Robust LMMSE Beamformer Design by Naive UL/DL Duality and Validation for Non-Cooperative Massive MIMO

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***Abstract*—Massive multiple input multiple output (MaMIMO) is key to enabling the 1000 fold capacity improvement promised by 5G, thanks to the linear increase in system capacity with the number of base station (BS) antennas. In this work, we consider a MaMIMO interfering broadcast channel (IBC) with no coordination among the BSs under Time Division Duplexing (TDD). Each BS only has partial channel state information (CSI) of its own UE. Under this setting, we propose a duality based approach to combine the received covariance information at the BS and the estimated uplink (UL) channel from the UE to derive the downlink (DL) beamformers. We also take into account the fact that the channel reciprocity is impacted by the effect of the transmit and receive chain and correct it using the estimated reciprocity calibration factors. Essentially, using reciprocity, we transform the partial CSI at the receiver in the UL to a partial CSI at the transmitter in the downlink (DL). The technique is evaluated on the Eurecom OpenAirInterface Massive MIMO hardware testbed [1].**

***Index Terms*—IBC, beamforming, LMMSE, Massive MIMO, TDD channel reciprocity, partial CSIT, duality.**

I. Introduction

Beamformer design that maximizes weighted sum rate (WSR) for a multiple input multiple output (MIMO) Interfering Broadcast Channel (IBC) is a well-researched topic, particularly for full channel state information at the transmitter (CSIT) [2], [3], [4]. However, global full CSIT is not practical and what is more realistic is to consider partial CSIT or imperfect knowledge. Several works have considered a Gaussian CSIT model for partial CSIT and designed beamformers that maximize the expected value of the WSR (EWSR). Typical approaches to EWSR maximisation are based on alternate metrics that approximate this EWSR [4], [5].

At the same time, Massive Massive MIMO (MaMIMO) has become a key component in driving the data rate increase targeted by 5G, thanks to the linear increase in system capacity with number of base station (BS) antennas [6]. Massive MIMO is typically used together with Time Division Duplexing (TDD), where reciprocity of the propagation channel is exploited to derive the downlink (DL) channel from the uplink channel (UL) that is relatively easier to obtain. This becomes clear once we see that in the UL with one single pilot transmission from a user equipment (UE) the channel for all the BS antennas may be derived, whereas, in the DL, separate pilots are needed for every BS antenna which becomes prohibitive in the MaMIMO scenario. However, as the Radio Frequency (RF) components are not reciprocal, the DL estimation requires additional compensation for the non-reciprocity of the RF front ends.

In this work, we consider a Massive MIMO IBC scenario under TDD. We propose a well-founded, but simple, DL beamformer design where each BS only has estimated channel state information (hence, partial CSIT) of its own intracell users. In reality, what the BS really has is partial CSIR (channel state information at the receiver) and we use reciprocity to transform this into partial CSIT. We then validate this approach on the Eurecom Massive MIMO hardware testbed [1].

The rest of the paper is organized as follows. We first present the system model and theory for the MIMO IBC beamformer design and channel reciprocity calibration in section II. Next, in III, we present the demo configuration that we used in our test environment and the results obtained in hardware. In this work, we use the terms cell and BS interchangeably.

II. System Model

Consider an IBC with C cells and a total of K UE. We shall consider a system-wide numbering of the UEs. UE k has N_k number of antennas and is served by BS b_k which has M_{b_k} number of antennas. Only one stream is transmitted per UE. The received signal at user k in cell b_k is,

$$\mathbf{y}_k = \underbrace{\mathbf{H}_{k,b_k}\,\mathbf{g}_k\,x_k}_{\text{signal}} + \underbrace{\sum_{\substack{i\neq k\\ b_i=b_k}} \mathbf{H}_{k,b_k}\,\mathbf{g}_i\,x_i}_{\text{intracell interf.}} + \underbrace{\sum_{j\neq b_k}\sum_{i:b_i=j} \mathbf{H}_{k,j}\,\mathbf{g}_i\,x_i}_{\text{intercell interf.}} + \mathbf{v}_k \quad (1)$$

where x_k is the intended (white, identity covariance) signal, $\mathbf{H}_{k,b_k}$ is the $N_k \times M_{b_k}$ channel from BS b_k to user k. BS b_k serves $K_{b_k} = \sum_{i:b_i=b_k} 1$ UEs. The noise $\mathbf{v}_k \sim \mathcal{CN}(0,\mathbf{I})$. The $M_{b_k} \times 1$ spatial transmit (Tx) filter or beamformer (BF) is $\mathbf{g}_k$. The scenario of interest is that of partial CSIT available with every BS for the intra-cell users. The Gaussian CSIT model for the partial CSIT is

$$\mathbf{H}_{k,b_k} = \overline{\mathbf{H}}_{k,b_k} + \widetilde{\mathbf{H}}_{k,b_k}\,\mathbf{C}^{1/2}_{t,k,b_k} \quad (2)$$

where $\overline{\mathbf{H}}_{k,b_k} = \mathbf{E}\mathbf{H}_{k,b_k}$, and $\mathbf{C}^{1/2}_{t,k,b_k}$ is the Hermitian square-root of the Tx side covariance matrix. The elements of $\widetilde{\mathbf{H}}_{k,b_k}$ are i.i.d. $\sim \mathcal{CN}(0,1)$.

$$\mathbf{E}_{\mathbf{H}_{k,b_k}|\overline{\mathbf{H}}_{k,b_k}}(\mathbf{H}_{k,b_k}-\overline{\mathbf{H}}_{k,b_k})(\mathbf{H}_{k,b_k}-\overline{\mathbf{H}}_{k,b_k})^H = \text{tr}\{\mathbf{C}_{t,k,b_k}\}\mathbf{I}_{N_k}.$$
$$\mathbf{E}_{\mathbf{H}_{k,b_k}|\overline{\mathbf{H}}_{k,b_k}}(\mathbf{H}_{k,b_k}-\overline{\mathbf{H}}_{k,b_k})^H(\mathbf{H}_{k,b_k}-\overline{\mathbf{H}}_{k,b_k}) = \mathbf{C}_{t,k,b_k}. \tag{3}$$

Note that the expectation is done over $\mathbf{H}_{k,b_k}$, for a known $\overline{\mathbf{H}}_{k,b_k}$. This is true for all the expectation operations done in this paper. However, as the parameter over which the expectation is done is clear from the context, henceforth, we just mention the expectation operator $\mathbf{E}$ to reduce notational overhead.

A. Expected WSR (EWSR)

Once the CSIT is imperfect, various optimization criteria could be considered, such as outage capacity. Here we shall consider the EWSR for a known channel mean $\overline{\mathbf{H}}$.

$$\begin{aligned}\text{EWSR}(\mathbf{g}) &= \mathbf{E}\sum_k u_k \ln|1+\mathbf{g}_k^H\mathbf{H}_{k,b_k}^H\mathbf{R}_{\overline{k}}^{-1}\mathbf{H}_{k,b_k}\mathbf{g}_k| \\ &= \mathbf{E}\sum_{k=1}^{K} u_k \left(\ln|\mathbf{R}_k| - \ln|\mathbf{R}_{\overline{k}}|\right).\end{aligned} \tag{4}$$

Here, $\mathbf{g}$ represents the collection of BFs $\mathbf{g}_k$, u_k are rate weights.

$$\begin{aligned}&\mathbf{R}_k = \mathbf{H}_{k,b_k}\mathbf{Q}_k\,\mathbf{H}_{k,b_k}^H + \mathbf{R}_{\overline{k}},\ \mathbf{Q}_i = \mathbf{g}_i\mathbf{g}_i^H, \\ &\mathbf{R}_{\overline{k}} = \sum_{i\neq k}\mathbf{H}_{k,b_i}\mathbf{Q}_i\,\mathbf{H}_{k,b_i}^H + \mathbf{I}_{N_k}.\end{aligned} \tag{5}$$

The EWSR cost function needs to be augmented with the power constraints $\sum_{k:b_k=j}\text{tr}\{\mathbf{Q}_k\} \leq P_j$.

B. EWSR Lower Bound: EWSMSE

The criterion $\text{EWSR}(\mathbf{g})$ is difficult to compute and to maximize directly due to the presence of the expectation operation. An alternative approach is to first introduce a receive (Rx) beamformer $\mathbf{f}_k$. The Rx filter output,

$$\hat{x}_k = \mathbf{f}_k^H\mathbf{y}_k = \mathbf{f}_k^H\mathbf{H}_{k,b_k}\,\mathbf{g}_k\,x_k + \sum_{i\neq k}\mathbf{f}_k^H\mathbf{H}_{k,b_i}\,\mathbf{g}_i\,x_i + \mathbf{f}_k^H\mathbf{v}_k \tag{6}$$

With this, the mean square error (MSE) may be obtained as,

$$\begin{aligned}e_k(\mathbf{f}_k,\mathbf{g}_k,\mathbf{H}) = 1 &- \mathbf{f}_k^H\mathbf{H}_{k,b_k}\,\mathbf{g}_k - \mathbf{g}_k^H\mathbf{H}_{k,b_k}\,\mathbf{f}_k \\ &+ \sum_i \mathbf{f}_k^H\mathbf{H}_{k,b_i}\,\mathbf{g}_i\mathbf{g}_i^H\mathbf{H}_{k,b_i}^H\,\mathbf{f}_k + ||\mathbf{f}_k||^2.\end{aligned} \tag{7}$$

Here, $\mathbf{H}$ refers to the collection of all $\mathbf{H}_{k,b_i}$. It turns out that it is much more attractive to consider $\mathbf{E}e_k(\mathbf{f}_k,\mathbf{g}_k,\mathbf{H})$ as in [7] since $e_k(\mathbf{f}_k,\mathbf{g}_k,\mathbf{H})$ is quadratic in $\mathbf{H}$. Hence consider optimizing the expected weighted sum MSE, $\mathbf{E}WSMSE(\mathbf{g},\mathbf{f},w,\mathbf{H})$.

$$\begin{aligned}&\min\nolimits_{\mathbf{f},w}\mathbf{E}_{\mathbf{H}|\overline{\mathbf{H}}}WSMSE(\mathbf{g},\mathbf{f},w,\mathbf{H}) \\ &\geq \mathbf{E}_{\mathbf{H}|\overline{\mathbf{H}}}\min\nolimits_{\mathbf{f},w}WSMSE(\mathbf{g},\mathbf{f},w,\mathbf{H}) = -EWSR(\mathbf{g})\end{aligned} \tag{8}$$

or hence

$$EWSR(\mathbf{g}) \geq -\min_{\mathbf{f},w}\mathbf{E}_{\mathbf{H}|\overline{\mathbf{H}}}WSMSE(\mathbf{g},\mathbf{f},w,\mathbf{H})\,. \tag{9}$$

Thus, this approach results in maximization of a lower bound of EWSR.

$$\begin{aligned}\mathbf{E}e_k = \widehat{e}_k = 1 - 2\Re\{\mathbf{f}_k^H\overline{\mathbf{H}}_{k,b_k}\mathbf{g}_k\} + \sum_{i=1}^{K}\mathbf{f}_k^H\overline{\mathbf{H}}_{k,b_i}\mathbf{g}_i\mathbf{g}_i^H\overline{\mathbf{H}}_{k,b_i}^H\mathbf{f}_k \\ + ||\mathbf{f}_k||^2\sum_{i=1}^{K}\mathbf{g}_i^H\mathbf{C}_{t,k,b_i}\mathbf{g}_i + ||\mathbf{f}_k||^2.\end{aligned}$$

where $\mathbf{C}_{t,k,b_i}$ are Tx side (LMMSE error) covariance matrices of $\mathbf{H}_{k,b_i}$. Note that the signal term disappears if $\overline{\mathbf{H}}_{k,b_k} = 0$. Hence the EWSMSE lower bound is very loose unless the Rice factor is high, and is useless in the absence of channel estimates. The overall algorithm for determining the beamformers is to perform alternating optimization amongst the following,

$$\begin{aligned}&\min_{w_k} EWSMSE \ \Rightarrow\ w_k = 1/\widehat{e}_k \\ &\min_{\mathbf{f}_k} EWSMSE \Rightarrow \mathbf{f}_k = \widehat{\mathbf{R}}_k^{-1}\overline{\mathbf{H}}_{k,b_k}\mathbf{g}_k \\ &\min_{\mathbf{g}_k} EWSMSE \Rightarrow \mathbf{g}_k = (\widehat{\mathbf{T}}_k + \lambda_{b_k}I_M)^{-1}\overline{\mathbf{H}}_{k,b_k}^H\mathbf{f}_k u_k w_k\end{aligned} \tag{10}$$

where

$$\begin{aligned}\widehat{\mathbf{R}}_k &= \textstyle\sum_i \overline{\mathbf{H}}_{k,b_i}\mathbf{g}_i\mathbf{g}_i^H\overline{\mathbf{H}}_{k,b_i}^H + \left(1+\sum_i \mathbf{g}_i^H\mathbf{C}_{t,k,b_i}\mathbf{g}_i\right)\mathbf{I}_{N_k} \\ \widehat{\mathbf{T}}_k &= \textstyle\sum_{i=1}^{K} u_i w_i(\overline{\mathbf{H}}_{i,b_k}^H\mathbf{f}_i\mathbf{f}_i^H\overline{\mathbf{H}}_{i,b_k} + ||\mathbf{f}_i||^2\mathbf{C}_{t,k,b_i})\,.\end{aligned} \tag{11}$$

Here, λ_{b_k} corresponds to the Lagrangian multiplier for the transmit power constraint at BS b_k. We remark here that a key interpretation of the EWSMSE equations is that the optimal transmit beamformer $\mathbf{g}_k$ has the form of a linear MMSE receiver for the dual UL.

C. MaMIMO limit and ESEI-WSR

If the number of Tx antennas M becomes very large, we get a convergence for any quadratic term of the form,

$$\mathbf{H}\mathbf{Q}\mathbf{H}^H \xrightarrow{M\rightarrow\infty} \mathbf{E}\,\mathbf{H}\mathbf{Q}\mathbf{H}^H = \overline{\mathbf{H}}\mathbf{Q}\overline{\mathbf{H}}^H + \text{tr}\{\mathbf{Q}\mathbf{C}_t\}. \tag{12}$$

Hence, we get the following MaMIMO limit matrices,

$$\begin{aligned}\breve{\mathbf{R}}_k &= \breve{\mathbf{R}}_{\overline{k}} + \overline{\mathbf{H}}_{k,b_k}\mathbf{Q}_k\overline{\mathbf{H}}_{k,b_k}^H + \text{tr}\{\mathbf{Q}_k\mathbf{C}_{t,k,b_k}\}\mathbf{I}_{N_k} \\ \breve{\mathbf{R}}_{\overline{k}} &= \mathbf{I}_{N_k} + \sum_{i\neq k}^{K}\left(\overline{\mathbf{H}}_{k,b_i}\mathbf{Q}_i\overline{\mathbf{H}}_{k,b_i}^H + \text{tr}\{\mathbf{Q}_i\mathbf{C}_{t,k,b_i}\}\mathbf{I}_{N_k}\right)\end{aligned}$$

Now, typical approaches to solve the WSR (eg. the DC approach in [3]) can be run to obtain the max EWSR BF. We shall refer to this approach as the ESEI-WSR approach as (channel dependent) signal and interference covariance matrices are replaced by their expected values. It was shown in [5] that the actual gap between these two criteria is zero at very low as well as very high SNR for a multi-user MISO scenario.

D. TDD channel reciprocity

Consider a system as in Fig. 1, where A represents a BS with M antennas and B represents a single antenna UE. We specifically consider a single antenna UE as it corresponds to our demo setting that will be described later. The channel, as observed in the digital domain, $\mathbf{h}_{A\rightarrow B}$ and $\mathbf{h}_{B\rightarrow A}$ can be represented by,

$$\mathbf{h}_{A\rightarrow B} = r_B\mathbf{c}_{A\rightarrow B}\mathbf{T}_A, \qquad \mathbf{h}_{B\rightarrow A} = \mathbf{R}_A\mathbf{c}_{B\rightarrow A}t_B, \tag{13}$$

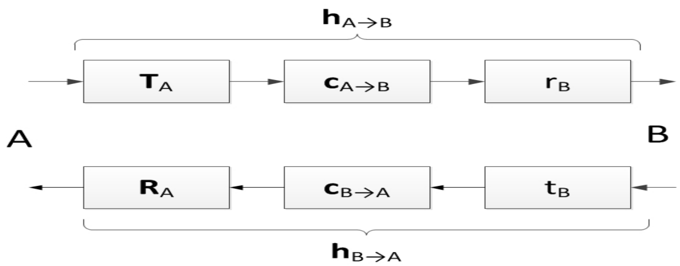

Fig. 1. Reciprocity Model

where matrices $\mathbf{T}_A$, $\mathbf{R}_A$ model the response of the transmit and receive RF front-ends at the BS, while $\mathbf{c}_{A\to B}$ and $\mathbf{c}_{B\to A}$ model the propagation channels, respectively from A to B and from B to A. The dimension of $\mathbf{T}_A$ and $\mathbf{R}_A$ are $M \times M$. Complex scalars $\mathbf{r}_B$, $\mathbf{t}_B$ model the response of the transmit and receive RF front-ends at the UE. The diagonal elements in the matrices $\mathbf{T}_A$, $\mathbf{R}_A$ represent the linear effects attributable to the impairments in the transmitter and receiver parts of the RF front-ends respectively, whereas the off-diagonal elements correspond to non-reciprocity in RF crosstalk and antenna mutual coupling. It is worth noting that although transmitting and receiving antenna mutual coupling are not generally reciprocal [8], theoretical modeling [9] and experimental results [10]–[12] both show that, in practice, RF crosstalk and antenna mutual coupling are sufficiently reciprocal to be ignored for the purpose of reciprocity calibration, which implies that $\mathbf{T}_A$, $\mathbf{R}_A$ can safely be assumed to be diagonal.

As the system is operating in TDD mode, the channel responses enjoy reciprocity within the channel coherence time, i.e., $\mathbf{c}_{A\to B} = \mathbf{c}^T_{B\to A}$. Therefore, we obtain the following relationship between the channels measured in both directions:

$$\mathbf{h}_{A\to B} = \underbrace{\mathbf{r}_B\mathbf{t}_B^{-1}}_{\mathbf{f}_B}\mathbf{h}^T_{B\to A}\underbrace{\mathbf{R}_A^{-T}\mathbf{T}_A}_{\mathbf{F}} = \mathbf{f}_B^{-T}\mathbf{h}^T_{B\to A}\mathbf{F}. \tag{14}$$

As the UE side calibration parameter is just a scaling factor, we can equivalently derive the DL channel as,

$$\mathbf{h}_{A\to B} = \mathbf{h}^T_{B\to A}\mathbf{F}. \tag{15}$$

E. Dual DL beamformer

At this point, we design the partial CSIT beamformer based on the EWSMSE approach for a specific case of $N_k = 1$. i.e., when there is only one antenna at the UE. Note that this assumption is not too restrictive as single antenna UE is the typical configuration in Massive MIMO. We shall design the partial CSIT beamformer based on a naive UL/DL duality. The relations between Rx $\mathbf{f}_k$ and Tx $\mathbf{g}_k$ in equation (10) represent a proper UL/DL duality as one can observe that the optimal DL BF $\mathbf{g}_k$ corresponds to a LMMSE Rx in a dual UL in which the UL channels would be $\mathbf{H}^H_{i,b_k}$, the UL Tx filters would be $\mathbf{f}_i$, the UL stream powers would be $u_i w_i$ and the white noise variance at the BS would be λ_{b_k}. These dual UL quantities are obviously different from corresponding actual UL transmission quantities. However, in order to largely simplify BF design and reduce signaling overhead, we propose a naive duality BF design in which we use the actual UL LMMSE Rx as DL BF. Note that one difference between actual and dual UL is a complex conjugation on the channel responses. Also, in the case of $N_k = 1$, we can ignore the UE side BF f_k. Note however that the resulting naive UL/DL duality BF design will converge to a matched filter at low SNR, and to a zero forcing (ZF) at high SNR. Hence the naive duality gives optimal results at both low or high SNR. Finally, for the partial CSI aspect, we shall replace statistical (channel) averaging by temporal averaging.

The received signal $\check{\mathbf{y}}_k$ at BS b_k may be written as,

$$\check{\mathbf{y}}_k = \check{\mathbf{h}}_{k,b_k} s_k + \check{\mathbf{v}}_{b_k}. \tag{16}$$

Here, $\check{\mathbf{v}}_{b_k}$ includes the AWGN noise as well as the received signal from all other users, both intracell and inter-cell. $\check{\mathbf{h}}_{k,b_k}$ denotes the uplink channel from the user k to BS b_k. Let $\mathbf{R}_{\check{y}\check{y}}$ be the uplink correlation matrix. Then the UL MMSE estimator is given by,

$$\mathbf{g}^{UL}_{MMSE} = \check{\mathbf{h}}^H_{k,b_k}\mathbf{R}^{-1}_{\check{y}\check{y}} \tag{17}$$

Note that we have used the covariance matrix $\mathbf{R}_{\check{y}\check{y}}$ as the partial CSIT is also local. I.e, each BS only has partial CSIT corresponding to its own UEs and not UEs corresponding to other BSs. Using reciprocity in TDD and accounting for the calibration factors, the DL MMSE estimator is given by

$$\begin{aligned}\mathbf{g}^{DL}_{MMSE} = \mathbf{R}^{-1}_{yy,dl}\mathbf{h}^H_{k,b_k} &= (\mathbf{F}^H\mathbf{R}_{\check{y}\check{y}}\mathbf{F})^{-1}\mathbf{F}^H\check{\mathbf{h}}^*_{k,b_k} \\ &= \mathbf{F}^{-1}\mathbf{R}^{-1}_{\check{y}\check{y}}\check{\mathbf{h}}^*_{k,b_k}.\end{aligned} \tag{18}$$

Here, $()^*$ denotes the conjugation operation. The covariance matrix is derived as a sample covariance as follows.

$$\mathbf{R}_{\check{y}\check{y}} = \frac{1}{L}\sum_{i=1}^{L}\check{\mathbf{y}}_k\check{\mathbf{y}}_k^H \tag{19}$$

A known issue with this approach is signal cancellation that occurs due to the mismatch between the estimated channel of the desired UE and the implicit component of the desired channel present in the sample covariance matrix [13]. A known solution in this context is the subtraction of the desired signal before computing the covariance matrix. This requires an iterative receiver for joint detection, channel estimation so that the BS can subtract out the contribution from its own UE before computing the covariance matrix.

III. Experimental Validation

Figure 2 shows an image of the MaMIMO prototype that is a part of the Eurecom OpenAirInterface platform. The Eurecom Massive MIMO array is constructed with several microstrip antenna cards, 12 of which are used in the current validation. Each such microstrip card, in turn, has 4 antennas. The 48 antennas are driven by 12 radio cards, where each radio card has 4 transceiver units. The transmission happens at carrier frequency $f_c = 2.66$GHz and the sampling frequency $f_s = 7.68$MHz corresponding to a 5MHz LTE transmission that uses 300 occupied subcarriers. The beamformer design is applied individually on every frequency subcarrier.

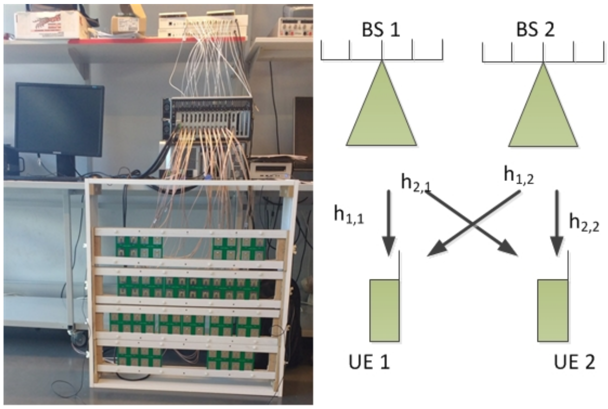

Fig. 2. Eurecom MaMIMO prototype and Demo set up

Figure 2 also illustrates the demo scenario. The two base BS units consist of 23 antennas each and the two UEs have one antenna each. Thus, 48 antennas of the MaMIMO antenna array are used to mimic the BS as well as the two single antenna UEs.

The demo exploits channel reciprocity to derive the DL beamformer weights based on the UL channel/covariance estimates. Hence, when the prototype is initialized, we perform a reciprocity calibration and store the reciprocity calibration parameters $\mathbf{F}$ in a file. Subsequently, this file is read to derive the DL beamformer using instantaneously estimated UL channel information as was given in (18). The instantaneous UL channel estimation is based on UL pilots. In our demo, we assume all the useful subcarriers as pilots in the UL. The quality of the channel estimates are further improved by exploiting the limited time domain spread of the channel taps. Our DL LMMSE design assumes no knowledge of the cross-links between the BS of one cell and UE of another. However, to serve as a reference, we also consider a ZF receiver that has full knowledge of all cross-links. In this case, let the UL channel matrix be,

$$\mathcal{H} = \begin{bmatrix}\breve{\mathbf{h}}_{1,1} & \breve{\mathbf{h}}_{2,1}\end{bmatrix} \tag{20}$$

Then,

$$\mathbf{g}_{ZF} = \mathbf{F}^{-1}(\mathcal{H}^{*}\mathcal{H}^{T})^{-1}\breve{\mathbf{h}}^{*}_{1,1}. \tag{21}$$

The other popular receiver in a MaMIMO scenario is the maximal ratio transmitter (MRT), which in this case would be,

$$\mathbf{g}_{MRT} = \mathbf{F}^{*}\breve{\mathbf{h}}^{*}_{1,1}. \tag{22}$$

The estimation of the covariance matrix needs significant averaging, particularly as the number of BS antennas increases. In our prototype, we exploit the low delay spread of the environment and compute the average covariance matrix across all the subcarriers.

Figure 3 shows the need for calibration by taking the example of MRT in a single BS single UE setting. The performance is measured on the basis of the ratio between the received signal power and the noise power (SNR) observed at the UE. The curve ”ideal” here refers to the case where the DL channel estimate is available and estimated directly. The curve ”calib” refers to the implementation of (22) and the curve ”no_calib” directly uses the estimated UL channel for DL beamforming without applying any reciprocity calibration. The SNR is shown for all the 300 occupied subcarriers of the 5MHz LTE orthogonal frequency division multiplexing (OFDM) symbol.

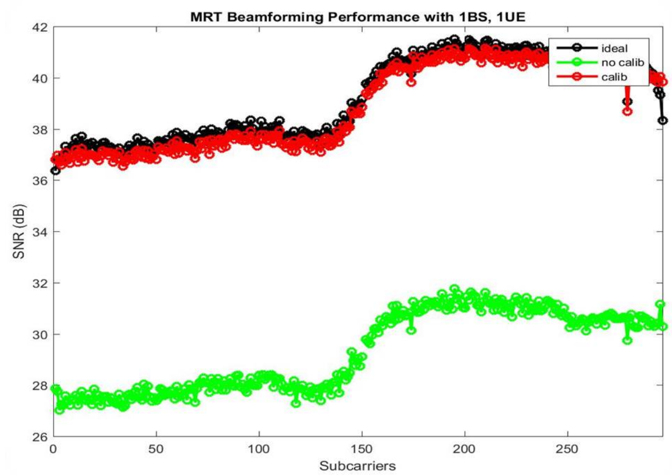

Fig. 3. Performance of MRT with and without calibration for a 23 antenna BS with a single UE.

Figure 4 shows the relative gains of MRT and ZF beamformers compared to no beamforming (omnidirectional antenna) by measuring the signal to interference plus noise ratio (SINR) at UE 1 as a result of using the different beamformer techniques. It is remarkable that the performance of the ZF beamformer is far superior to that of the MRT which is the most widespread beamforming technique used for MaMIMO. In Figure 5, covariance matrix is estimated for the interfering links in the UL and the DL MMSE BF is derived based on the UL covariance estimates and the reciprocity calibration parameters. The curve ZF serves as a reference where the UL channels of the interfering links are known, so that the DL ZF beamforming can be done with the help of reciprocity calibration as shown in (21). The curve MMSE_Ryy is the scenario where the BS computes the covariance based on the total received signal from both its own UE and the interfering UE. We are limited here by the accuracy of the channel estimation and the averaging required for the covariance estimation. For the massive MIMO BS configuration, the averaging requirement for the covariance matrix estimation is very stringent as the dimension of the covariance matrix grows proportionally to the square of the number of BS antennas. Due to inaccuracy in channel estimation, signal cancellation occurs between the channel estimate (in MF) and the channel contribution in $\mathbf{R}_{\breve{y}\breve{y}}$. The curve MMSE_Ryy_IntfOnly corresponds to the scenario where the covariance of the transmission from the interfering UE is used for DL MMSE BF along with the reciprocity parameters. This approach avoids the signal cancellation issue. Hence, we

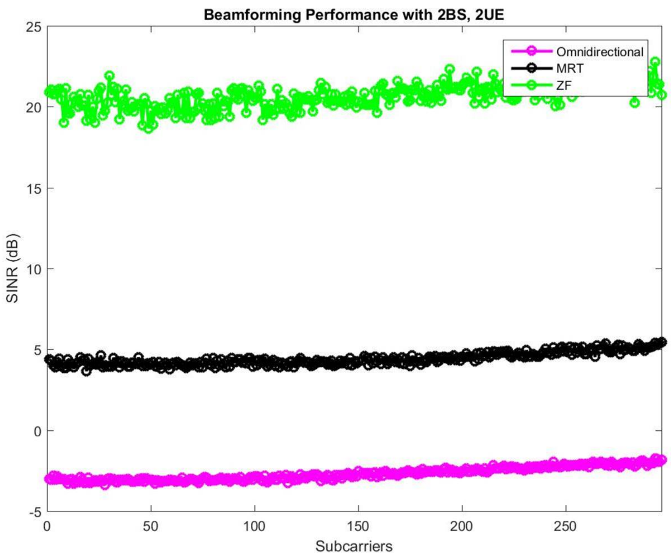

Fig. 4. Performance of MRT and ZF beamformers compared to no beamforming

observe that the performance of MMSE_Ryy is much poorer compared to that of the curve MMSE_Ryy_IntfOnly for the massive MIMO BS. In fact, the performance of the curve MMSE_Ryy_IntfOnly is quite close to that of the ZF which has knowledge of the interfering links as well.

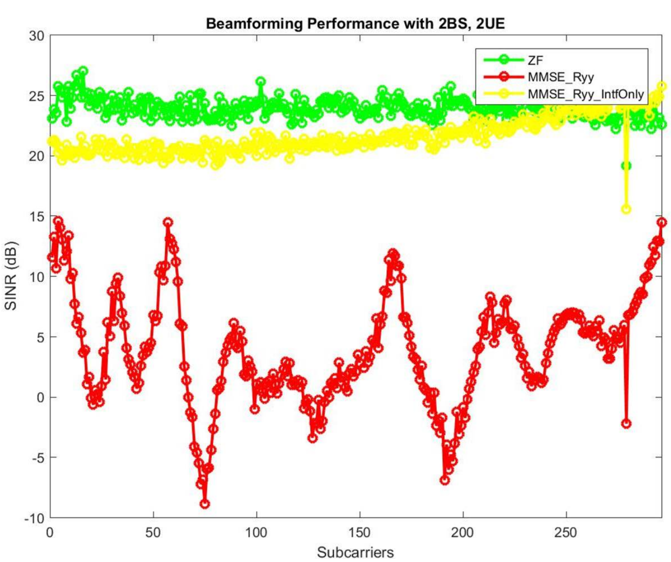

Fig. 5. Comparison of the performance of partial CSIT LMMSE beamformer with that of ZF which requires full information of cross-links.

IV. Conclusion

In this work, we have designed a DL beamformer for a Massive MIMO IBC scenario with partial channel state information. In particular, there is no cooperation assumed across the different BSs. Using TDD channel reciprocity we effectively converted a partial CSIR problem to that of partial CSIT. We then considered the estimation of the UL covariance matrix while avoiding the signal cancellation issue as part of the overall beamformer design. The resulting beamformer was then validated on Eurecom's Massive MIMO OpenAirInterface platform and shows a performance close to that of a ZF beamformer that needs information of the cross-links.

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