values from practical systems [33] are considered, where $P_{amp} = P_t/\nu$ is defined as the power required by an amplifier with $\nu = 0.35$ efficiency and transmitted power $P_t = 30dBm$ and

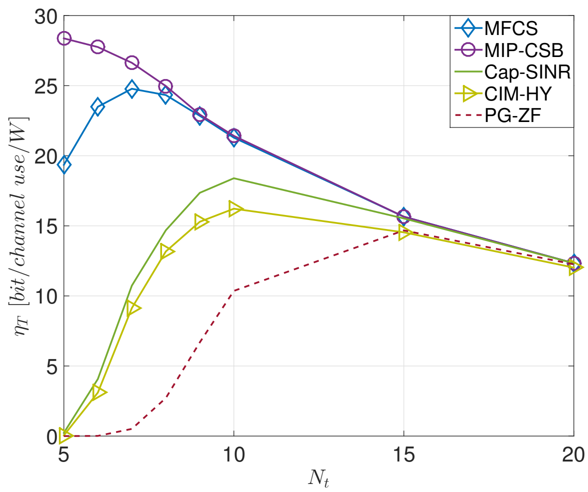

Fig. 3: 4-PSK Power Efficiency over Throughput when $K = 5$, $N = 100$ with perfect CSI and $SNR = 5dB$.

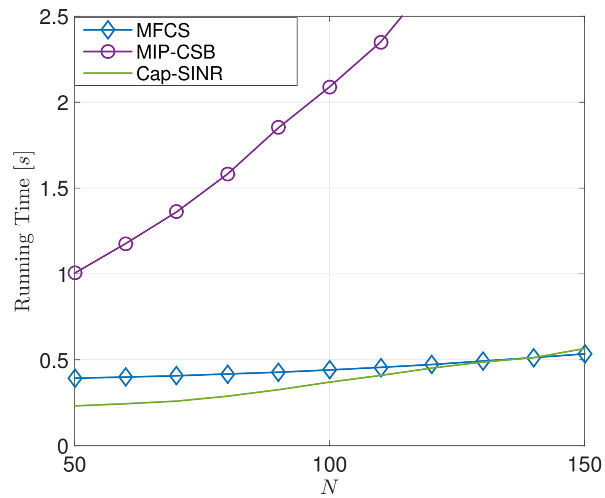

Fig. 4: Frame Running time when $K = 5$, $N_t = 5$ and $T_{DL} = 4$.

$P_{RF} = 65.9mW$. Power efficiency performances for the 4-PSK case are presented in Fig.3 as a function of the subset size N_t when considering an $SNR = 5dB$.

Schemes from the literature are all outperformed by both MIP-CSB and MFCS for all the spectrum of subset size N_t values. At the same time, it is interesting to notice that the proposed MFCS scheme is characterized by maximum value of η_T when $N_t = 7$, hence showing that 2 additional antennas in the subset size N_t could further improve the SER performances shown in Fig.2 with a beneficial trade-off between hardware complexity and power consumptions.

C. Complexity Analysis

The computational burdens of the proposed schemes are evaluated in terms of running times with increasing antenna array sizes at the BS. For the sake of a fair comparison, running times are computed within a coherence time, during which the channel state information $h_{n,k}, \forall n \in \{1,..,N\}, k \in \{1,..,K\}$ is constant. This is due to the fact that the proposed schemes require to evaluate TAS-beamforming at a symbol-by-symbol rate, while the Cap TAS scheme needs to be performed on a coherence time basis. In the proposed study, a Time Division Duplexing (TDD) scenario [34] is considered, where coherence time T_{cohe} represents the total number of data-symbols that can be transmitted while considering $\mathbf{H}$ constant. The TDD assumption is common on M-MIMO, as it allows to exploit the channel reciprocity property, which causes the number of slots used for CSI acquisition T_{CSI} to be directly proportional to the number of users K. As it follows, the number of slots allocated to data transmission and reception can be computed as

$$T_{data} = T_{cohe} - \mu K, \tag{17}$$

where $\mu \geq 1$ is direct proportionality parameter that defines the number of pilot symbols used for CSI per user. Accordingly, the final number of data symbols to be allocated for downlink within a single coherence time is defined as

$$T_{DL} = \eta_{DL} \left(T_{cohe} - T_{CSI} \right), \tag{18}$$

where $\eta_{DL} \in [0,1]$ is the proportionality factor used to identify the downlink portion of T_{data}.

Running times are presented in Fig.4 for a realistic scenario where $T_{cohe} = 10$ symbols, $T_{CSI} = 5$ symbols and $T_{DL} = 4$ symbols, in line with the work by [34]. It can be seen that the proposed schemes are overall affected by longer computational times. Such behavior is caused by the fact that the proposed schemes require a symbol-by-symbol update, in contrast with conventional TAS schemes from the literature. Nonetheless, the proposed MFCS scheme is characterized by running times that are comparable to the ones of Cap-SINR. This is particularly important, as it proves that such approach is able to achieve interesting performances in terms of SER and power efficiency with non-significant additional computational costs, when compared to Cap-SINR. The proposed MFCS scheme is almost unaffected by the increase in array sizes, while Cap-SINR is instead characterized by increasing running times for larger arrays. Accordingly, it can be inferred that the proposed MFCS scheme is expected to be characterized by similar or lower complexity for larger systems, when compared to existing state-of-the-art TAS schemes, such as Cap-SINR. On the other hand, the MIP-CSB curve shows that the MIP approach is characterized by higher computational times, which are caused by its trellis search-based solution.

However, Cap-SINR based schemes require further operations at the receiver in order to recover the data. In fact, the BS is required to feed-forward the mobile users with $\mathbf{h}_k^T \mathbf{p}_k, \forall k$ in order to equalize the received signal. On the contrary, such feedback is not required by the proposed approaches, where the computational complexity fully resides at the BS and additional operations at the receiver, such as estimation and equalization, are not necessary.

V. Conclusions

In this paper, transmit antenna selection and beamforming based on constructive MUI are jointly performed in order to improve power efficiency performances of future M-MU-MIMO systems. The presented numerical studies show that constructive MUI at the receiver side can be efficiently optimized and exploited by simultaneously identifying a subset of transmitting antennas and the precoded signal. The proposed schemes are evaluated in terms of symbol error rate, power efficiency and computational complexity and their performances are compared with state-of-the-art schemes from the literature, showing the significantly positive trade-offs introduced by the proposed schemes.

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