Beam Allocation based on Spatial Compatibility for Hybrid Beamforming C-RAN Networks

Lászlon R. Costa[*†], Yuri C. B. Silva[†], F. Rafael M. Lima[†], and Anja Klein[*]
[*]Communications Engineering Lab, Technische Universität Darmstadt, Merckstrasse 25, 64283 Darmstadt, Germany
[†]Wireless Telecom Research Group (GTEL), Federal University of Ceará, Fortaleza, Brazil
{l.rodrigues, a.klein}@nt.tu-darmstadt.de, {laszlon, yuri, rafaelm}@gtel.ufc.br

***Abstract*—In this work, we consider a beam allocation problem in hybrid beamforming (HBF) cloud radio access networks (C-RANs) to maximize the sum rate. The problem considers a codebook-based analog beamforming performed at the remote radio head (RRH) and digital beamforming at the baseband unit (BBU). Differently from previous works, we assume that a given user equipment (UE) can be served by multiple beams, without being specifically associated to a given RRH. Due to the relation between the analog beam allocation and the digital beamforming, we consider a new metric based on channel correlation and channel attenuation for the analog beamforming solution. The metric measures the spatial compatibility in order to improve the spectral efficiency before the digital beamforming solution. In order to evaluate the proposed metric we present a low-complexity greedy algorithm, which is shown to provide a reasonable performance/complexity trade-off.**

***Index Terms*—hybrid beamforming, Cloud-RAN, Beam allocation**

I. Introduction

In order to attend to the increasing demands for high data rates, low latency and massive connectivity in upcoming 5G wireless networks [1], some key technologies will be required. Among these we can mention millimeter wave transmission [2], massive multiple-input multiple-output (MIMO) [3] and cloud radio access network (C-RAN) [4]. The use of mmWave massive MIMO leads to increased bandwidth and high data rates, by efficiently using the spatial and frequency domains. In order to reduce the complexity and energy consumption of massive MIMO, hybrid beamforming (HBF) techniques [5]–[7] split the precoding design into two parts, with the analog processing being carried out at the radio frequency (RF) stage and the digital part at the baseband. The deployment of C-RAN, on the other hand, allows the signal processing functions to be split among the baseband unit (BBU) and the remote radio heads (RRHs).

By using C-RAN solutions, it is possible to improve the effectiveness of interference management and cooperative techniques. The implementation of mmWave massive MIMO in a C-RAN context may lead to benefits in terms of cost and capacity, as shown in [8]. In such scenario, the analog beamforming can be implemented at the RRHs and the digital part at the BBU. Considering a fast enough backhaul among the BBU and RRHs, it becomes possible to implement joint transmission techniques [9] for C-RAN scenarios, as considered in [10], [11].

In this paper, we consider the problem of beam allocation in HBF C-RAN networks. The problem is formulated for the joint precoding scenario, assuming a codebook-based analog beamforming at the RRHs and digital beamforming at the BBU. Differently from previous works, such as [10]–[12], we assume that a given user equipment (UE) can be served by multiple beams, without being specifically associated to a given RRH. The formulated problem is non-convex and composed of integer and continuous variables, which makes the optimal solution impractical for realistic scenarios.

Therefore, we propose to separate the analog beamforming allocation from the digital precoder solutions as two steps performed sequentially. The first step is the analog beam allocation and the second step is the digital precoder calculated considering the analog beamforming solution. In order to improve the system capacity, we propose an analog beam allocation method based on a spatial compatibility metric, which considers the channel correlation and attenuation. The solution considering this metric will create an equivalent channel with low spatial correlation, therefore, the interfering signals can be easily isolated by the digital beamforming. It is worth to mention that the components of the proposed metric, i.e., channel correlation and attenuation, can be acquired from channel statistics, thus avoiding the need for instantaneous channel state information (CSI). In order to evaluate our beam allocation metric, we propose a low-complexity greedy algorithm.

The paper is organized as follows. In section II, the considered system model is presented. The considered beam allocation problem and the proposed suboptimal solutions are detailed in section III. Next, simulation results are presented and discussed in section IV, while conclusions and next steps are provided in section V.

II. System Model

We consider M base stations (BSs), each one corresponding to an RRH equipped with N antennas and B RF chains, simultaneously serving K single-antenna UEs. Each BS has a fully connected hybrid beamforming architecture, which means that all antennas from one BS are connected to all B RF chains. Therefore, each BS is capable to design B analog beams using N antennas, where $B < N$. For simplicity, let us consider $MB \geq K$, which means that it is possible to point at least one analog beam to each UE considering the overall system. The BSs are connected to a cloud processing unit, which performs all baseband processing.

Let us define $\mathbf{h}_{m,k} \in \mathbb{C}^{1\times N}$ as the downlink channel between BS m and UE k. The concatenation of channel vectors between a UE k and all BSs is represented as

$$\mathbf{h}_k = [\mathbf{h}_{1,k}, \cdots, \mathbf{h}_{M,k}] \in \mathbb{C}^{1\times MN}. \tag{1}$$

The analog beams designed by each BS m are composed of phase shifts applied at each transmit antenna, which is represented by the vector $\mathbf{f}^A_{m,b} \in \mathbb{C}^{N\times 1}$, b is the index of the RF chain associated with the analog beam. Each element of $\mathbf{f}^A_{m,b}$ is given by $\frac{1}{\sqrt{N}}e^{j\theta}$, where θ is a quantized angle. The complete analog beamforming matrix considering all beams of BS m is defined as

$$\mathbf{F}^A_m = \left[\mathbf{f}^A_{m,1}, \cdots, \mathbf{f}^A_{m,B}\right] \in \mathbb{C}^{N\times B}, \tag{2}$$

where each column of matrix $\mathbf{F}^A_m$ represents an analog beam. The block diagonal matrix containing the analog beamforming matrices from all cooperative BSs is defined as

$$\mathbf{F}^A = \text{blkdiag}\left(\mathbf{F}^A_1, \cdots, \mathbf{F}^A_M\right) \in \mathbb{C}^{MN\times MB}. \tag{3}$$

Let $\mathbf{f}^D_{m,k} \in \mathbb{C}^{B\times 1}$ be the digital beamforming designed by the cloud processing unit to UE k in BS m. The overall digital beamformer to UE k considering the RF chains from all BSs can be expressed as

$$\mathbf{f}^D_k = \left[(\mathbf{f}^D_{1,k})^H, \cdots, (\mathbf{f}^D_{M,k})^H\right]^H \in \mathbb{C}^{MB\times 1}. \tag{4}$$

Therefore, the SINR of a UE k can be calculated by

$$\gamma_k = \frac{|\mathbf{h}_k \mathbf{F}^A \mathbf{f}^D_k|^2}{\sum_{\substack{i=1\\ i\neq k}}^{K} |\mathbf{h}_k \mathbf{F}^A \mathbf{f}^D_i|^2 + \sigma^2}, \tag{5}$$

where σ^2 is the received additive white Gaussian noise (AWGN) power and the transmit symbols are assumed to have unit variance.

III. Beam Allocation Problem

In this work, we consider that each analog beam at each BS is taken from a predefined codebook composed of C codewords. Each codeword corresponds to a set of phase shifts at the transmit antennas, representing an analog beam. The codebook is denoted by matrix $\mathbf{C} \in \mathbb{C}^{N\times C}$, where each column is an analog beam. For simplicity, let us consider the same codebook for all BSs and let the matrix $\overline{\mathbf{F}}^A$ be a block diagonal matrix containing the codebooks from all BSs, defined as

$$\overline{\mathbf{F}}^A = \mathbf{I}_M \otimes \mathbf{C} \in \mathbb{C}^{MN\times MC}, \tag{6}$$

where $\mathbf{I}_M$ is an identity matrix of dimension M and $\otimes$ is the Kronecker product.

In order to maximize the capacity of the network, the cloud processing unit has to determine the analog beams of each BS, as well as design a digital beam for each UE. Further, let $\mathbf{U}_{m,k} \in \mathbb{B}^{C\times B}$ be the binary matrix representing beam allocations of BS m to UE k. The columns of $\mathbf{U}_{m,k}$ correspond to the different RF chains, while its rows represent codeword indices from the codebook, i.e., a given column from matrix $\mathbf{C}$. When the codeword c is associated with the RF chain b from a BS m in order to point a beam towards UE k, the matrix element $[\mathbf{U}_{m,k}]_{c,b}$ is set to 1, otherwise it is zero.[1]

The block diagonal matrix containing all beams allocated to UE k is given by

$$\mathbf{U}_k = \text{blkdiag}\left(\mathbf{U}_{1,k}, \cdots, \mathbf{U}_{M,k}\right) \in \mathbb{B}^{MC\times MB}, \tag{7}$$

and the total beams allocated in the system considering all BSs is given by matrix

$$\mathbf{U} = \sum_{k=1}^{K} \mathbf{U}_k, \in \mathbb{B}^{MC\times MB}. \tag{8}$$

The optimization problem of joint analog beam allocation and digital beamforming design to maximize the sum capacity of the network can be formulated as:

$$\max_{\mathbf{U}_k, \mathbf{f}^D_k} \sum_{k=1}^{K} \log_2 \left(1 + \frac{|\mathbf{h}_k \overline{\mathbf{F}}^A \mathbf{U} \mathbf{f}^D_k|^2}{\sum_{\substack{i=1\\ i\neq k}}^{K} |\mathbf{h}_k \overline{\mathbf{F}}^A \mathbf{U} \mathbf{f}^D_i|^2 + \sigma^2}\right) \tag{9a}$$

$$\text{s.t} \sum_{k=1}^{K} \left\|\mathbf{F}^A_m \mathbf{f}^D_{m,k}\right\|^2 \leq P_{\text{max}} \quad \forall m, \tag{9b}$$

$$\sum_{k=1}^{K} \mathbf{1}^T_{MC} \mathbf{U}_k = \mathbf{1}^T_{MB}, \tag{9c}$$

$$\mathbf{1}^T_{MC} \mathbf{U}_k \mathbf{1}_{MB} \geq 1 \quad \forall k. \tag{9d}$$

where $\mathbf{1}_X$ represents a column vector of ones containing X elements. The objective function in (9a) maximizes the sum capacity of the system, considering the capacity achieved by each UE. Constraint (9b) is the power constraint of each BS, whose maximum transmit power is given by P_{max}. Considering that it might be impractical to have one RF chain generate more than one analog beam, constraint (9c) guarantees that one codeword will be associated to each RF chain and that the system will operate using all MB RF chains. Finally, constraint (9d) guarantees that at least one analog beam will be pointed to each UE in the system.

Note that problem (9) has a nonconvex objective function and that there exists a coupling among the binary variables $\mathbf{U}_k$ and continuous variables $\mathbf{f}^D_k$, which leads to an intractable problem. For these reasons, we consider the solution to be decoupled into three separate lower-complexity steps: analog beam allocation; digital precoder design and power allocation. The focus of this work is on the analog beam allocation step. In this step, we evaluate low-complexity solutions considering different criteria to estimate the final network performance. The first solution is based on capacity estimates without considering the digital precoder. The second one is a solution that takes into account the correlation among interfering effective channels created by the analog beam selection.

[1]The operator $[\mathbf{A}]_{c,b}$ represents the element in row c and column b of matrix $\mathbf{A}$.

A. Low-Complexity Sub-Optimal Solutions

For the analog beam allocation solution, let us consider the equivalent channel between a BS m and a UE k, which is created by the analog beam matrix. The equivalent channel can be defined as $\mathbf{h}_{m,k}\mathbf{F}_m^A \in \mathbb{C}^{1\times B}$. Note that, considering the codebook $\mathbf{C}$ and the binary beam allocation matrix $\mathbf{U}_{m,k}$, we can define the analog beams of BS m as

$$\mathbf{F}_m^A = \mathbf{C}\sum_{k=1}^{K}\mathbf{U}_{m,k}. \tag{10}$$

Based on the equivalent channel, we evaluate two low-complexity sub-optimal solutions for the analog beam allocation step. The first approach is an adaptation of the solution proposed in [11], while the second one corresponds to our correlation-based proposal. In order to describe the algorithms, let us define some auxiliary sets and variables. Let $\mathcal{C}_m$ be the set of codewords available to be allocated at BS m. Further, let $\mathcal{M}$ and $\mathcal{K}$ be the sets of BSs and UEs available to the beam allocation, respectively. Finally, let the vector $\mathbf{c}_i \in \mathbb{C}^{N\times 1}$ define the i-th column of matrix $\mathbf{C}$.

The algorithms follow the structured pseudo-code presented in Algorithm 1. The main difference between them is the criterion to select the tuple codeword-BS-UE to be assigned.

Algorithm 1 Greedy Algorithms

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Define 𝓜 and 𝓚 as the set of all BSs and UEs in the system, respectively.
Define 𝓒_m as the set of all codewords for all BSs.
Select the tuple codeword-BS-UE (c*, m*, k*) which has the largest expected received power |h_{m*,k*} c_{c*}|^2
Set [U_{m*,k*}]_{c*,b*} = 1, where b* is the index of an RF chain at BS m without assigned codeword.
Update 𝓚 = 𝓚 \ {k*} and 𝓒_m = 𝓒_m \ {c*}.
In the specific case that M = 1, do 𝓜 = 𝓜 \ {m*}.
while 𝓚 ≠ ∅ do
    Select the tuple codeword-BS-UE (c*, m*, k*) which maximizes metric (11) or minimizes (13).
    Set [U_{m*,k*}]_{c*,b*} = 1
    𝓚 = 𝓚 \ {k*} and 𝓒_m = 𝓒_m \ {c*}.
    If all RF chains in BS m* have been allocated, do 𝓜 = 𝓜 \ {m*}.
end while
If 𝓚 = ∅ and the constraint (9c) is not fulfilled, update 𝓚 to incllude all UEs of the system.
while There are RF chains in any BS without codeword do
    Select the tuple codeword-BS-UE (c*, m*, k*) which maximizes metric (11) or minimizes (13).
    Set [U_{m*,k*}]_{c*,b*} = 1
    𝓒_m = 𝓒_m \ {c*}.
    If all RF chains in BS m* have been allocated, do 𝓜 = 𝓜 \ {m*}.
end while
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The first algorithm is the Greedy-Estimated Capacity, which is an adaption of the algorithm proposed in [11]. This algorithm considers the expected signal power after the beam allocation in order to predict system capacity. The Greedy-Estimated Capacity expression is defined as

$$f_{\text{CAP}}\left(c^*, m^*, k^*\right) = \sum_{k=1}^{K}\log_2\left(1+\frac{\|\mathbf{h}_k\overline{\mathbf{F}}^A\mathbf{U}_k\|^2}{\sum_{\substack{j=1\\ j\neq k}}^{K}\|\mathbf{h}_k\overline{\mathbf{F}}^A\mathbf{U}_j\|^2+\sigma^2}\right), \tag{11}$$

where the relation between intended signal power and interfering signal power is made considering the equivalent channel without digital precoder. The main goal of the greedy algorithm is to select tuples codeword-BS-UE which maximize metric (11).

The second criterion is the Greedy-Correlation, which considers the correlation between equivalent channels after the beam allocation and the channel atenuation on the tuple codeword-BS-UE (c^*, m^*, k^*). Hence, we can calculate the complete downlink equivalent channel to one given UE considering the beam allocation of all BSs by

$$\overline{\mathbf{h}}_k = \mathbf{h}_k\overline{\mathbf{F}}^A\mathbf{U}. \tag{12}$$

The Greedy-Correlation criterion is defined by

$$f_{\text{COR}}\left(c^*, m^*, k^*\right) = \|\mathbf{h}_{k^*,m^*}\mathbf{c}_{c^*}\|^{-2}\sum_{k}^{K}\sum_{\substack{j=1\\ j\neq k}}^{K}\frac{|\overline{\mathbf{h}}_k\overline{\mathbf{h}}_j^H|}{\|\overline{\mathbf{h}}_k\|\|\overline{\mathbf{h}}_j\|}, \tag{13}$$

where the term $\|\mathbf{h}_{k^*,m^*}\mathbf{c}_{c^*}\|^{-2}$ is the channel attenuation and the term $\frac{|\overline{\mathbf{h}}_k\overline{\mathbf{h}}_j^H|}{\|\overline{\mathbf{h}}_k\|\|\overline{\mathbf{h}}_j\|}$ is the channel correlation. The vector $\mathbf{c}_{c^*}$ is the column vector c^* from the codebook matrix $\mathbf{C}$.

Note that, in order to efficiently isolate the signals in the space domain by digital beamforming, an analog beam assignment that reduces the correlation among the downlink equivalent channels is desired. In addition, analog beams which have good signal reception by UEs are also desired. Therefore, the greedy-correlation method will allocate analog beams with low correlation and good channel quality. It is worth to mention that the correlation and channel antennuation could be acquired by evaluating the statistics of the channel, therefore, the metric could be calculated without considering the instantaneous CSI.

In order to evaluate the performance of the greedy analog beam allocation, we consider a classical regularized zero-forcing (RZF) digital precoder that acts on the equivalent channel created by analog beamforming. In this approach, we calculate the digital precoders for each UE k considering the equivalent channels after the analog beamforming solution. The RZF for the HBF scenario is given by

$$\mathbf{F}^D = \zeta\left(\left(\mathbf{F}^A\right)^H\mathbf{H}^H\mathbf{H}\mathbf{F}^A+\alpha MB\mathbf{I}_{MB}\right)^{-1}\left(\mathbf{F}^A\right)^H\mathbf{H}^H, \tag{14}$$

where $\mathbf{H}$ is the concatenation of all UE channels $\mathbf{H} = \left[\mathbf{h}_1^H, \cdots, \mathbf{h}_K^H\right]^H$, ζ is a normalization parameter to fulfill the

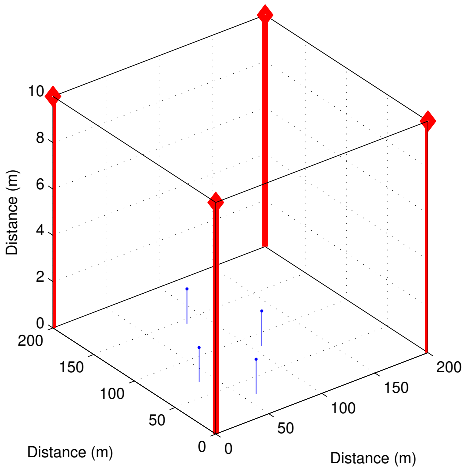

Fig. 1. Simulated environment with 4 UEs in the system.

power constraint at each BS and α is the regularization parameter which controls the interference. In this work, we consider the corresponding $\alpha = (K\sigma^2)/(P_{\max} MB)$, which maximizes the signal-to-interference-plus-noise ratio (SINR) in a single-cell scenario [13], [14]. Each column of $\mathbf{F}^D$ results in a single precoder vector for each UE, $\mathbf{F}^D = [\mathbf{f}_1^D, \cdots \mathbf{f}_K^D] \in \mathbb{C}^{MB\times K}$.

For the power allocation step, we have to consider the per-BS power constraints. This means that if we normalize each BS digital precoder, the total combined digital precoder will no longer cancel the interference [9], [15]. Therefore, we apply a sub-optimal power allocation according to

$$\zeta = \left\{ \min_{m=1,\cdots,M} \sqrt{\frac{P_{\max}}{\|\mathbf{F}_m^A \mathbf{F}_m^D\|_F^2}} \right\}, \tag{15}$$

where $\mathbf{F}_m^D = [\mathbf{f}_{m,1}^D, \cdots, \mathbf{f}_{m,K}^D] \in \mathbb{C}^{B\times K}$.

This power allocation solution guarantees that all BSs will fulfill the power transmit budget without modifying the zero forcing effect. However, this solution will typically lead to a situation where only one BS transmits with total power, which is not optimal in terms of system capacity.

IV. Numerical Results

In order to evaluate the analog beamforming solutions presented in Section III-A, simulations have been performed. The simulated system corresponds to an outdoor micro-cell environment deployed in a system with 100 MHz bandwidth and line-of-sight (LOS) at all links. We have simulated a scenario considering a square environment with 4 BSs at each vertex and with the antenna array pointing towards the center of the square, as illustrated in Figure 1. We have considered 24 dBm of transmission power budget, which is equally divided among 124 frequency resources. The proposed algorithm has been applied at each frequency resource, considering the amount of power reserved to it. The main parameters used in the simulations are specified in Table I.

TABLE I
General simulation parameters

Num. of BSs	4
Num. of BS antennas	64
Num. of UE antennas	1
BS height	10 m
UE height	1.5 m
BS antenna downtilt	12 degrees
Distance between neighbor BSs	200 meters
BS antenna array	uniform rectangular array (URA) 8x8
UE antenna array	Omni-directional
Maximum transmit power	24 dBm
Num. frequency resources	124
Noise density	-174 dBm/Hz
Monte Carlo samples	500
Digital Precoder	RZF
Codebook	discrete fourier transform (DFT)
Channel Model	QuaDRiGA [16]
Scenario	3GPP 3D Urban Micro-Cell LOS

In our simulations we consider two reference cases: Full digital and Random. The full digital approach is the method where we consider that each BS has the number of RF chains equal to the number of antennas. In this case, only the digital precoder with all antennas is considered. As for the random approach, it considers a random allocation of codewords and users, while satisfying the constraints in problem (9). The position of each UE is generated randomly considering a uniform area distribution.

Figure 2 shows the sum data rate at one frequency resource of the system for the simulated algorithms, considering 2 RF chains per BS and that the number of UEs increases from 2 to 8.

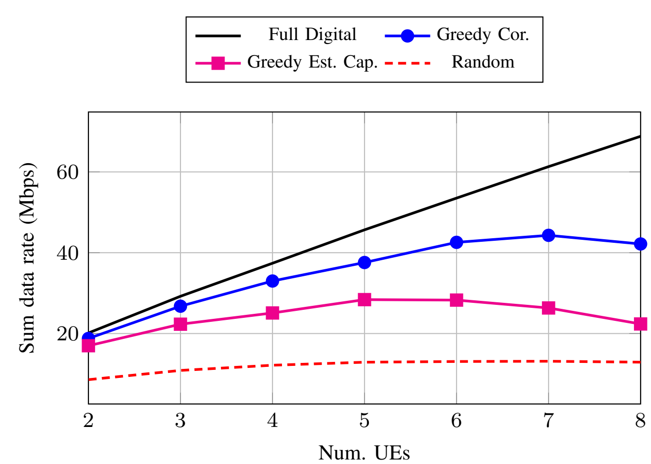

Fig. 2. Sum data rate in one frequency resource versus number of UEs considering 2 RF chains per BS.

It can be seen that the full digital and random approaches present the upper and lower performances, respectively, as expected. The sum data rate increases with the number of UEs up to a certain point, after which it starts to decrease for all HBF algorithms. The reason is that the maximum number of possible analog beams in the system is limited to 8 and thus the algorithm does not have enough degrees of freedom to find a

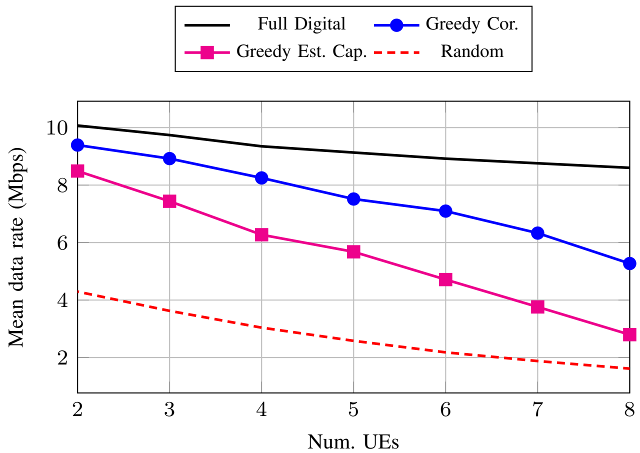

Fig. 3. Mean data rate in a frequency resource versus number of UEs considering 2 RF chains per BS.

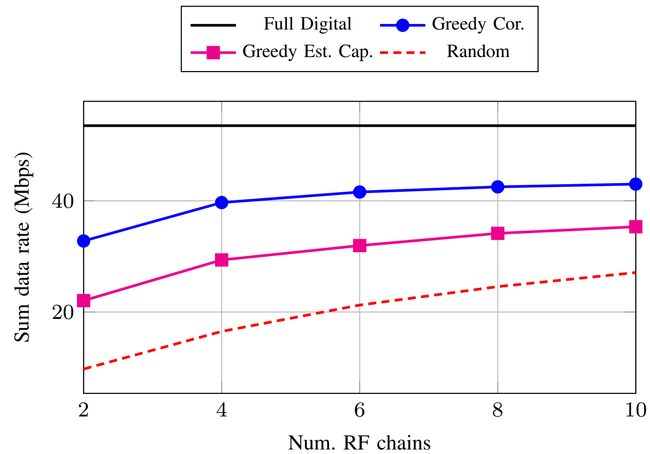

Fig. 4. Sum data rate in a frequency resource versus number of RF chains considering 6 UEs in the system.

good solution when the number of users tends to 8, in the sense that each UE will be allocated to each possible RF chain. This limitation is not present in the fully digital case, and this can be noted by its increased capacity with the number of UEs. The greedy correlation solution presents a better performance than the greedy estimated capacity. This behavior demonstrates that the spatial compatibility has an impact on the system capacity and should not be neglected.

Figure 3 shows the mean user data rate considering 2 RF chains per BS when the number of UEs increases. In this result, it can be noted that the mean data rate is reduced when the number of UEs increases for all simulated algorithms. The reason for that is the decreased orthogonality of the channels when the number of UEs in the system increases. The mean capacity achieves better performance for the Greedy-Correlation solution, when compared with all other hybrid beamforming algorithms.

In Figure 4, the performance considering 6 UEs and increasing the number of RF chains per BS is shown. It can be seen that the Greedy-Correlation still outperforms the other HBF solutions for all numbers of RF chains.

In all simulations the Greedy-Correlation outperforms the Greedy-Estimated Capacity, demonstrating that the correlation metric has a significant effect on the simulated scenarios. Note that the Greedy-Correlation considers the orthogonality of the spaces and the channel quality to perform the analog beam allocation, which improves the ability of the digital beamforming to isolate the interfering signals.

V. Conclusions

In this work, we have presented an analog beam allocation problem in Hybrid Beamforming C-RAN networks. The presented problem considers codebook-based analog beamforming applied at the RF chains of each BS and the digital beamforming performed at the C-RAN BBU.

The presented problem is non-convex and cannot be directly solved. Therefore, we have presented a solution decoupling the analog beam allocation, digital beamforming design and power allocation. The focus of this work is on the analog beam allocation step, for which we evaluate a new metric that considers both channel correlation and attenuation. The metric measures the spatial orthogonality created by the analog beamforming, which can improve the interfering signal isolation performed by the digital precoder. In order to analyze its performance, a greedy algorithm from the literature has been adapted to the considered scenario, taking into account different beam allocation criteria.

Our results have demonstrated that the correlation criterion has a considerable influence on the simulated scenario, leading to gains with regard to the other analyzed approach (Greedy-Estimated Capacity).

Acknowledgment

This work was financed in part by the Coordenação de Aperfeiçoamento de Pessoal de Nível Superior - Brasil (CAPES) - Finance Code 001 and by DAAD with funds from the Federal Ministry of Education and Research (BMBF). This work has been performed in the context of the DFG Collaborative Research Center (CRC) 1053 MAKI - subprojects C1 and C7. Yuri C. B. Silva also acknowledges the support of CNPq.

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